

## SECTION 10

### HIGH SPEED SIGNAL AMPLIFICATION

- HIGH SPEED AMPLIFIER ARCHITECTURES
- CURRENT FEEDBACK AMPLIFIERS
- COMPARISON BETWEEN VOLTAGE FEEDBACK AND CURRENT FEEDBACK OP AMPS
- HIGH SPEED BUFFER AMPLIFIERS
- HIGH SPEED OP AMP NOISE MODELS
- OP AMP DC MODEL
- LEVEL SHIFTING HIGH SPEED SIGNALS USING OP AMPS



## SECTION 10

# HIGH SPEED SIGNAL AMPLIFICATION

*Walt Kester*

The amplification of high speed signals while preserving fidelity presents a significant design challenge. Although there is an overwhelming number of op amps from which to select, only a few satisfy the stringent requirements of high speed signal processing systems. Figure 10.1 lists the selection criteria in the approximate order of decreasing importance. Dynamic range (measured in terms of distortion and noise) and bandwidth are major driving forces in high speed signal processing systems. These parameters are especially important in applications (such as spectral

analysis) where the preservation of spectral purity is paramount. Settling time is important in pulse analysis, while video applications require high bandwidths (flat over a wide range of frequencies) and low levels of differential gain and phase.

In the majority of high speed systems, ac performance usually takes priority over dc precision. Nevertheless, modern high speed amplifiers achieve levels of dc accuracy which are usually more than adequate for most applications.

**10**

### HIGH SPEED AMPLIFIER SELECTION CRITERIA

- Distortion Under Load
- Bandwidth and Bandwidth Flatness
- Broadband Noise
- DC and Capacitive Output Drive Capability
- Settling Time
- Video-Specifications: Differential Gain and Phase
- DC Offset, Drift, and Input Bias Current

Figure 10.1

## HIGH SPEED AMPLIFIER ARCHITECTURES

Although it is possible to select high speed amplifiers without much understanding of their internal architectures or circuits, the following discussion will allow the designer to make much more intelligent tradeoffs and perhaps complete the system design in a more timely manner with better performance. In recent years, the differences in amplifier architectures are becoming

somewhat overshadowed by the similarities in final performance. For example, *current feedback* op amps (sometimes called *transimpedance* amplifiers) are not necessarily the best choice in all high speed applications. There are a number of applications where optimized fixed-gain *voltage feedback* buffers provide the best solution.

## POPULAR HIGH SPEED AMPLIFIER ARCHITECTURES

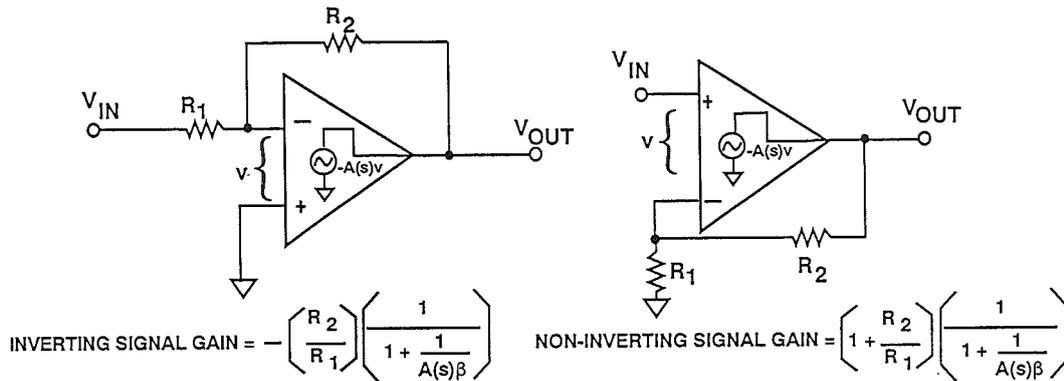
- Voltage Feedback
- Current Feedback
- Optimized Fixed-Gain Closed-Loop Buffers

Figure 10.2

The equivalent circuit for a typical voltage feedback op amp is shown in Figure 10.3. The input voltage is multiplied by the open loop gain  $A(s)$  to yield the output voltage. The feedback at-

tenuation factor is  $\beta$ . The equations which relate the input and output voltage for the inverting and non-inverting mode are also given in Figure 10.3.

VOLTAGE FEEDBACK OP AMP EQUIVALENT CIRCUITS



$A(s)$  = OPEN LOOP VOLTAGE GAIN

$$\beta = \text{FEEDBACK FACTOR} = \frac{R_1}{R_1 + R_2}$$

$$\frac{1}{\beta} = \text{NOISE GAIN} = 1 + \frac{R_2}{R_1}$$

$$\text{LOOP GAIN} = A(s)\beta$$

Figure 10.3

The term  $A(s)\beta$  is referred to as the *loop gain*. It is the value of the loop gain at a specified frequency which will determine the overall accuracy of the op amp at that frequency. The so-called *curative* effects of feedback at any frequency are determined by the available amount of

loop gain at that frequency. Loop gain affects gain accuracy and stability, linearity, distortion, input impedance, and output impedance. The corresponding Bode plot for a single pole rolloff with fixed compensation is shown in Figure 10.4.

LOG-LOG BODE PLOT FOR VOLTAGE FEEDBACK OP AMP

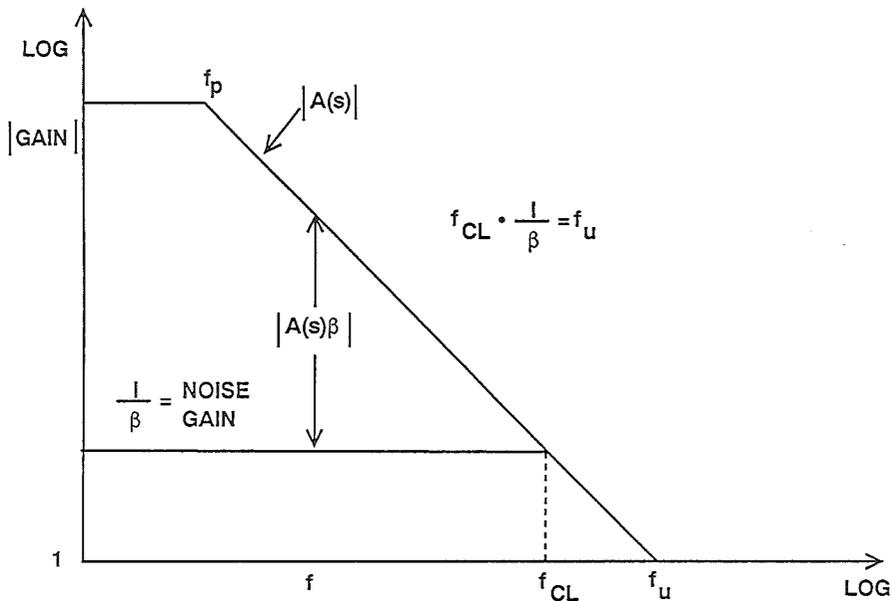


Figure 10.4

The *gain-bandwidth product* of an op amp is simply the product of the closed loop gain and the corresponding bandwidth at a specified frequency. For a voltage feedback amplifier which has a single-pole frequency response, this product is constant over a wide range of frequencies (see Figure 10.4). If the op amp is stable at unity gain, the frequency at which the open loop response crosses unity gain is called the *unity gain bandwidth frequency*. The gain-bandwidth specification may thus be used to calculate the closed-loop bandwidth for various values of closed-loop gain.

bandwidth is that the closed loop gain,  $A_{CL}$ , refers to the *noise gain*,  $1/\beta$ , and not the *signal gain* (see Figure 10.5). For instance, in the non-inverting mode, the dc *signal gain*  $(1+R_2/R_1)$  is equal to the dc noise gain. In the inverting mode, however, the noise gain remains  $1+R_2/R_1$ , but the signal gain is now  $-R_2/R_1$ . For example, if an op amp has a gain-bandwidth product of 10MHz, the closed loop bandwidth for a non-inverting unity gain configuration is 10MHz, while that of a unity-gain inverter is only 5MHz.

A key point often confused in selecting voltage feedback op amps based on

RELATIONSHIP BETWEEN NOISE GAIN, SIGNAL GAIN, AND BANDWIDTH FOR OP AMP WITH UNITY GAIN BW OF 10MHZ

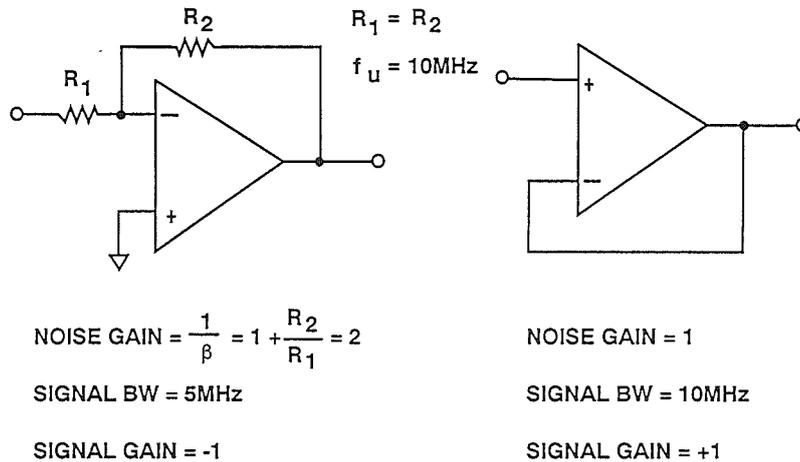


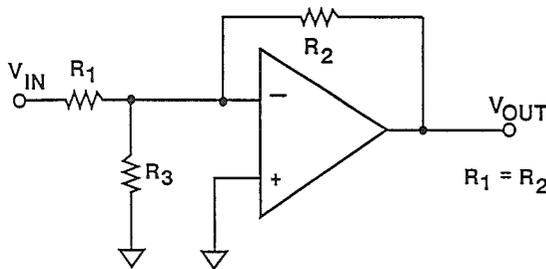
Figure 10.5

10

Another important point is that some op amps are optimized to operate at high gains, and are not stable under low- or unity-gain conditions. With these op amps, the gain-bandwidth product is meaningful only over the region of stable closed-loop gains. This type of amplifier can, however, be used at low inverting signal gains by the addition of a shunt resistor to ground as

shown in Figure 10.6. The extra resistor is chosen such that the noise gain is greater than the minimum value required for stability. The penalty is increased sensitivity to input offset voltage and input noise voltage as well as lower signal bandwidth. A capacitor may be used in series with the resistor to avoid increasing the dc noise gain.

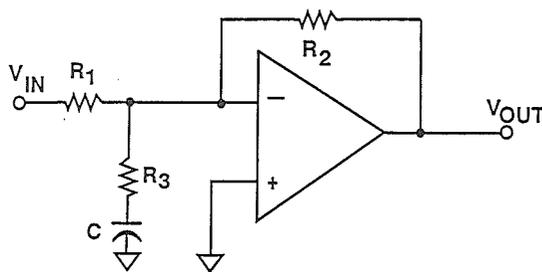
## OPERATING NON-UNITY GAIN STABLE OP AMPS AT UNITY GAIN IN THE INVERTING MODE



$$\text{SIGNAL GAIN} = - \frac{R_2}{R_1} = -1$$

$$\text{NOISE GAIN} = 1 + \frac{R_2}{R_1 \parallel R_3}$$

CHOOSE  $R_3$  FOR MINIMUM  
STABLE NOISE GAIN



LARGE C GIVES HIGH AC  
NOISE GAIN, BUT DC  
NOISE GAIN REMAINS LOW

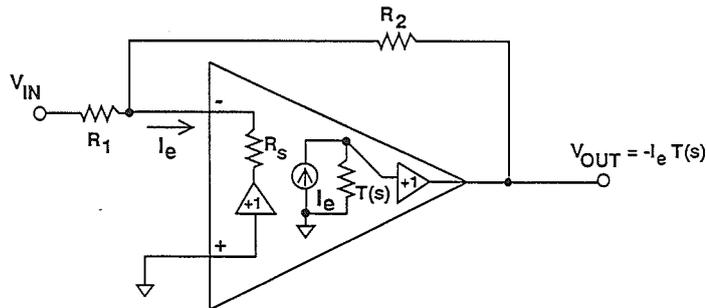
Figure 10.6

### CURRENT FEEDBACK AMPLIFIERS

The equivalent circuit for a current feedback amplifier is shown in Figure 10.7. The signal at the non-inverting input is applied to the inverting input through a unity-gain buffer with an output impedance  $R_s$  (usually between 10 and 100 $\Omega$ ). The current entering the inverting input is multiplied by the *transimpedance open loop gain*,  $T(s)$ , to yield the output voltage. The feedback attenuation factor of the current feedback amplifier is different from the

voltage feedback amplifier because of the low inverting input impedance,  $R_s$ . Solving the feedback equation yields the transfer function shown in Figure 10.7. But although the expression for the current feedback amplifier loop gain is different from a voltage feedback amplifier, it can be used in exactly the same manner in determining the accuracy of the amplifier closed loop gain at any specific frequency.

CURRENT FEEDBACK OP AMP EQUIVALENT CIRCUIT



$$\text{INVERTING SIGNAL GAIN} = - \left( \frac{R_2}{R_1} \right) \left( \frac{1}{1 + \frac{1}{\text{LG}}} \right) \quad \text{NON-INVERTING SIGNAL GAIN} = \left( 1 + \frac{R_2}{R_1} \right) \left( \frac{1}{1 + \frac{1}{\text{LG}}} \right)$$

$T(s)$  = OPEN LOOP TRANSIMPEDANCE GAIN

$$\beta_{cf} = \text{FEEDBACK FACTOR} = \frac{R_s \parallel R_1}{R_s \parallel R_1 + R_2}$$

$$A_{cf}(s) = \text{OPEN LOOP VOLTAGE GAIN} = \frac{T(s)}{R_s}$$

$$\text{LOOP GAIN} = \text{LG} = A_{cf}(s) \beta_{cf} = \frac{T(s) \{ R_s \parallel R_1 \}}{R_s \{ R_s \parallel R_1 + R_2 \}}$$

Figure 10.7

COMPARISON BETWEEN VOLTAGE FEEDBACK AND CURRENT FEEDBACK OP AMPS

The inverting mode transfer functions for the voltage feedback amplifier and the current feedback amplifier are compared in Figure 10.8. Notice that for the voltage feedback amplifier, the frequency-dependent term,  $1/A(s)$ ,

is multiplied by the noise gain,  $(1 + R_2/R_1)$ . This implies that the closed loop bandwidth is approximately inversely proportional to the noise gain, hence, the product of the noise gain and the closed loop bandwidth is constant.

## VOLTAGE FEEDBACK AND CURRENT FEEDBACK INVERTER CLOSED LOOP GAIN EQUATIONS

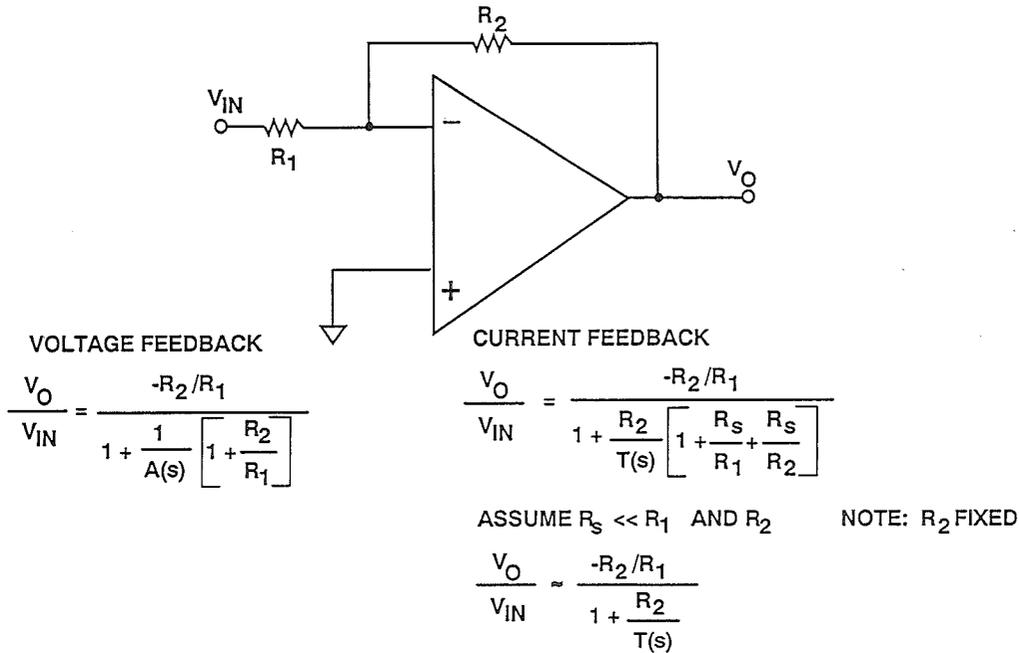


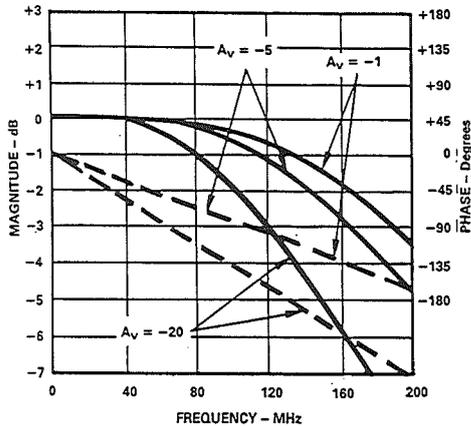
Figure 10.8

In the current feedback amplifier, however, if  $R_S \ll R_1$  and  $R_2$ , the closed loop bandwidth is independent of the gain,  $R_2/R_1$ , and depends only upon the feedback resistor  $R_2$ . Furthermore, most current feedback amplifiers are optimized for maximum bandwidth with a particular value of  $R_2$ . This implies that the closed loop bandwidth of a current feedback amplifier will remain fairly constant regardless of closed loop gain, provided the gain is changed by varying only  $R_1$ .

Current feedback (or transimpedance) op amps therefore have bandwidths which are relatively independent of closed loop gains (assuming the feedback resistor value remains constant). Therefore, it is inappropriate to refer to the gain-bandwidth product of this type

of amplifier. For instance, the signal bandwidth of the AD9617 with a 400Ω feedback resistor is approximately 190MHz for a closed loop signal gain of -1, and approximately 165MHz for a closed loop-signal gain of -5 (see Figure 10.9). In the first case, the so-called gain-bandwidth product would be 190MHz (1 × 190MHz), while in the second case it would be 825MHz (5 × 165MHz). In addition, current feedback amplifiers are usually optimized for a fixed value of feedback resistor. Increasing the feedback resistor lowers the bandwidth proportionally, while decreasing the value may lead to instability. The closed loop signal bandwidth for current feedback amplifiers should therefore be determined from curves on the data sheet.

## GAIN AND PHASE RESPONSE FOR AD9617 CURRENT FEEDBACK OP AMP



GAIN	BW	PRODUCT
-1	190MHz	190MHz
-5	165MHz	825MHz

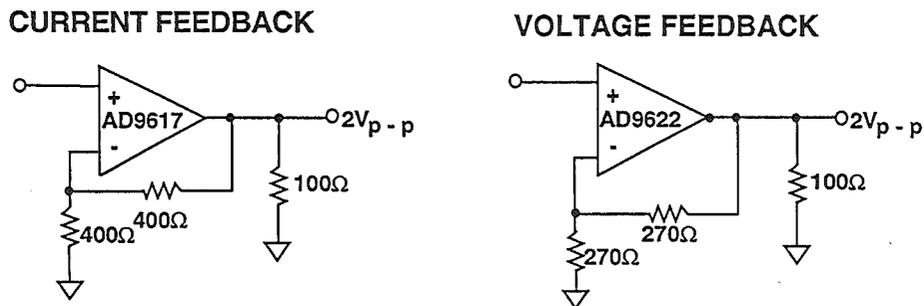
Figure 10.9

The requirement of a fixed, low-value (typically  $400\Omega$  to  $1000\Omega$ ) feedback resistor becomes a significant disadvantage when using a current feedback op amp in the inverting mode at large gains. For example, with the AD9617 optimum feedback resistor of  $400\Omega$ , the feedforward resistor must be  $40\Omega$  to achieve an inverting gain of 10. Driving the low value feedforward resistor may

become a significant problem. Therefore, the non-inverting configuration is generally preferable when using current feedback amplifiers at high gains.

In order to evaluate the performance differences between voltage and current feedback op amps, consider the simple gain-of-2 circuits shown in Figure 10.10.

## HIGH SPEED, GAIN-OF-TWO, 200MHz OP AMP BUFFER



	AD9617	AD9622
DISTORTION AT 1 MHz	-92dBc	-76dBc
DISTORTION AT 20 MHz	-67dBc	-66dBc
RMS OUTPUT NOISE (0.1Hz TO 200MHz)	172μV rms	98μV rms

Figure 10.10

One circuit uses the AD9622 voltage feedback op amp, while the other uses the AD9617 current feedback op amp. Both op amps were designed on the same complementary bipolar (CB) process. Notice that both op amps have a closed loop bandwidth of about 200MHz and 0.1% settling time of

about 10ns. The important performance differences relate to noise and distortion. Notice that although the AD9622 op amp distortion is close to that of the AD9617 at 20MHz, it becomes much worse at lower frequencies (see Figure 10.12).

VOLTAGE FEEDBACK VERSUS CURRENT FEEDBACK OP AMP COMPARISON FOR GAIN OF +2 CONFIGURATION

	Current Feedback (AD9617)	Voltage Feedback (AD9622)
Distortion @ 20MHz	- 67dBc	- 66dBc
Distortion @ 1MHz	- 92dBc	- 76dBc
Input Current Noise	29pA / $\sqrt{\text{Hz}}$ ( - Input)	3.2pA / $\sqrt{\text{Hz}}$ ( + and - Input)
Input Voltage Noise	1.2nV / $\sqrt{\text{Hz}}$	3.5nV / $\sqrt{\text{Hz}}$
RMS Output Noise	172 $\mu\text{V}$ rms, 0.1 to 200MHz	98 $\mu\text{V}$ rms, 0.1 to 200MHz
Bandwidth (SSBW)	200MHz	220MHz
Slew Rate	1400V/ $\mu\text{s}$	1500V/ $\mu\text{s}$
Settling Time to 0.1%	10ns	8ns
Input Offset Voltage	1mV	2mV

Figure 10.11

DISTORTION OF AD9622 VOLTAGE FEEDBACK OP AMP AND AD9617 CURRENT FEEDBACK OP AMP FOR

$R_L = 100\Omega$  AND  $V_{OUT} = 2V_{p-p}$ , NOISE GAIN = 2, BANDWIDTH = 200MHz

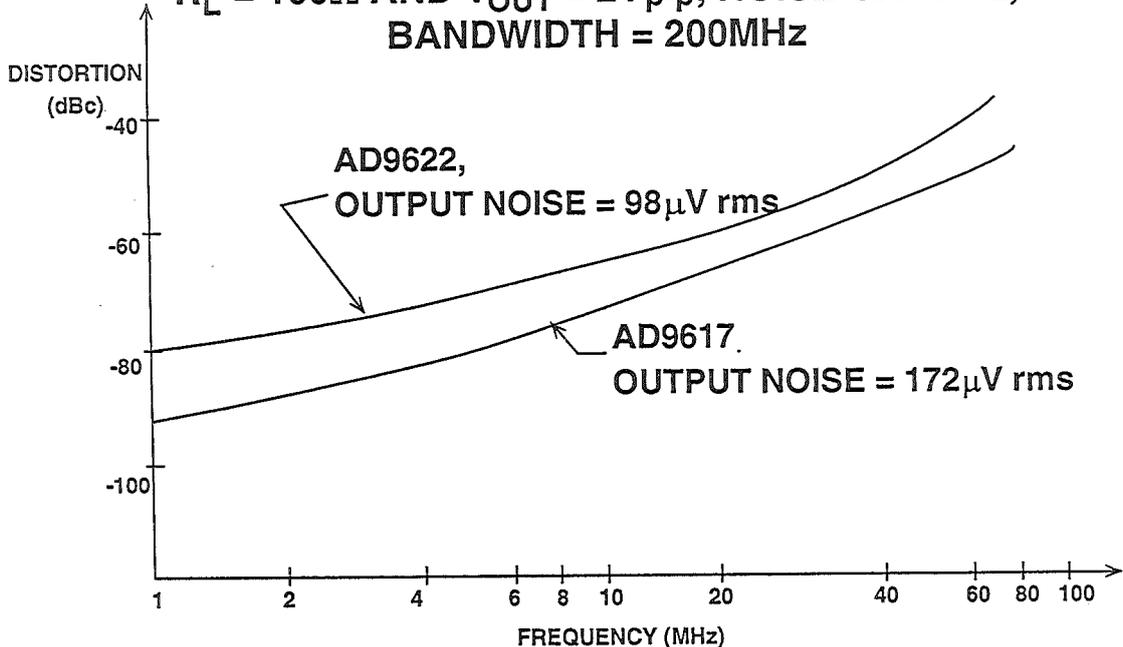


Figure 10.12

The other important difference is that the total output noise of the AD9622 is much less than that of the AD9617. The dominant source of output noise for the AD9622 is the input voltage noise. On the other hand, the AD9617 inverting input current noise (which flows through the feedback resistor) is the dominant contributor to the total output noise. In both circuits, the Johnson noise of the resistors may be neglected.

The inverting and non-inverting frequency response of the AD9622 voltage feedback op amp is shown in Figure 10.13. Note that the bandwidth is approximately inversely proportional to the gain as would be expected for the voltage feedback architecture.

### AD9622 VOLTAGE FEEDBACK OP AMP FREQUENCY RESPONSE

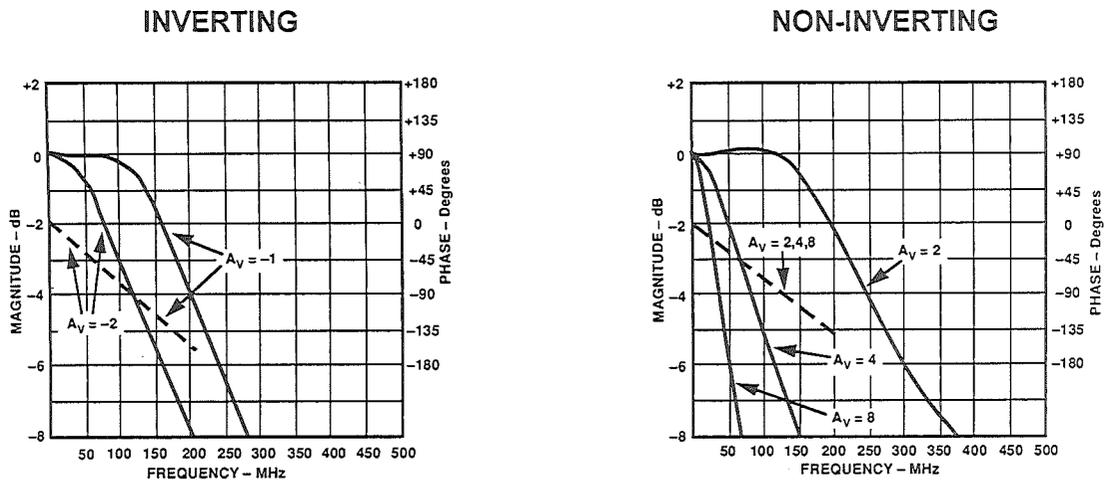


Figure 10.13

In op amps such as the AD9617 and AD9622, settling time is often of interest. The short and long term settling time of the AD9622 is shown in Figure 10.14. Notice that the long term settling time exhibits a thermal tail of about 0.03%. This is common in high speed

amplifiers because of the temperature change caused by the power changing in the output stage. (The data in Figure 10.14 was taken with a 100Ω load). The thermal tail is almost non-existent if the load is greater than about 500Ω.

## AD9622 OP AMP SHORT AND LONG-TERM SETTLING TIME

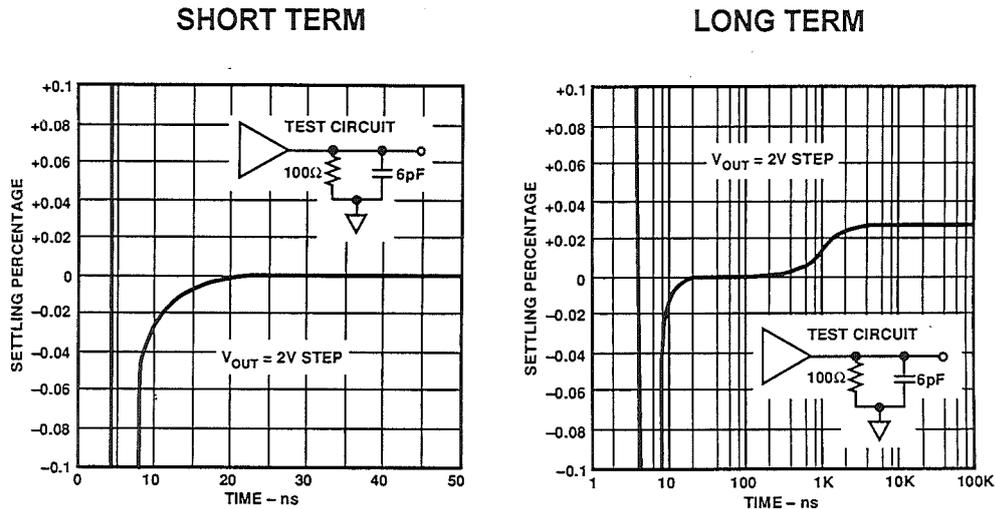


Figure 10.14

From the discussions so far, it should be clear that the differences between voltage and current feedback op amp performance are somewhat more subtle than one would suspect from recent literature. Other than the tradeoffs between noise and distortion discussed above, current feedback amplifiers are uniquely suited to applications where constant bandwidth must be achieved over a fairly wide range of gain settings. (This is true only if the value of the feedback resistor remains constant! If the feedback resistor is increased above the nominal value, the bandwidth is

approximately inversely proportional to the resistor increase. Lower than optimum feedback resistors usually cause instability in current feedback op amps).

In order to take advantage of the lower current noise and increased flexibility of voltage feedback amplifiers, the same basic circuit design may be optimized for various gains. This was done in the AD9621, AD9622, AD9623, and AD9624 op amp family. The performance of these amplifiers is summarized in Figure 10.15.

## AD9621,AD9622,AD9623,AD9424 VOLTAGE FEEDBACK OP AMPS

Parameter	AD9621	AD9622	AD9623	AD9624	Units
Minimum Stable Gain	+1	+2	+4	+6	V/V
Harmonic Distortion (20MHz)	-52	-66	-64	-66	dB
Large Signal Bandwidth (4V p-p)	130	160	190	200	MHz
SSBW (0.5Vp-p)	350	220	270	300	MHz
Slew Rate	1200	1500	2100	2200	V/ $\mu$ s
Rise / Fall Time (0.5V Step)	2.4	1.7	1.6	1.5	ns
Settling Time (to 0.1%/0.01%)	7/11	8/14	8/14	8/14	ns
Input Noise (0.1MHz - 200MHz)	80	49	36	32	$\mu$ Vrms

Figure 10.15

In order to achieve these levels of high speed performance, however, careful attention must be given to good high speed circuit layout, grounding, and decoupling techniques. Figure 10.16 shows the standard inverting and noninverting connections for the AD9622. Notice that since the inputs are symmetrical, the effects of the input bias currents can be minimized by the addition of the proper resistor in series with the noninverting input. The diagrams of Figure 10.16 show a bootstrap capacitor,  $C_B$  (normally 0.001 $\mu$ F), which may be added to further enhance settling time. This capacitor connects to the internal high impedance nodes of the amplifier. Using this capacitor will reduce the large signal (4V) step output

settling time by 3ns to 5ns for 0.05% or greater accuracy. For settling accuracy less than 0.05% or for smaller step sizes, its effect will be less apparent. Under fast slew conditions, this capacitor forces the internal signal (initial step) amplitude to be controlled by the "on" (slewed) transistor, preventing its complement from completely turning off. This allows for faster settling time of these internal nodes and also the output. In the frequency domain, total (high frequency) distortion will be approximately the same with or without  $C_B$ . Typically, the 3rd harmonic will be greater than the 2nd without  $C_B$ . This condition will be reversed with  $C_B$  in place.

## PROPER LAYOUT AND DECOUPLING IS CRITICAL TO HIGH SPEED OP AMP PERFORMANCE

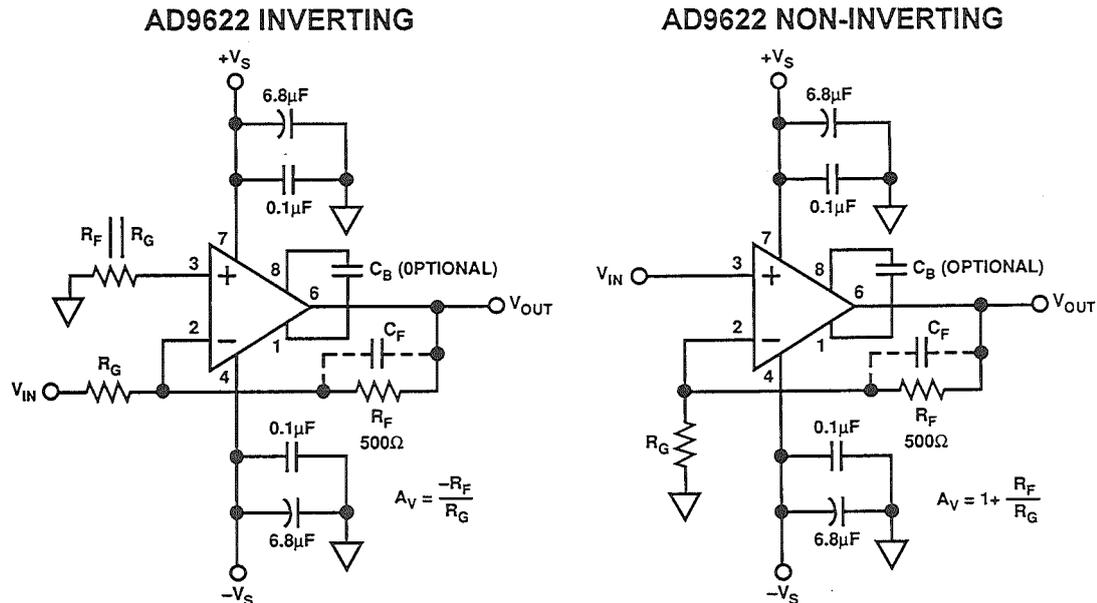


Figure 10.16

As with all wide bandwidth components, good PC board layout is critical to obtain the best dynamic performance with these amplifiers. The ground plane in the area of the op amp and its associated components should cover as much of the component side of the board as possible (or first interior ground layer of a multilayer board).

The ground plane should be removed in the area of the inputs and  $R_F$  and  $R_G$  to minimize stray capacitance at the input. The same precaution should be used for  $C_B$  if it is used. Each power supply trace should be decoupled close to the package with a  $0.1\mu\text{F}$  ceramic (preferably surface mount), plus a  $6.8\mu\text{F}$  tantalum capacitor within 0.5".

All lead lengths for input, output, and feedback resistor should be kept as short as possible. All gain setting resis-

tors should be chosen for low values of parasitic capacitance and inductance, i.e., microwave resistors (buffed metal film rather than laser-trimmed spiral-wound) and/or carbon resistors.

Microstrip techniques should be used for all input and output lead lengths in excess of one inch. Sockets should be avoided if at all possible because of their parasitic capacitance and inductance. If sockets are necessary, individual *pin sockets* such as AMP p/n 6-330808-3 should be used. These contribute far less stray capacitance and inductance than molded socket assemblies.

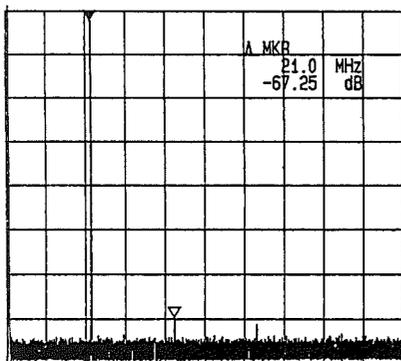
The effects of inadequate decoupling on harmonic distortion performance are dramatically illustrated in Figure 10.17. Photo A shows the spectral output of the AD9617 driving a  $100\Omega$

load with proper decoupling. If the decoupling is removed, the distortion is greatly increased as shown in Photo B of the same figure. Unlike lower frequency amplifiers, the power supply rejection ratio of high frequency amplifiers is generally fairly poor at high frequencies. For example, at 100MHz, the power supply rejection ratio of both the AD9617 and the AD9622 is less

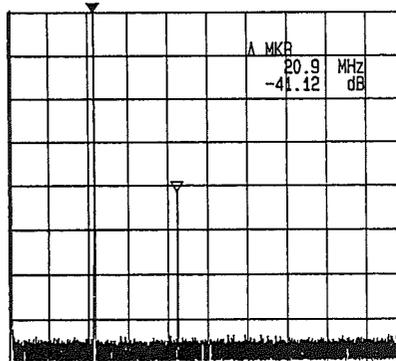
than 20dB. This is the primary reason for the degradation in performance with inadequate decoupling. The change in output signal produces a corresponding signal-dependent load current change. The corresponding change in power supply voltage due to inadequate decoupling produces a signal-dependent error in the output which manifests itself as an increase in distortion.

### EFFECTS OF INADEQUATE DECOUPLING ON HARMONIC DISTORTION PERFORMANCE OF AD9617

(A) PROPER DECOUPLING



(B) NO DECOUPLING



VERTICAL SCALE: 10dB/div.  
HORIZONTAL SCALE: 10MHz/div.

Figure 10.17

Inadequate decoupling can also severely affect the pulse response of high speed amplifiers such as the AD9617. Figure 10.19 shows the effects of removing all

decoupling capacitors on the AD9717 in its evaluation board. Notice the severe ringing on the pulse response.

**AD9617 POWER SUPPLY REJECTION RATIO (PSRR)  
AND COMMON MODE REJECTION RATIO (CMRR)**

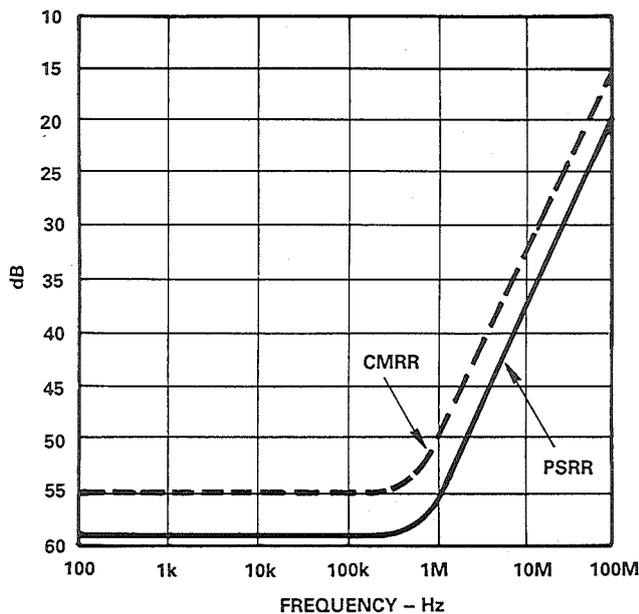
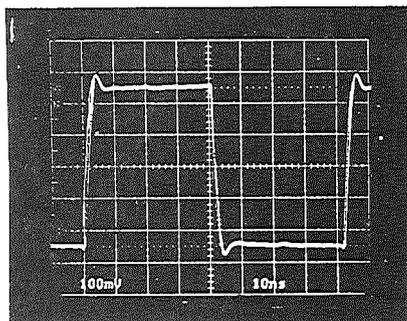


Figure 10.18

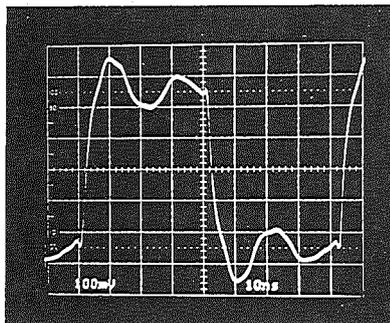
10

**EFFECT OF INADEQUATE DECOUPLING  
ON PULSE RESPONSE OF AD9617 OP AMP**

PROPER DECOUPLING



NO DECOUPLING



VERTICAL SCALE: 100mV/div.  
HORIZONTAL SCALE: 10ns/div.

Figure 10.19

The effects of stray capacitance on the inverting input of the AD9617 is shown in Figure 10.20. Careful examination of the waveform indicates sustained oscillation at about 650MHz. This was verified using a spectrum analyzer. High frequency oscillation at hundreds of megahertz is a good indicator of poor layout, grounding and decoupling practices. Unfortunately, you may never actually observe it unless you have a scope or spectrum analyzer which has sufficient bandwidth. Un-

wanted oscillations at RF frequencies will probably be rectified and averaged by devices to which the oscillating signal is applied. This is referred to as *RF rectification* and will create small unexplained dc offsets which may even be a function of moving your hand over the PC board. It is absolutely essential when building circuits using high frequency components to have high bandwidth test equipment and use it to check for oscillation at frequencies well beyond the signals of interest.

### EFFECT OF 4.7pF STRAY INVERTING INPUT CAPACITANCE ON AD9617 SHOWING 667MHz OSCILLATION

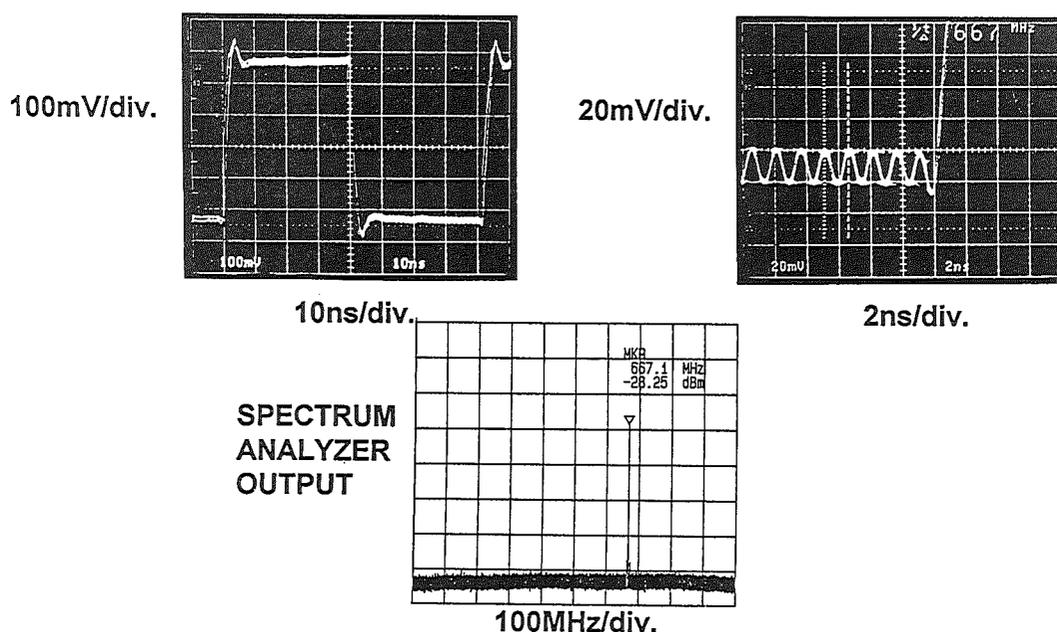


Figure 10.20

Many of these problems occur in the prototype phase due to a disregard for high frequency layout and decoupling techniques. Figure 10.21 shows a hand-wired prototype board which gives excellent performance in spite of its lack of esthetic appeal. The op amp is

mounted upside down on a solid copper-clad board with the leads bent back. The signals are connected with short point-to-point wiring. The characteristic impedance of a wire over a ground plane is about 120Ω, although this may vary as much as ±40% depending on the

distance from the plane. The decoupling capacitors are connected directly from the op amp power pins to the copper-clad ground. When working at frequencies of several hundred MHz, it is a good idea to use only one side of the board for ground. Many people drill holes in the board and connect both sides together with short pieces of wire

soldered to both sides of the board. If care is not taken, this may result in unexpected ground loops between the two sides of the board, especially at RF frequencies. This approach to prototyping is often called the *deadbug* method because the upside-down ICs look like deceased insects.

### PHOTOGRAPH OF A GOOD HIGH SPEED OP AMP PROTOTYPE LAYOUT

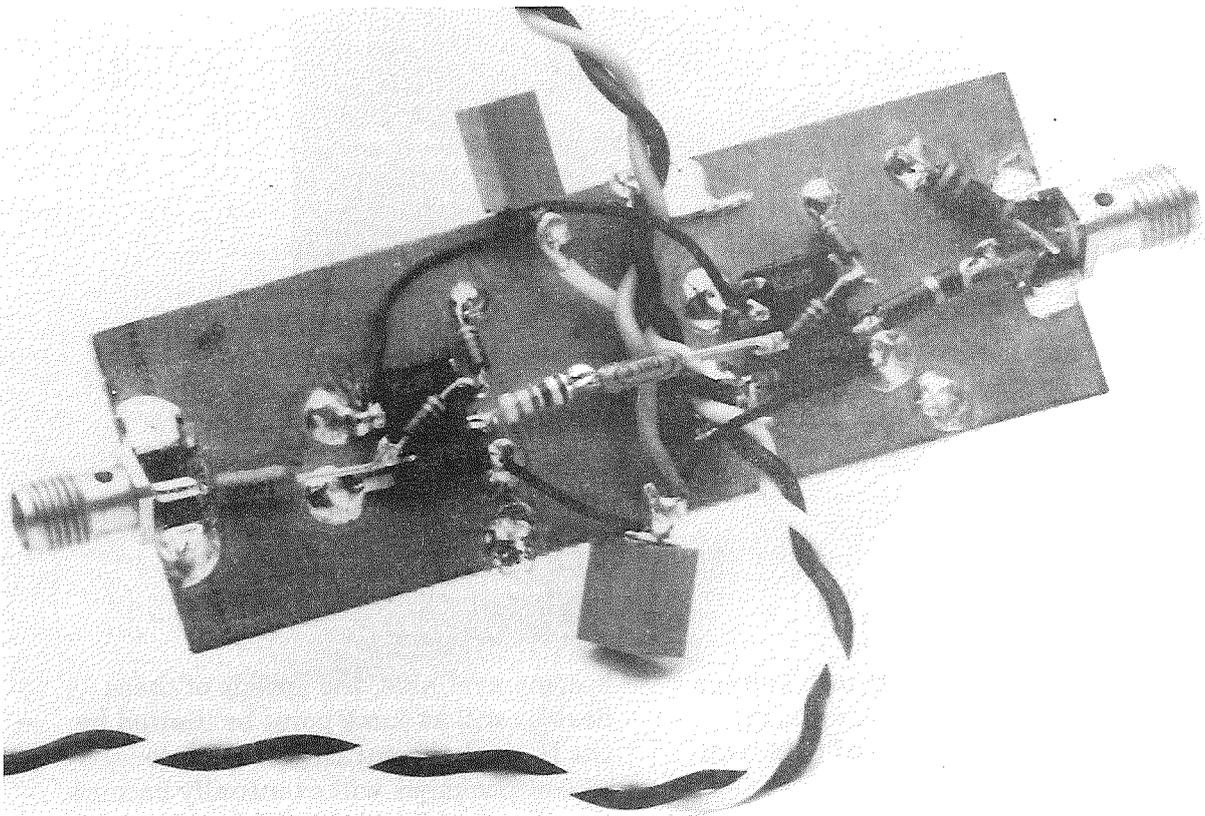


Figure 10.21

In order to properly evaluate the performance of high speed amplifiers, Analog Devices supplies evaluation boards at a nominal charge. The AD9617 evaluation board is shown in Figure 10.22. Use of these boards is encouraged during the initial system design phases.

The actual artwork for the layouts is available free of charge from Analog Devices. In many cases, the artwork appears on the amplifier datasheets, and sometimes PC graphics files are also available

## PHOTOGRAPH OF AD9617 OP AMP EVALUATION BOARD

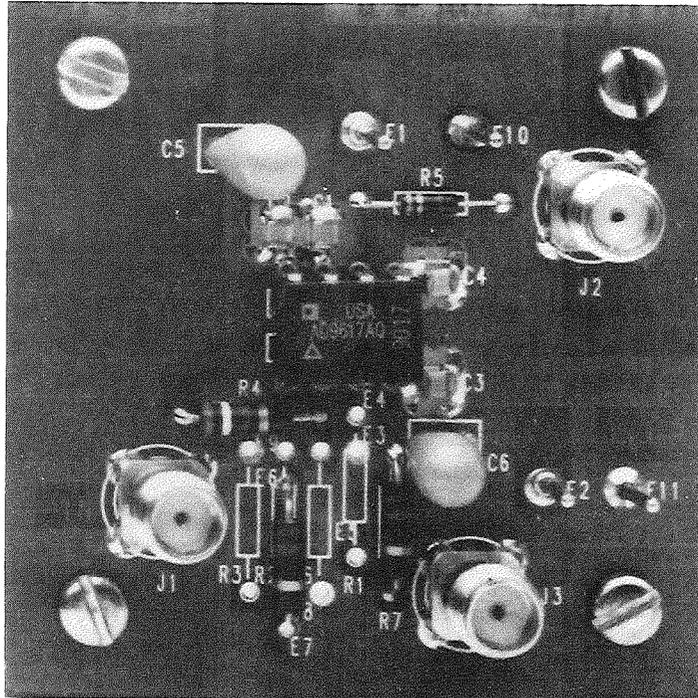


Figure 10.22

So far, we have examined both voltage and current feedback amplifiers, discussed their advantages and disadvantages, and presented some guidelines for successful application based on system requirements. It should be clear that as higher frequency processes

become available, the speeds of these devices will at some point be limited by the IC package parasitics. Surface mount packages will eventually become the standard for high performance ultra high speed op amps.

## SUMMARY OF HIGH SPEED VOLTAGE FEEDBACK OP AMP CHARACTERISTICS

- Symmetrical Inputs
- Equalization of Source Resistances Generally Reduces Effects of Input Bias Currents
- Largest Noise Source is Input Voltage Noise
- Flexible Feedback Networks Allow Many Tradeoffs
- Constant Gain-Bandwidth Product
- May be Used as Integrators in Active Filters
- Sensitive to Stray Capacitance on Inputs and Outputs
- Proper Layout, Grounding, and Decoupling is Essential to Achieve Specified Performance

Figure 10.23

## SUMMARY OF HIGH SPEED CURRENT FEEDBACK OP AMPS CHARACTERISTICS

- Non-Symmetrical Inputs
- Input Bias Current Cancellation Schemes Don't Work
- Inverting Input Current Noise Usually Dominates
- Feedback Network Fixed for Optimum Performance
- Difficult to Use as Integrators
- Bandwidth Remains Relatively Constant for Different Gains
- Sensitive to Stray Capacitance on Inputs and Outputs
- Proper Layout, Grounding, and Decoupling is Essential to Achieve Specified Performance

Figure 10.24

## HIGH SPEED BUFFER AMPLIFIERS

In the early days of high speed circuits, simple emitter followers were often used as high speed buffers. The term *buffer* was generally accepted to mean a unity-gain open loop amplifier such as the National LH-0033 hybrid. This device was a complementary emitter follower with an input FET source

follower. The schematic of this device and a purely bipolar version is shown in Figure 10.25. Both devices achieved bandwidths of about 100MHz at fairly respectable levels of harmonic distortion. However, they suffered from dc and ac non-linearities when loaded with impedances much less than 500Ω.

### EARLY OPEN-LOOP, 100MHz BANDWIDTH HYBRID BUFFERS

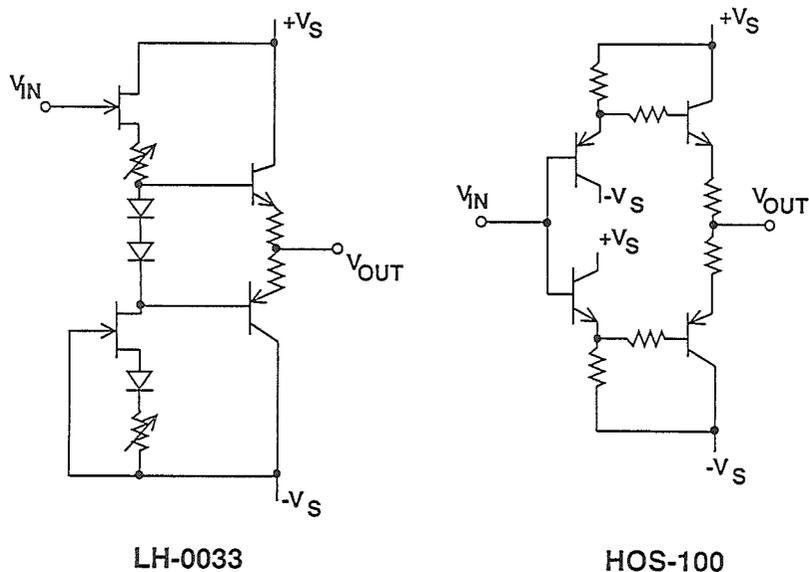


Figure 10.25

An early IC implementation of these functions was the Precision Monolithic, Inc. BUF-03 shown in Figure 10.26.

(PMI is now a division of Analog Devices.) This open-loop IC buffer achieves a bandwidth of about 50MHz.

## EARLY IC OPEN-LOOP, 50MHz BUFFER, THE BUF-03

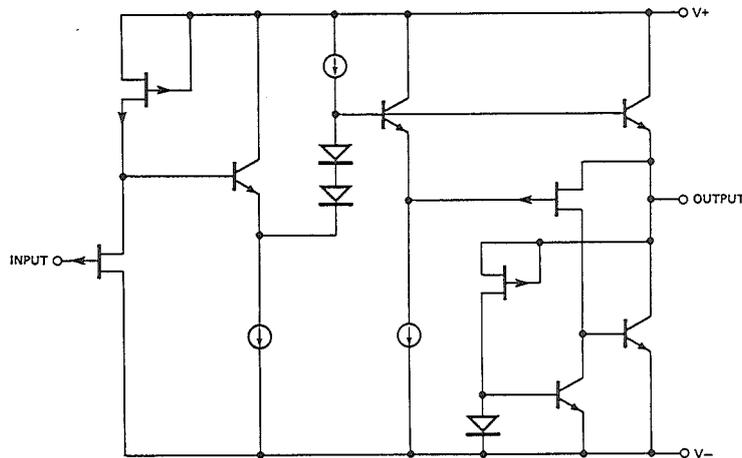


Figure 10.26

One of the problems with open loop buffers is that although high bandwidths may be achieved, these devices cannot take advantage of the “curative” effects of negative feedback. Distortion and DC performance suffers considerably when open loop buffers are loaded with typical video impedance levels of 50, 75, or 100Ω.

As IC processes evolved and high speed general-purpose op amps became available, it became possible to build relatively high speed buffers using op amps as building blocks. Usually, however, the general-purpose op amps are compensated to operate over a wide range of gains and feedback conditions. Therefore, bandwidth suffers somewhat at low gains, especially in the unity-gain non-inverting mode, because of the additional external compensation usually required.

The AD9620 is a 600MHz voltage feedback op amp which is optimized for maximum performance as a unity-gain buffer. The feedback resistor is on chip for minimum parasitic effects. The impressive specifications are summarized in Figure 10.27. Although the open loop dc gain of the AD9620 is only about 2000, the proprietary voltage feedback circuit design of the AD9620 (Patents Pending) insures that this value is relatively constant under a variety of load and frequency conditions. DC endpoint linearity for a 100Ω load is better than 60ppm (85dB) as shown in Figure 10.28. Frequency response and settling time for the AD9620 is shown in Figure 10.29, and harmonic distortion performance in Figure 10.30.

### AD9620 ULTRALOW DISTORTION 600MHz BUFFER KEY SPECIFICATIONS

- Gain Accuracy: 0.994V/V
- Wide Bandwidth: 600MHz
- Slew Rate: 2200V/ $\mu$ s
- Ultralow Distortion: -73dBc @ 20MHz, -91dBc @ 2.3MHz
- Fast Settling Time: 8ns to 0.02%
- $\pm$ 40mA Output Current
- Low Noise: 2nV/ $\sqrt$ Hz

Figure 10.27

### AD9620 BUFFER DC ENDPOINT LINEARITY ERROR FOR 100 $\Omega$ AND 200 $\Omega$ LOADS INDICATES GOOD OPEN-LOOP GAIN STABILITY OVER SIGNAL RANGE

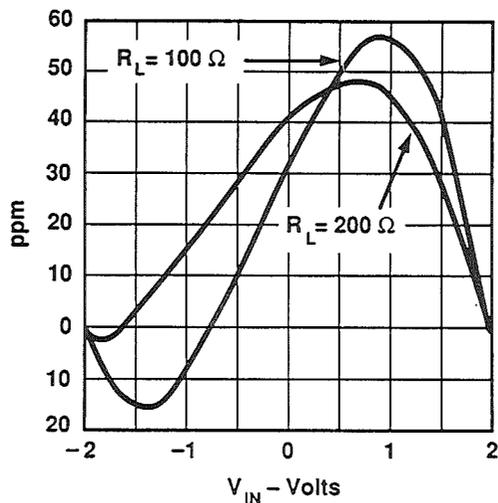


Figure 10.28

## AD9620 BUFFER FREQUENCY RESPONSE AND SETTLING TIME

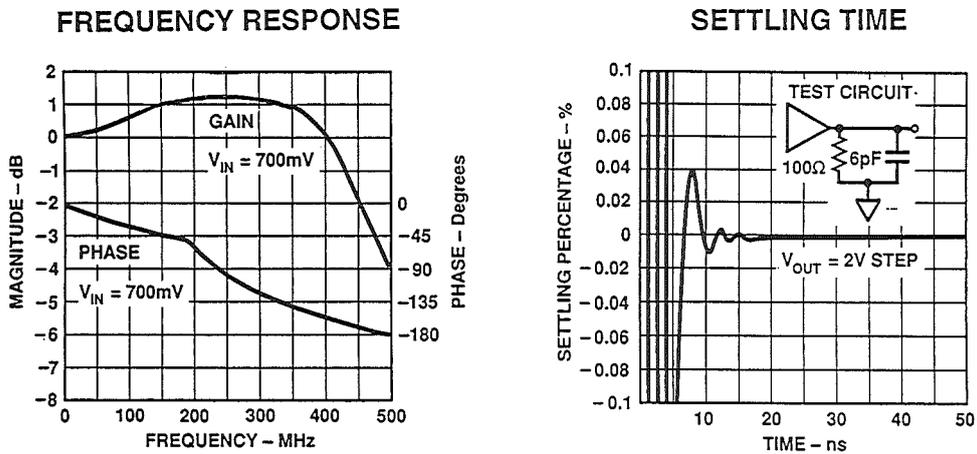


Figure 10.29

## AD9620 BUFFER HARMONIC DISTORTION PERFORMANCE

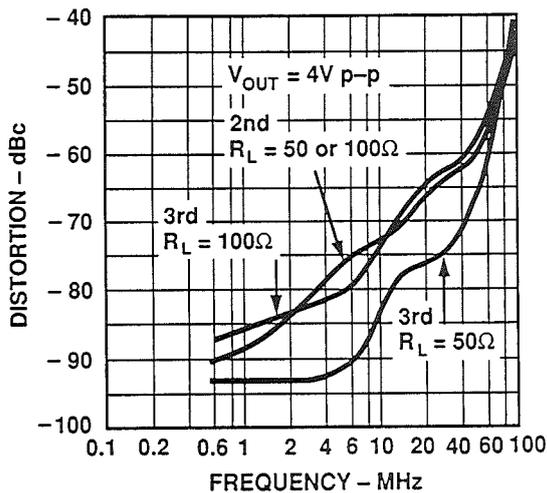


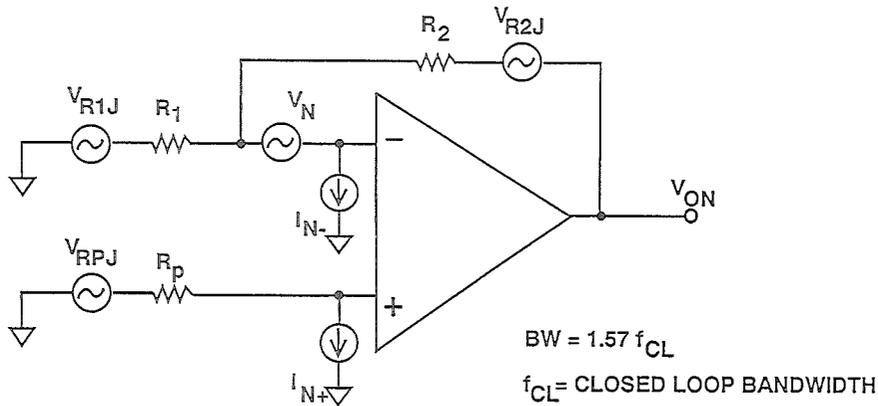
Figure 10.30

## HIGH SPEED OP AMP NOISE MODELS

A generalized noise model for an op amp is shown in Figure 10.31. This model is especially suited to high speed op amp circuits where low frequency 1/f noise may usually be neglected. In order for the model to accurately predict results, the closed-loop frequency re-

sponse should be flat within a few dB up to the closed-loop bandwidth frequency. In calculating the total output rms noise, the high frequency noise will dominate, and any significant peaking in the frequency response must be considered in the calculation.

### OP AMP NOISE MODEL



$$V_{ON} = \sqrt{BW} \sqrt{I_{N-}^2 R_2^2 + I_{N+}^2 R_p^2 \left[1 + \frac{R_2}{R_1}\right]^2 + V_N^2 \left[1 + \frac{R_2}{R_1}\right]^2 + 4kTR_2 + 4kTR_1 \left[\frac{R_2}{R_1}\right]^2 + 4kTR_p \left[1 + \frac{R_2}{R_1}\right]^2}$$

Figure 10.31

There are a number of useful simplifications which can be made when estimating the output noise of a wideband op amp. Because the individual noise sources contributing to the output noise add on an rms basis, sources which are 3 to 5 times less than the largest may

be neglected in the calculation. In high frequency circuits such as those previously discussed, resistor values seldom exceed 500Ω, therefore resistor Johnson noise is usually neglected. The useful simplifications are summarized in Figure 10.32.

## SIMPLIFICATIONS IN ESTIMATING RMS OUTPUT NOISE FOR HIGH SPEED OP AMPS

- Neglect Resistor Johnson Noise if  $R < 500\Omega$
- For Voltage Feedback Op Amps, Major Noise Source is Input Voltage Noise. Reflect to Output by Multiplying by Noise Gain.
- For Current Feedback Op Amps, Major Noise Source is Inverting Input Current Noise. Reflect to Output by Multiplying by Feedback Resistor Value.
- Use Closed-Loop Small Signal Bandwidth in Calculation

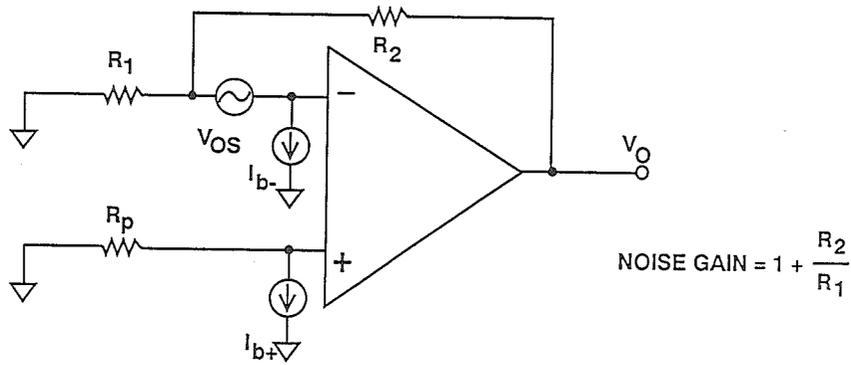
Figure 10.32

### OP AMP DC MODEL

A generalized model for calculating the total output offset voltage of an op amp is shown in Figure 10.33. These equations are applicable for either voltage or current feedback op amps. However, the inputs of a voltage feedback op amp are symmetrical; therefore, the input bias currents are approximately equal

provided internal bias compensation is not used. Due to the non-symmetrical input structure of the current feedback op amp, however, the bias currents are usually different, and therefore, appropriate values as given in the data sheet must be used for  $I_{b+}$  and  $I_{b-}$ .

OP AMP OFFSET VOLTAGE MODEL



$$V_O = \pm V_{OS} \left[ 1 + \frac{R_2}{R_1} \right] \pm I_{b+} R_p \left[ 1 + \frac{R_2}{R_1} \right] \pm I_{b-} R_2$$

Figure 10.33

Although small dc offset shifts are not a problem in most high frequency applications, significant shifts may produce clipping in an ADC system as shown in Figure 10.34. The output drift of the

driving op amp must be summed with the offset drift of the ADC. If the drift is large, the gain must be reduced to prevent clipping, thereby reducing the overall dynamic range of the system.

## LARGE OFFSET SHIFTS REQUIRE GAIN REDUCTION TO PREVENT ADC CLIPPING

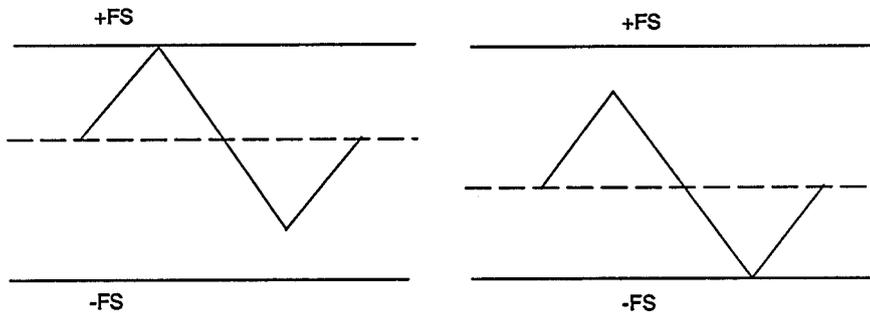


Figure 10.34

## LEVEL SHIFTING HIGH SPEED SIGNALS USING OP AMPS

In addition to providing gain, op amps are often used as level shifters in signal processing systems. Figure 10.35 shows two configurations for the inverting mode. Injecting an offset current into the inverting input is probably the simplest method, but the penalty is an

increase in noise gain due to  $R_3$  and the potentiometer resistance. The resulting increase in noise gain may be reduced by making  $V_R$  large enough so that  $R_3$  can be made much greater than  $R_1$  and  $R_2$ .

## INVERTING OP AMP LEVEL SHIFTERS

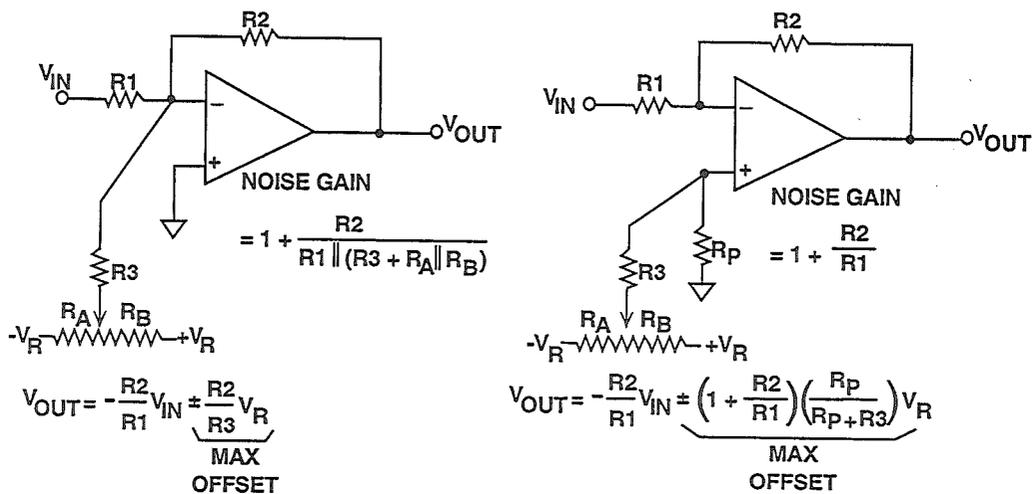


Figure 10.35

The second circuit in Figure 10.35 shows how to create an output offset by injecting the offset current into the non-inverting input. This circuit results in no increase in noise gain, but requires the addition of  $R_p$  which may affect the circuit bandwidth if it is not adequately bypassed.

The circuit shown in Figure 10.36 is used to level shift the output when

using the op amp in the non-inverting mode. This circuit works well for small offsets where  $R_3$  can be made much greater than  $R_1$ . Otherwise, the signal gain will be affected as the offset potentiometer is adjusted. The gain may be stabilized, however, if  $R_3$  is connected to a fixed low impedance reference voltage source.

## NON-INVERTING OP AMP LEVEL SHIFTERS

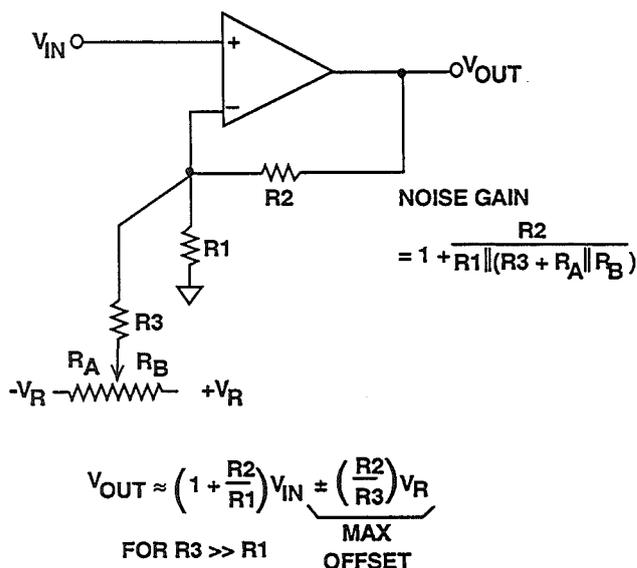


Figure 10.36

Many op amps have offset null pins available for correcting small offsets. The AD845 CBFET op amp is a good example of such an op amp. Key specifications are given in Figure 10.37 and a functional schematic in Figure 10.38. The AD845 is laser trimmed for a maximum offset voltage of  $250\mu\text{V}$  and an offset temperature coefficient of  $5\mu\text{V}/^\circ\text{C}$ . The configuration shown in Figure 10.38 may be used to null out

this small offset voltage, but the resulting imbalance in the current in the input differential FET pair produces an additional  $4\mu\text{V}/^\circ\text{C}$  temperature for every millivolt nulled. Therefore, nulling out  $250\mu\text{V}$  of offset voltage will increase the offset temperature coefficient by approximately  $1\mu\text{V}/^\circ\text{C}$ . In no case should the offset null pins be used to correct for large system offsets!

## AD845 CBFET KEY SPECIFICATIONS

- 0.25mV Max Input Offset Voltage
- $5\mu\text{V}/^\circ\text{C}$  Max Offset Voltage Drift
- 0.5nA Input Bias Current
- 350ns Settling Time to 0.01%
- 16MHz Unity-Gain Bandwidth
- $25\text{nV}/\sqrt{\text{Hz}}$ ,  $2\text{pA}/\sqrt{\text{Hz}}$  Noise at 1kHz

Figure 10.37

**CAREFUL USE OF OP AMP NULL PINS IS EFFECTIVE  
IN REDUCING SMALL OFFSET VOLTAGES BUT MAY  
INCREASE TEMPERATURE COEFFICIENT**

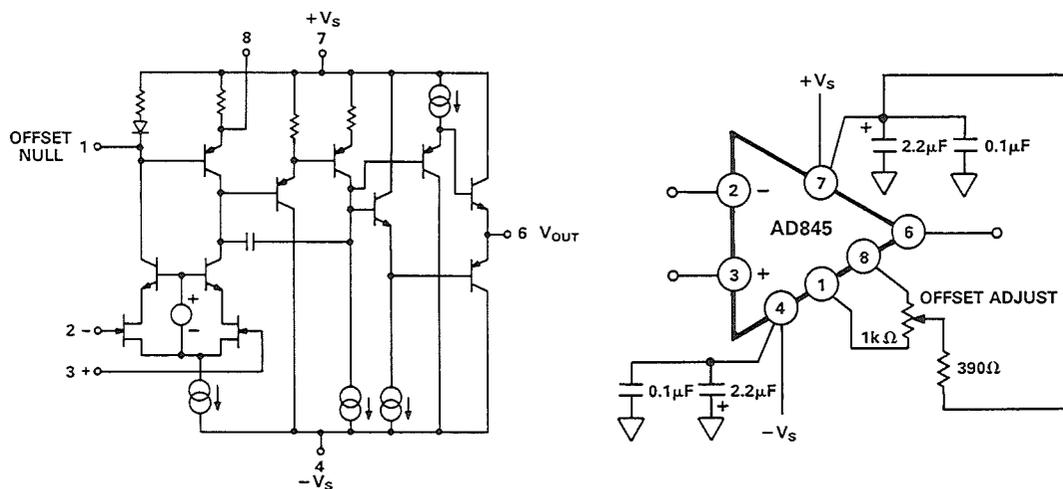


Figure 10.38

**REFERENCES**

1. Walt Jung, "Op Amps in Line-Driver and Receiver Circuits Part 1, Video Applications," *Analog Dialogue* 26-2, pp. 13-20, 1992.
2. Thomas M. Frederiksen, *Intuitive Operational Amplifiers*, McGraw Hill, 1988.
3. D. Stout, M. Kaufman, *Handbook of Operational Amplifier Circuit Design*, New York, McGraw-Hill, 1976.
4. Joe Buxton, *Careful Design Tames High-Speed Op Amps*, *Electronic Design*, April 11, 1991.
5. J. Dostal, *Operational Amplifiers*, Elsevier Scientific Publishing, New York, 1981.
6. *High Speed Design Seminar Notes*, 1990, Analog Devices.
7. *1992 Amplifier Applications Guide*, 1992, Analog Devices.
8. Walter G. Jung, *IC Op Amp Cookbook, Third Edition*, SAMS (Division of Macmillan, Inc.), 1986.

