

SECTION 1

HIGH SPEED OPERATIONAL AMPLIFIERS

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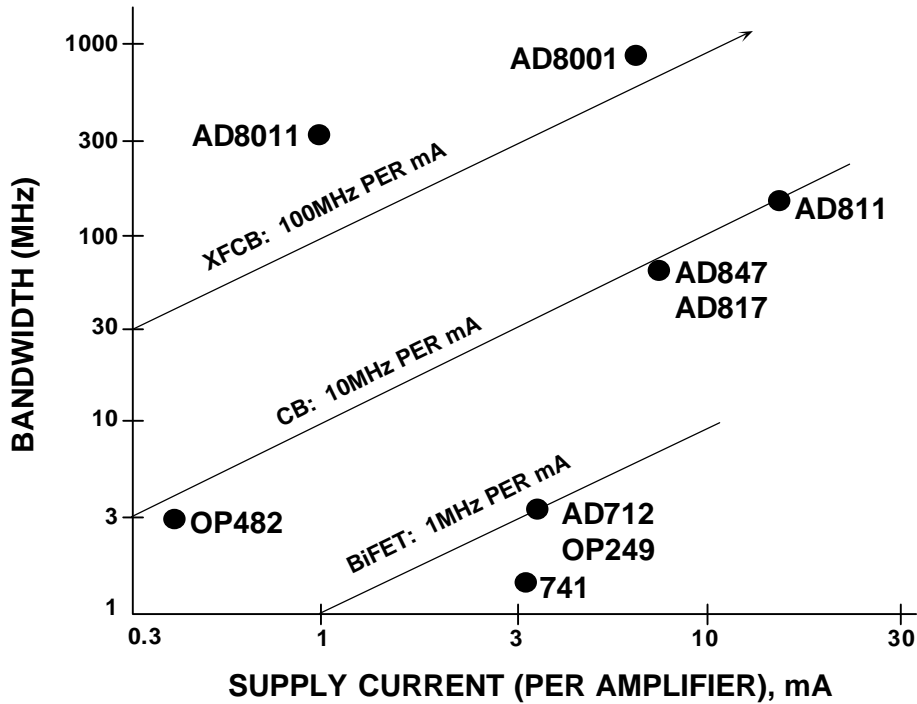
INTRODUCTION

High speed analog signal processing applications, such as video and communications, require op amps which have wide bandwidth, fast settling time, low distortion and noise, high output current, good DC performance, and operate at low supply voltages. These devices are widely used as gain blocks, cable drivers, ADC pre-amps, current-to-voltage converters, etc. Achieving higher bandwidths for less power is extremely critical in today's portable and battery-operated communications equipment. The rapid progress made over the last few years in high-speed linear circuits has hinged not only on the development of IC processes but also on innovative circuit topologies.

The evolution of high speed processes by using amplifier bandwidth as a function of supply current as a figure of merit is shown in Figure 1.1. (In the case of duals, triples, and quads, the current per amplifier is used). Analog Devices BiFET process, which produced the AD712 and OP249 (3MHz bandwidth, 3mA current), yields about 1MHz per mA. The CB (Complementary Bipolar) process (AD817, AD847, AD811, etc.) yields about 10MHz/mA of supply current. Ft's of the CB process PNP transistors are about 700MHz, and the NPN's about 900MHz.

The latest generation complementary bipolar process from Analog Devices is a high speed dielectrically isolated process called XFCB (eXtra Fast Complementary Bipolar). This process (2-4 GHz Ft matching PNP and NPN transistors), coupled with innovative circuit topologies allow op amps to achieve new levels of cost-effective performance at astonishing low quiescent currents. The approximate figure of merit for this process is typically 100MHz/mA, although the AD8011 op amp is capable of 300MHz bandwidth on 1mA of supply current due to its unique two-stage current-feedback architecture.

AMPLIFIER BANDWIDTH VERSUS SUPPLY CURRENT FOR ANALOG DEVICES' PROCESSES



1.1

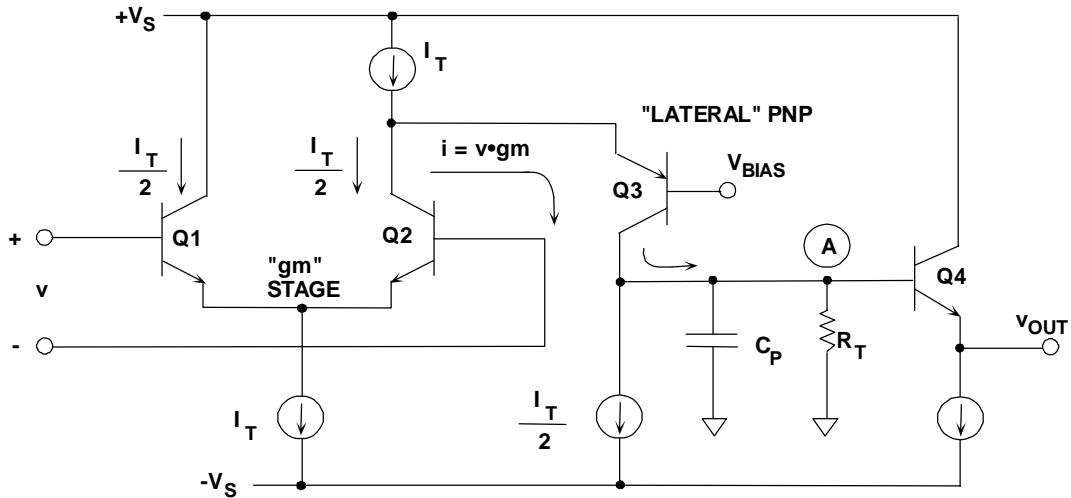
In order to select intelligently the correct op amp for a given application, an understanding of the various op amp topologies as well as the tradeoffs between them is required. The two most widely used topologies are voltage feedback (VFB) and current feedback (CFB). The following discussion treats each in detail and discusses the similarities and differences.

VOLTAGE FEEDBACK (VFB) OP AMPS

A voltage feedback (VFB) op amp is distinguished from a current feedback (CFB) op amp by circuit topology. The VFB op amp is certainly the most popular in low frequency applications, but the CFB op amp has some advantages at high frequencies. We will discuss CFB in detail later, but first the more traditional VFB architecture.

Early IC voltage feedback op amps were made on "all NPN" processes. These processes were optimized for NPN transistors, and the "lateral" PNP transistors had relatively poor performance. Lateral PNPs were generally only used as current sources, level shifters, or for other non-critical functions. A simplified diagram of a typical VFB op amp manufactured on such a process is shown in Figure 1.2.

VOLTAGE FEEDBACK (VFB) OP AMP DESIGNED ON AN "ALL NPN" IC PROCESS



$$\blacksquare g_m = I_C \cdot \frac{q}{kT}$$

$$\blacksquare v_{OUT} = \frac{i}{j\omega C_P} = v \cdot \frac{g_m}{j\omega C_P} @ HF$$



1.2

The input stage is a differential pair consisting of either a bipolar pair (Q1, Q2) or a FET pair. This "g_m" (transconductance) stage converts the small-signal differential input voltage, v, into a current, i, i.e., its transfer function is measured in units of conductance, 1/Ω, (or mhos). The small signal emitter resistance, r_e, is approximately equal to the reciprocal of the small-signal g_m. The formula for the small-signal g_m of a single bipolar transistor is given by the following equation:

$$g_m = \frac{1}{r_e} = \frac{q}{kT} (I_C) = \frac{q}{kT} \left(\frac{I_T}{2} \right), \text{ or}$$

$$g_m \approx \left(\frac{1}{26\text{mV}} \right) \left(\frac{I_T}{2} \right).$$

I_T is the differential pair tail current, I_C is the collector bias current (I_C = I_T/2), q is the electron charge, k is Boltzmann's constant, and T is absolute temperature. At +25°C, V_T = kT/q = 26mV (often called the Thermal Voltage, V_T).

As we will see shortly, the amplifier unity gain-bandwidth product, f_u, is equal to g_m/2πC_p, where the capacitance C_p is used to set the dominant pole frequency. For this reason, the tail current, I_T, is made proportional to absolute temperature (PTAT). This current tracks the variation in r_e with temperature thereby making g_m independent of temperature. It is relatively easy to make C_p reasonably constant over temperature.

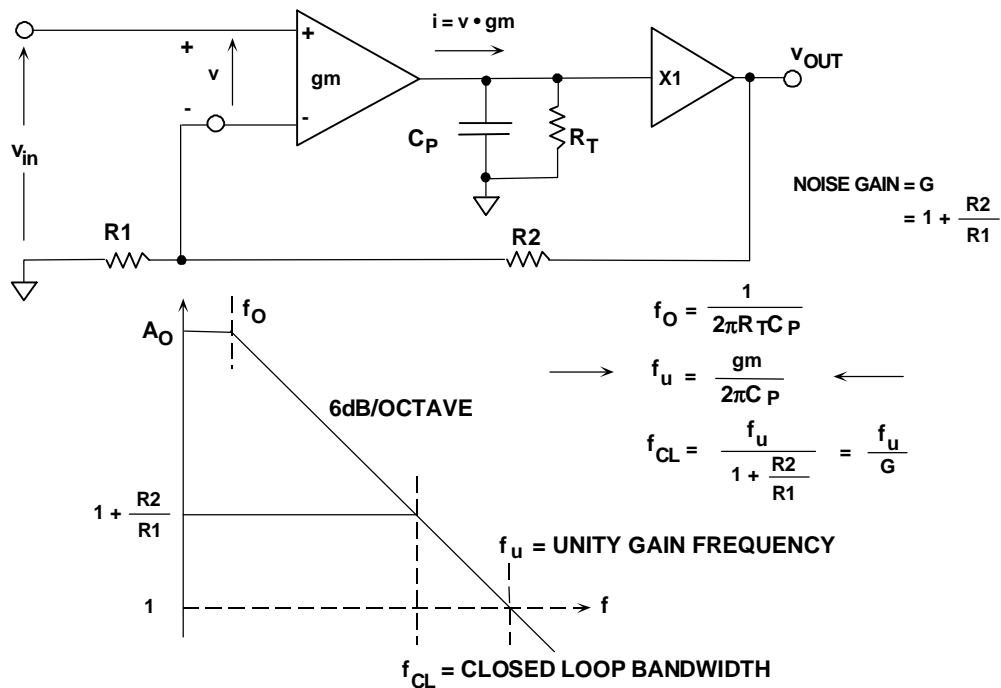
The output of one side of the g_m stage drives the emitter of a lateral PNP transistor (Q3). It is important to note that Q3 is not used to amplify the signal, only to level

shift, i.e., the signal current variation in the collector of Q2 appears at the collector of Q3. The output collector current of Q3 develops a voltage across high impedance node A. Cp sets the dominant pole of the frequency response. Emitter follower Q4 provides a low impedance output.

The effective load at the high impedance node A can be represented by a resistance, R_T , in parallel with the dominant pole capacitance, C_p . The small-signal output voltage, v_{out} , is equal to the small-signal current, i , multiplied by the impedance of the parallel combination of R_T and C_p .

Figure 1.3 shows a simple model for the single-stage amplifier and the corresponding Bode plot. The Bode plot is constructed on a log-log scale for convenience.

MODEL AND BODE PLOT FOR A VFB OP AMP



1.3

The low frequency breakpoint, f_0 , is given by:

$$f_0 = \frac{1}{2\pi R_T C_p}$$

Note that the high frequency response is determined solely by g_m and C_p :

$$v_{out} = v \cdot \frac{g_m}{j\omega C_p}$$

The unity gain-bandwidth frequency, f_u , occurs where $|v_{out}| = |v|$. Solving the above equation for f_u , assuming $|v_{out}| = |v|$:

$$f_u = \frac{g_m}{2\pi C_p}.$$

We can use feedback theory to derive the closed-loop relationship between the circuit's signal input voltage, v_{in} , and its output voltage, v_{out} :

$$\frac{v_{out}}{v_{in}} = \frac{1 + \frac{R_2}{R_1}}{1 + \frac{j\omega C_p}{g_m} \left(1 + \frac{R_2}{R_1}\right)}.$$

At the op amp 3dB closed-loop bandwidth frequency, f_{cl} , the following is true:

$$\frac{2\pi f_{cl} C_p}{g_m} \left(1 + \frac{R_2}{R_1}\right) = 1, \text{ and hence}$$

$$f_{cl} = \frac{g_m}{2\pi C_p} \left(\frac{1}{1 + \frac{R_2}{R_1}} \right), \text{ or}$$

$$f_{cl} = \frac{f_u}{1 + \frac{R_2}{R_1}}.$$

This demonstrates a fundamental property of VFB op amps: *The closed-loop bandwidth multiplied by the closed-loop gain is a constant*, i.e., the VFB op amp exhibits a constant gain-bandwidth product over most of the usable frequency range.

Some VFB op amps (called *de-compensated*) are unstable at unity gain and are designed to be operated at some minimum amount of closed-loop gain. For these op amps, the gain-bandwidth product is still relatively constant over the region of allowable gain.

Now, consider the following typical example: $I_T = 100\mu\text{A}$, $C_p = 2\text{pF}$. We find that:

$$g_m = \frac{I_T / 2}{V_T} = \frac{50\mu\text{A}}{26\text{mV}} = \frac{1}{520\Omega}$$

$$f_u = \frac{g_m}{2\pi C_p} = \frac{1}{2\pi(520)(2 \cdot 10^{-12})} = 153\text{MHz}.$$

Now, we must consider the large-signal response of the circuit. The slew-rate, SR, is simply the total available charging current, $I_T/2$, divided by the dominant pole capacitance, C_p . For the example under consideration,

$$I = C \frac{dv}{dt}, \quad \frac{dv}{dt} = SR, \quad SR = \frac{I}{C}$$

$$SR = \frac{I_T / 2}{C_p} = \frac{50\mu A}{2pF} = 25V / \mu s.$$

The full-power bandwidth (FPBW) of the op amp can now be calculated from the formula:

$$FPBW = \frac{SR}{2\pi A} = \frac{25V / \mu s}{2\pi \cdot 1V} = 4MHz,$$

where A is the peak amplitude of the output signal. If we assume a 2V peak-to-peak output sine wave (certainly a reasonable assumption for high speed applications), then we obtain a FPBW of only 4MHz, even though the small-signal unity gain-bandwidth product is 153MHz! For a 2V p-p output sine wave, distortion will begin to occur much lower than the actual FPBW frequency. We must increase the SR by a factor of about 40 in order for the FPBW to equal 153MHz. The only way to do this is to increase the tail current, I_T , of the input differential pair by the same factor. This implies a bias current of 4mA in order to achieve a FPBW of 160MHz. We are assuming that C_p is a fixed value of 2pF and cannot be lowered by design.

VFB OP AMP BANDWIDTH AND SLEW RATE CALCULATION

- Assume that $I_T = 100\mu\text{A}$, $C_p = 2\text{pF}$
- $g_m = \frac{I_c}{V_T} = \frac{50\mu\text{A}}{26\text{mV}} = \frac{1}{520\Omega}$
- $f_u = \frac{g_m}{2\pi C_p} = 153\text{MHz}$
- Slew Rate = $SR = \frac{I_T / 2}{C_p} = 25\text{V} / \mu\text{s}$
BUT FOR 2V PEAK-PEAK OUTPUT ($A = 1\text{V}$)
- $FPBW = \frac{SR}{2\pi A} = 4\text{MHz}$
- Must increase I_T to 4mA to get $FPBW = 160\text{MHz}!!$
- Reduce g_m by adding emitter degeneration resistors



1.4

In practice, the FPBW of the op amp should be approximately 5 to 10 times the maximum output frequency in order to achieve acceptable distortion performance (typically 55-80dBc @ 5 to 20MHz, but actual system requirements vary widely).

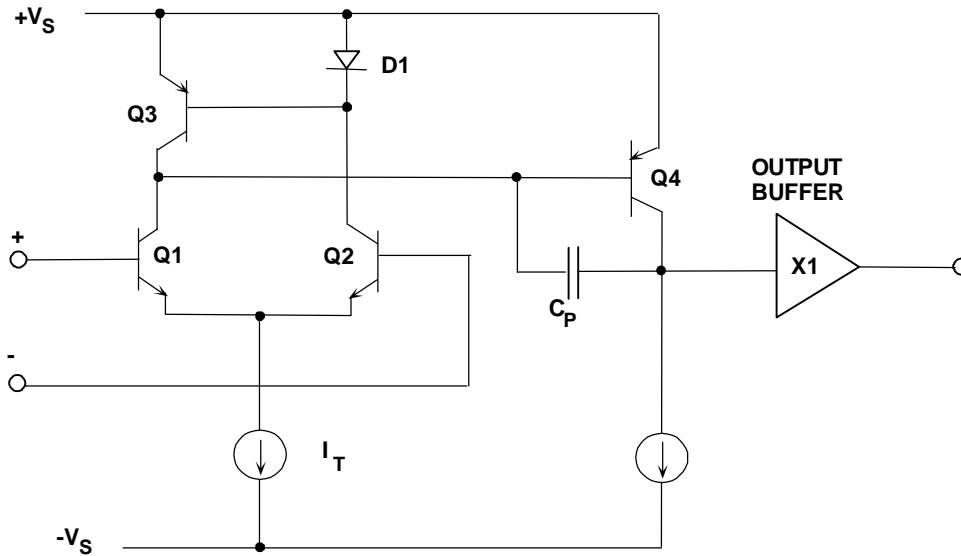
Notice, however, that increasing the tail current causes a proportional increase in g_m and hence f_u . In order to prevent possible instability due to the large increase in f_u , g_m can be reduced by inserting resistors in series with the emitters of Q1 and Q2 (this technique, called *emitter degeneration*, also serves to linearize the g_m transfer function and lower distortion).

A major inefficiency of conventional bipolar voltage feedback op amps is their inability to achieve high slew rates without proportional increases in quiescent current (assuming that C_p is fixed, and has a reasonable minimum value of 2 or 3pF). This of course is not meant to say that high speed op amps designed using this architecture are deficient, it's just that there are circuit design techniques available which allow equivalent performance at lower quiescent currents. This is extremely important in portable battery operated equipment where every milliwatt of power dissipation is critical.

VFB Op Amps Designed on Complementary Bipolar Processes

With the advent of complementary bipolar (CB) processes having high quality PNP transistors as well as NPNs, VFB op amp configurations such as the one shown in the simplified diagram (Figure 1.5) became popular.

VFB OP AMP USING TWO GAIN STAGES

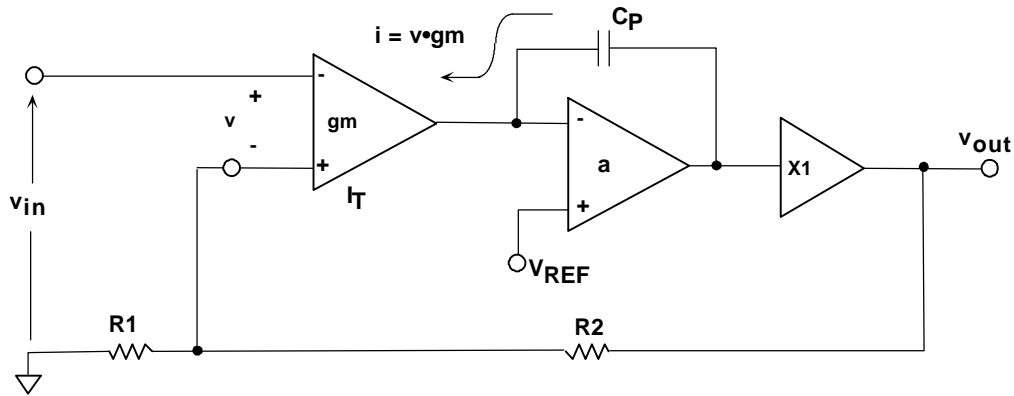


1.5

Notice that the input differential pair (Q1, Q2) is loaded by a current mirror (Q3 and D1). We show D1 as a diode for simplicity, but it is actually a diode-connected PNP transistor (matched to Q3) with the base and collector connected to each other. This simplification will be used in many of the circuit diagrams to follow in this section. The common emitter transistor, Q4, provides a *second* voltage gain stage. Since the PNP transistors are fabricated on a complementary bipolar process, they are high quality and matched to the NPNs and suitable for voltage gain. The dominant pole of the amplifier is set by C_p, and the combination of the gain stage, Q4, and C_p is often referred to as a *Miller Integrator*. The unity-gain output buffer is usually a complementary emitter follower.

The model for this two-stage VFB op amp is shown in Figure 1.6. Notice that the unity gain-bandwidth frequency, f_{u} , is still determined by the g_m of the input stage and the dominant pole capacitance, C_p. The second gain stage increases the DC open-loop gain, but the maximum slew rate is still limited by the input stage tail current: $SR = I_T/C_p$.

MODEL FOR TWO STAGE VFB OP AMP



$$\blacksquare f_u = \frac{gm}{2\pi C_p}$$

$$\blacksquare f_{CL} = \frac{f_u}{1 + \frac{R_2}{R_1}}$$

$$\blacksquare SR = \frac{I_T}{C_p}$$

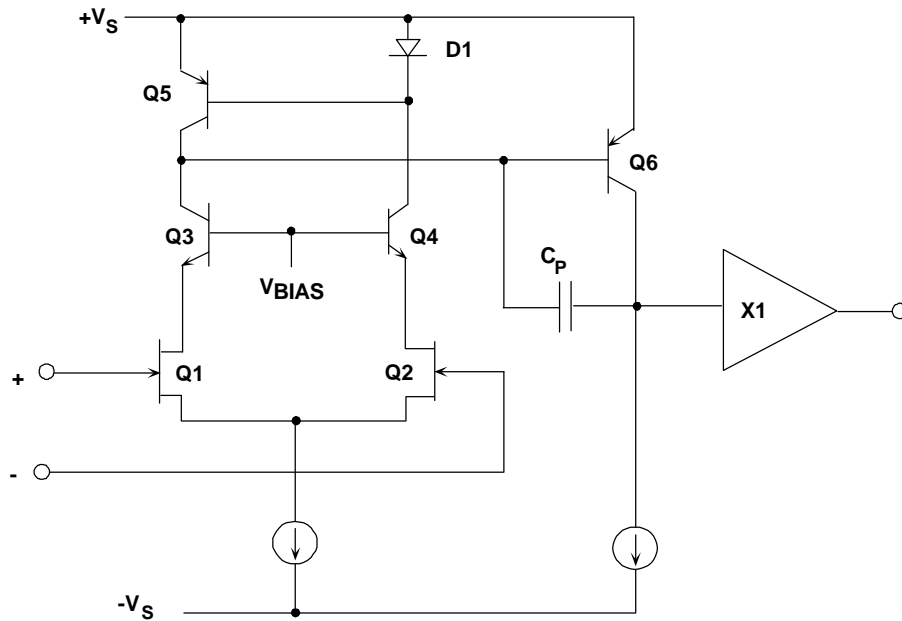


1.6

The two-stage topology is widely used throughout the IC industry in VFB op amps, both precision and high speed.

Another popular VFB op amp architecture is the *folded cascode* as shown in Figure 1.7. An industry-standard video amplifier family (the AD847) is based on this architecture. This circuit takes advantage of the fast PNPs available on a CB process. The differential signal currents in the collectors of Q1 and Q2 are fed to the emitters of a PNP cascode transistor pair (hence the term *folded cascode*). The collectors of Q3 and Q4 are loaded with the current mirror, D1 and Q5, and Q4 provides voltage gain. This single-stage architecture uses the junction capacitance at the high-impedance node for compensation (and some variations of the design bring this node to an external pin so that additional external capacitance can be added).

AD845 BiFET 16MHz OP AMP SIMPLIFIED CIRCUIT

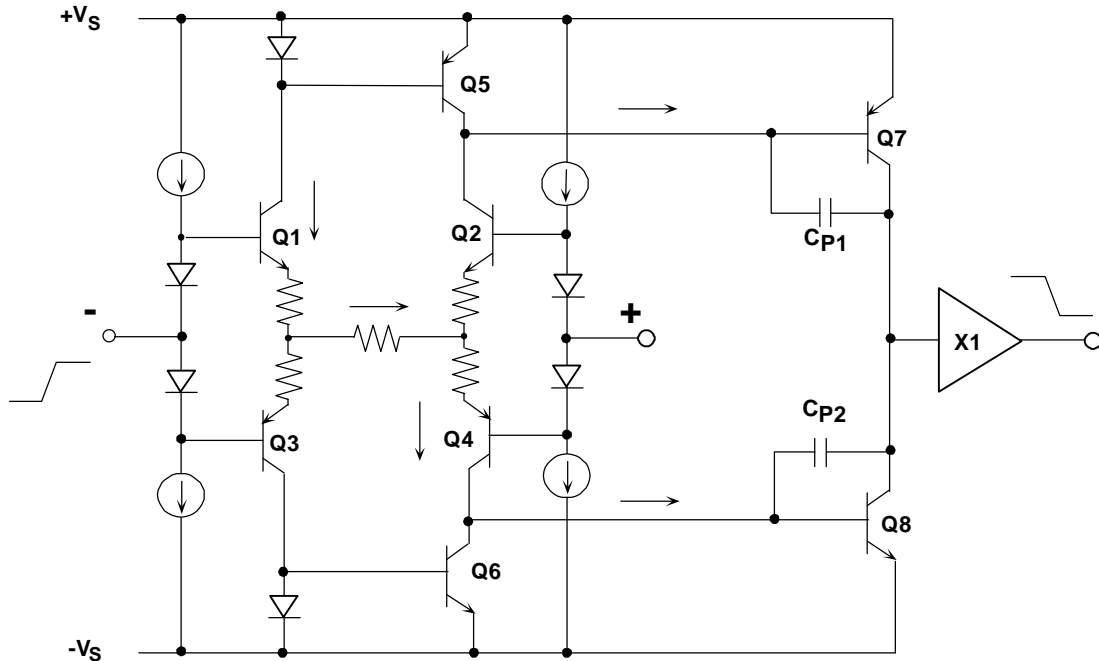


1.8

A New VFB Op Amp Architecture for "Current-on-Demand" Performance, Lower Power, and Improved Slew Rate

Until now, op amp designers had to make the above tradeoffs between the input g_m stage quiescent current and the slew-rate and distortion performance. Analog Devices' has patented a new circuit core which supplies *current-on-demand* to charge and discharge the dominant pole capacitor, C_p , while allowing the quiescent current to be small. The additional current is proportional to the fast slewing input signal and adds to the quiescent current. A simplified diagram of the basic core cell is shown in Figure 1.9.

"QUAD-CORE" VFB gm STAGE FOR CURRENT-ON-DEMAND



1.9

The *quad-core* (g_m stage) consists of transistors Q1, Q2, Q3, and Q4 with their emitters connected together as shown. Consider a positive step voltage on the inverting input. This voltage produces a proportional current in Q1 which is mirrored into Cp1 by Q5. The current through Q1 also flows through Q4 and Cp2. At the dynamic range limit, Q2 and Q3 are correspondingly turned off. Notice that the charging and discharging current for Cp1 and Cp2 is not limited by the quad core bias current. In practice, however, small current-limiting resistors are required forming an "H" resistor network as shown. Q7 and Q8 form the second gain stage (driven differentially from the collectors of Q5 and Q6), and the output is buffered by a unity-gain complementary emitter follower.

The quad core configuration is patented (Roy Gosser, U.S. Patent 5,150,074 and others pending), as well as the circuits which establish the quiescent bias currents (not shown in the diagram). A number of new VFB op amps using this proprietary configuration have been released and have unsurpassed high frequency low distortion performance, bandwidth, and slew rate at the indicated quiescent current levels (see Figure 1.10). The AD9631, AD8036, and AD8047 are optimized for a gain of +1, while the AD9632, AD8037, and AD8048 for a gain of +2. The same quad-core architecture is used as the *second* stage of the AD8041 rail-to-rail output, zero-volt input single-supply op amp. The input stage is a differential PNP pair which allows the input common-mode signal to go about 200mV below the negative supply rail. The AD8042 and AD8044 are dual and quad versions of the AD8041.

"QUAD-CORE" TWO STAGE XFCB VFB OP AMPS AC CHARACTERISTICS VERSUS SUPPLY CURRENT

PART #	I_{SY} / AMP	BANDWIDTH	SLEW RATE	DISTORTION
AD9631/32	17mA	320MHz	1300V/ μ s	-72dBc@20MHz
AD8036/37 Clamped	20mA	240MHz	1200V/ μ s	-72dBc@20MHz
AD8047/48	5.8mA	250MHz	750V/ μ s	-66dBc@5MHz
AD8041 (1)	5.2mA	160MHz	160V/ μ s	-69dBc@10MHz
AD8042 (2)	5.2mA	160MHz	200V/ μ s	-64dBc@10MHz
AD8044 (4)	2.75mA	150MHz	170V/ μ s	-75dBc@5MHz
AD8031 (1)	0.75mA	80MHz	30V/ μ s	-62dBc@1MHz
AD8032 (2)	0.75mA	80MHz	30V/ μ s	-62dBc@1MHz

Number in () indicates single, dual, or quad



1.10

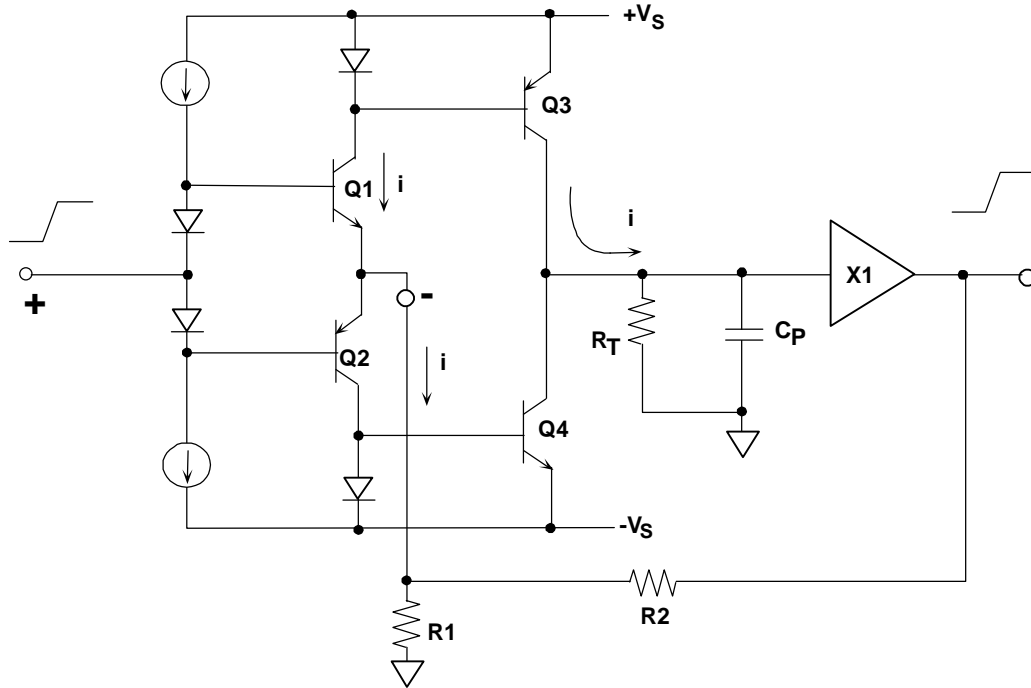
CURRENT FEEDBACK (CFB) OP AMPS

We will now examine the current feedback (CFB) op amp topology which has recently become popular in high speed op amps. The circuit concepts were introduced many years ago, however modern high speed complementary bipolar processes are required to take full advantage of the architecture.

It has long been known that in bipolar transistor circuits, currents can be switched faster than voltages, other things being equal. This forms the basis of non-saturating emitter-coupled logic (ECL) and devices such as current-output DACs. Maintaining low impedances at the current switching nodes helps to minimize the effects of stray capacitance, one of the largest detriments to high speed operation. The current mirror is a good example of how currents can be switched with a minimum amount of delay.

The current feedback op amp topology is simply an application of these fundamental principles of current steering. A simplified CFB op amp is shown in Figure 1.11. The non-inverting input is high impedance and is buffered directly to the inverting input through the complementary emitter follower buffers Q1 and Q2. Note that the inverting input impedance is very low (typically 10 to 100 Ω), because of the low emitter resistance. In the ideal case, it would be zero. This is a fundamental difference between a CFB and a VFB op amp, and also a feature which gives the CFB op amp some unique advantages.

SIMPLIFIED CURRENT FEEDBACK (CFB) OP AMP



1.11

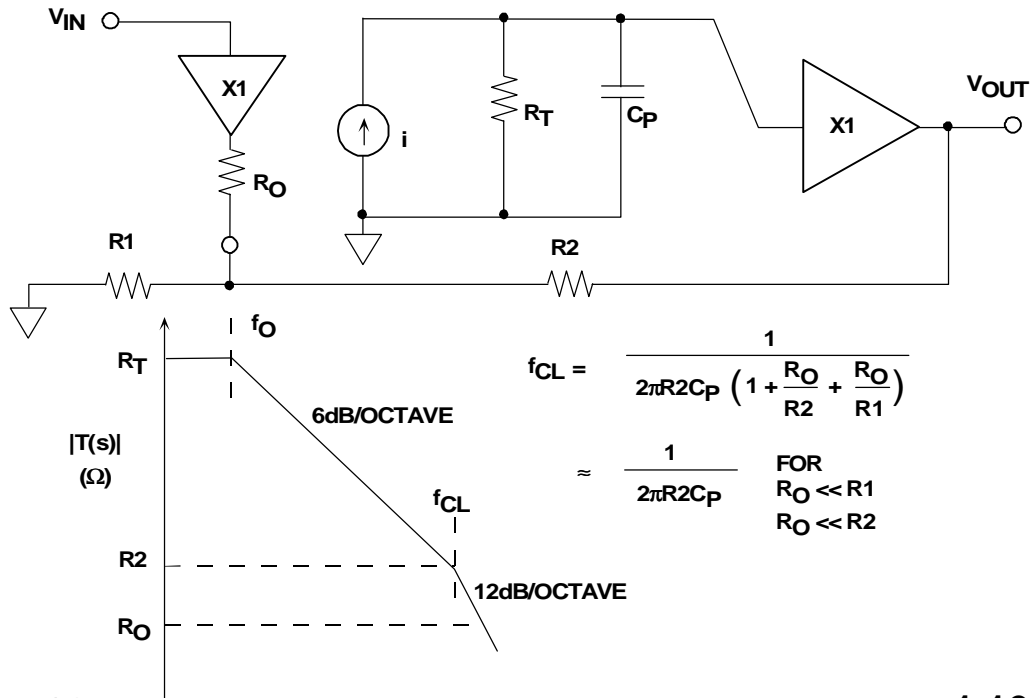
The collectors of Q1 and Q2 drive current mirrors which mirror the inverting input current to the high impedance node, modeled by R_T and C_p . The high impedance node is buffered by a complementary unity gain emitter follower. Feedback from the output to the inverting input acts to force the inverting input *current* to zero, hence the term *Current Feedback*. (In the ideal case, for zero inverting input impedance, no small signal voltage can exist at this node, only small signal current).

Consider a positive step voltage applied to the non-inverting input of the CFB op amp. Q1 immediately sources a proportional current into the external feedback resistors creating an *error current* which is mirrored to the high impedance node by Q3. The voltage developed at the high impedance node is equal to this current multiplied by the equivalent impedance. This is where the term *transimpedance op amp* originated, since the transfer function is an impedance, rather than a unitless voltage ratio as in a traditional VFB op amp.

Note that the error current is not limited by the input stage bias current, i.e., *there is no slew-rate limitation in an ideal CFB op amp*. The current mirrors supply *current-on-demand* from the power supplies. The negative feedback loop then forces the output voltage to a value which reduces the inverting input error current to zero.

The model for a CFB op amp is shown in Figure 1.12 along with the corresponding Bode plot. The Bode plot is plotted on a log-log scale, and the open-loop gain is expressed as a transimpedance, $T(s)$, with units of ohms.

CFB OP AMP MODEL AND BODE PLOT



1.12

The finite output impedance of the input buffer is modeled by R_o . The input error current is i . By applying the principles of negative feedback, we can derive the expression for the op amp transfer function:

$$\frac{v_{out}}{v_{in}} = \frac{1 + \frac{R_2}{R_1}}{1 + j\omega C_p R_2 \left(1 + \frac{R_o}{R_2} + \frac{R_o}{R_1}\right)}$$

At the op amp 3db closed-loop bandwidth frequency, f_{cl} , the following is true:

$$2\pi f_{cl} C_p R_2 \left(1 + \frac{R_o}{R_2} + \frac{R_o}{R_1}\right) = 1$$

Solving for f_{cl} :

$$f_{cl} = \frac{1}{2\pi C_p R_2 \left(1 + \frac{R_o}{R_2} + \frac{R_o}{R_1}\right)}$$

For the condition $R_o \ll R_2$ and R_1 , the equation simply reduces to:

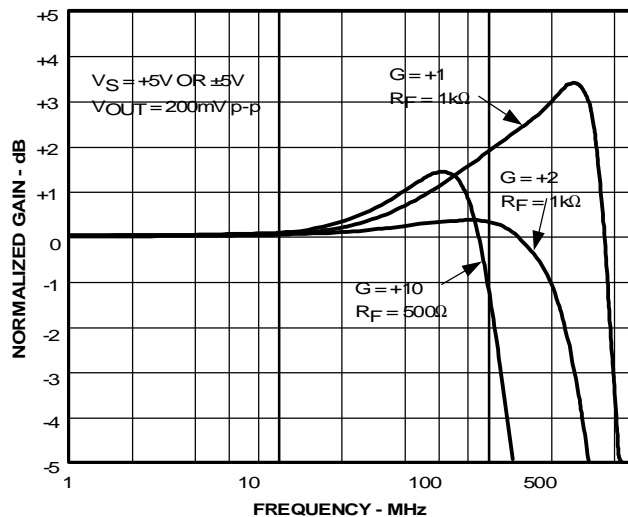
$$f_{cl} = \frac{1}{2\pi C_p R_2}$$

Examination of this equation quickly reveals that *the closed-loop bandwidth of a CFB op amp is determined by the internal dominant pole capacitor, C_p , and the external feedback resistor R_2 , and is independent of the gain-setting resistor, R_1* . This ability to maintain constant bandwidth independent of gain makes CFB op amps ideally suited for wideband programmable gain amplifiers.

Because the closed-loop bandwidth is inversely proportional to the external feedback resistor, R_2 , a CFB op amp is usually optimized for a specific R_2 . Increasing R_2 from it's optimum value lowers the bandwidth, and decreasing it may lead to oscillation and instability because of high frequency parasitic poles.

The frequency response of the AD8011 CFB op amp is shown in Figure 1.13 for various closed-loop values of gain (+1, +2, and +10). Note that even at a gain of +10, the closed loop bandwidth is still greater than 100MHz. The peaking which occurs at a gain of +1 is typical of wideband CFB op amps when used in the non-inverting mode and is due primarily to stray capacitance at the inverting input. The peaking can be reduced by sacrificing bandwidth and using a slightly larger feedback resistor. The AD8011 CFB op amp represents state-of-the-art performance, and key specifications are shown in Figure 1.14.

AD8011 FREQUENCY RESPONSE G = +1, +2, +10



1.13

AD8011 CFB OP AMP KEY SPECIFICATIONS

- 1mA Power Supply Current (+5V or ±5V)
- 300MHz Bandwidth (G = +1)
- 2000 V/μs Slew Rate
- 29ns Settling Time to 0.1%
- Video Specifications (G = +2)
Differential Gain Error 0.02%

Differential Phase Error 0.06°
25MHz 0.1dB Bandwidth

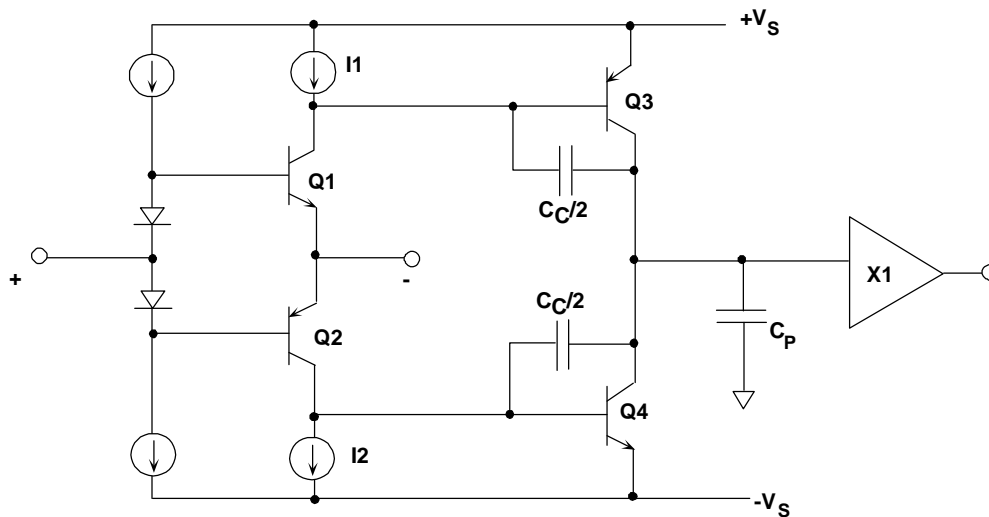
- Distortion
 - 70dBc @ 5MHz
 - 62dBc @ 20MHz
- Fully Specified for ±5V or +5V Operation



1.14

Traditional current feedback op amps have been limited to a single gain stage, using current-mirrors as previously described. The AD8011 (and also others in this family: AD8001, AD8002, AD8004, AD8005, AD8009, AD8013, AD8072, AD8073), unlike traditional CFB op amps uses a two-stage gain configuration as shown in Figure 1.15. Until now, fully complementary two-gain stage CFB op amps have been impractical because of their high power dissipation. The AD8011 employs a second gain stage consisting of a pair of complementary amplifiers (Q3 and Q4). Note that they are not connected as current mirrors but as grounded-emitters. The detailed design of current sources (I1 and I2), and their respective bias circuits (Roy Gosser, patent-applied-for) are the key to the success of the two-stage CFB circuit; they keep the amplifier's quiescent power low, yet are capable of supplying *current-on-demand* for wide current excursions required during fast slewing.

SIMPLIFIED TWO-STAGE CFB OP AMP



1.15

A further advantage of the two-stage amplifier is the higher overall bandwidth (for the same power), which means lower signal distortion and the ability to drive heavier external loads.

Thus far, we have learned several key features of CFB op amps. The most important is that for a given complementary bipolar IC process, *CFB generally always yields higher FPBW (hence lower distortion) than VFB for the same amount of quiescent supply current*. This is because there is practically no slew-rate limiting in CFB. Because of this, the full power bandwidth and the small signal bandwidth are approximately the same.

The second important feature is that the *inverting input impedance of a CFB op amp is very low*. This can be advantageous when using the op amp in the inverting mode as an I/V converter, because there is much less sensitivity to inverting input capacitance than with VFB.

The third feature is that *the closed-loop bandwidth of a CFB op amp is determined by the value of the internal C_p capacitor and the external feedback resistor R_2 and is relatively independent of the gain-setting resistor R_1* . We will now examine some typical applications issues and make further comparisons between CFBs and VFBs.

CURRENT FEEDBACK OP AMP FAMILY

PART	I_{SY}/AMP	BANDWIDTH	SLEW RATE	DISTORTION
AD8001 (1)	5.5mA	880MHz	1200V/ μ s	-65dBc@5MHz
AD8002 (2)	5.0mA	600MHz	1200 V/ μ s	-65dBc@5MHz
AD8004 (4)	3.5mA	250MHz	3000 V/ μ s	-78dBc@5MHz
AD8005 (1)	0.4mA	180MHz	500 V/ μ s	-53dBc@5MHz
AD8009 (1)	11mA	1000MHz	7000 V/ μ s	-80dBc@5MHz
AD8011 (1)	1mA	300MHz	2000 V/ μ s	-70dBc@5MHz
AD8012 (2)	1mA	300MHz	1200 V/ μ s	-66dBc@5MHz
AD8013 (3)	4mA	140MHz	1000 V/ μ s	$\Delta G=0.02\%$, $\Delta\phi=0.06^\circ$
AD8072 (2)	5mA	100MHz	500 V/ μ s	$\Delta G=0.05\%$, $\Delta\phi=0.1^\circ$
AD8073 (3)	5mA	100MHz	500 V/ μ s	$\Delta G=0.05\%$, $\Delta\phi=0.1^\circ$

Number in () Indicates Single, Dual, Triple, or Quad



1.16

SUMMARY: CURRENT FEEDBACK OP AMPS

- CFB yields higher FPBW and lower distortion than VFB for the same process and power dissipation

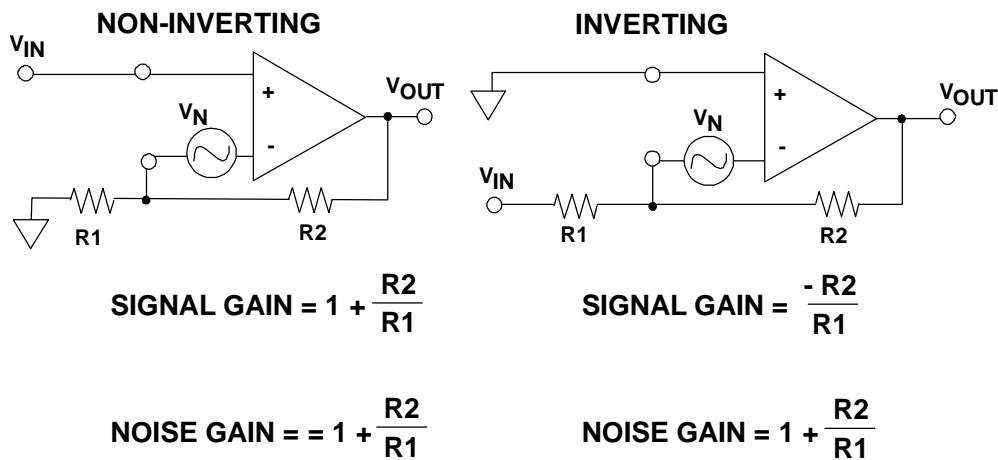
- Inverting input impedance of a CFB op amp is low, non-inverting input impedance is high
- Closed-loop bandwidth of a CFB op amp is determined by the internal dominant-pole capacitance and the external feedback resistor, independent of the gain-setting resistor

EFFECTS OF FEEDBACK CAPACITANCE IN OP AMPS

At this point, the term *noise gain* needs some clarification. Noise gain is the amount by which a small amplitude noise voltage source in series with an input terminal of an op amp is amplified when measured at the output. The input voltage noise of an op amp is modeled in this way. It should be noted that the DC noise gain can also be used to reflect the input offset voltage (and other op amp input error sources) to the output.

Noise gain must be distinguished from *signal gain*. Figure 1.18 shows an op amp in the inverting and non-inverting mode. In the non-inverting mode, notice that noise gain is equal to signal gain. However, in the inverting mode, the noise gain doesn't change, but the signal gain is now $-R2/R1$. Resistors are shown as feedback elements, however, the networks may also be reactive.

NOISE GAIN AND SIGNAL GAIN COMPARISON



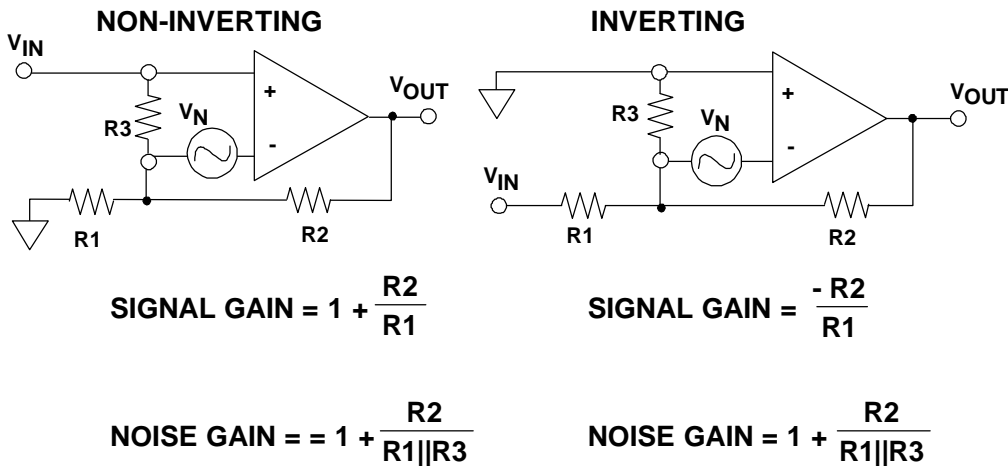
FOR VFB OP AMP:

$$\text{CLOSED-LOOP BW} = \frac{\text{UNITY GAIN BANDWIDTH FREQUENCY}}{\text{NOISE GAIN}}$$

$$f_{CL} = \frac{f_u}{G}$$

Two other configurations are shown in Figure 1.19 where the noise gain has been increased independent of signal gain by the addition of R3 across the input terminals of the op amp. This technique can be used to stabilize de-compensated op amps which are unstable for low values of noise gain. However, the sensitivity to input noise and offset voltage is correspondingly increased.

INCREASING THE NOISE GAIN WITHOUT AFFECTING SIGNAL GAIN



1.19

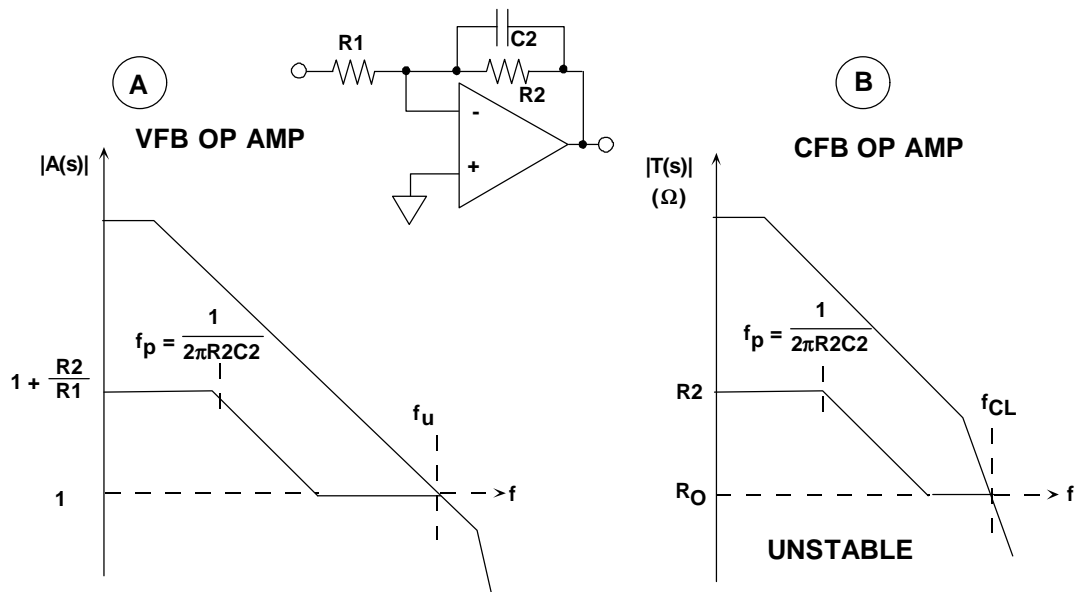
Noise gain is often plotted as a function of frequency on a Bode plot to determine the op amp stability. If the feedback is purely resistive, the noise gain is constant with frequency. However, reactive elements in the feedback loop will cause it to change with frequency. Using a log-log scale for the Bode plot allows the noise gain to be easily drawn by simply calculating the breakpoints determined by the frequencies of the various poles and zeros. The point of intersection of the noise gain with the open-loop gain not only determines the op amp *closed-loop bandwidth*, but also can be used to analyze stability.

An excellent explanation of how to make simplifying approximations using Bode plots to analyze gain and phase performance of a feedback networks is given in Reference 4.

Just as signal gain and noise gain can be different, so can the *signal bandwidth* and the *closed-loop bandwidth*. The op amp closed-loop bandwidth, f_{CL} , is always determined by the intersection of the noise gain with the open-loop frequency response. The signal bandwidth is equal to the closed-loop bandwidth only if the feedback network is purely resistive.

It is quite common to use a capacitor in the feedback loop of a VFB op amp to shape the frequency response as in a simple single-pole lowpass filter (see Figure 1.20a). The resulting noise gain is plotted on a Bode plot to analyze stability and phase margin. Stability of the system is determined by the net slope of the noise gain and the open loop gain where they intersect. For unconditional stability, the noise gain plot must intersect the open loop response with a net slope of less than 12dB/octave. In this case, the net slope where they intersect is 6dB/octave, indicating a stable condition. Notice for the case drawn in Figure 1.20a, the second pole in the frequency response occurs at a considerably higher frequency than f_u .

NOISE GAIN STABILITY ANALYSIS FOR VFB AND CFB OP AMPS WITH FEEDBACK CAPACITOR



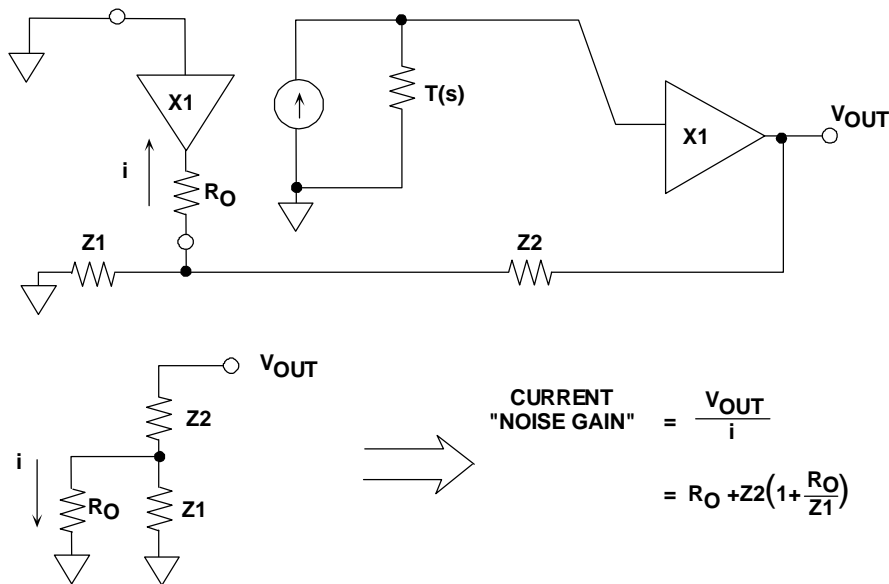
1.20

In the case of the CFB op amp (Figure 1.20b), the same analysis is used, except that the open-loop transimpedance gain, $T(s)$, is used to construct the Bode plot. The definition of *noise gain* (for the purposes of stability analysis) for a CFB op amp, however, must be redefined in terms of a *current* noise source attached to the inverting input (see Figure 1.21). This current is reflected to the output by an impedance which we define to be the "current noise gain" of a CFB op amp:

"CURRENT NOISE GAIN"

$$\equiv R_0 + Z_2 \left(1 + \frac{R_0}{Z_1} \right)$$

CURRENT "NOISE GAIN" DEFINITION FOR CFB OP AMP FOR USE IN STABILITY ANALYSIS



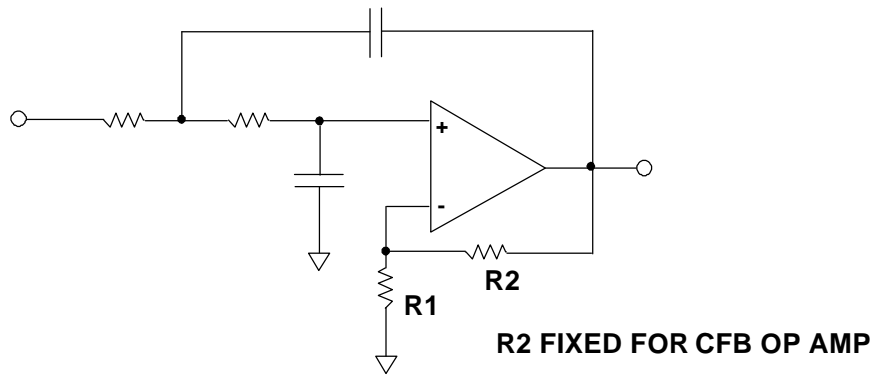
1.21

Now, return to Figure 1.20b, and observe the CFB *current noise gain* plot. At low frequencies, the CFB current noise gain is simply R_2 (making the assumption that R_0 is much less than Z_1 or Z_2). The first pole is determined by R_2 and C_2 . As the frequency continues to increase, C_2 becomes a short circuit, and all the inverting input current flows through R_0 (refer back to Figure 1.21).

The CFB op amp is normally optimized for best performance for a fixed feedback resistor, R_2 . Additional poles in the transimpedance gain, $T(s)$, occur at frequencies above the closed loop bandwidth, f_{CL} , (set by R_2). Note that the intersection of the CFB current noise gain with the open-loop $T(s)$ occurs where the slope of the $T(s)$ function is 12dB/octave. This indicates instability and possible oscillation.

It is for this reason that *CFB op amps are not suitable in configurations which require capacitance in the feedback loop*, such as simple active integrators or lowpass filters. They can, however, be used in certain active filters such as the Sallen-Key configuration shown in Figure 1.22 which do not require capacitance in the feedback network.

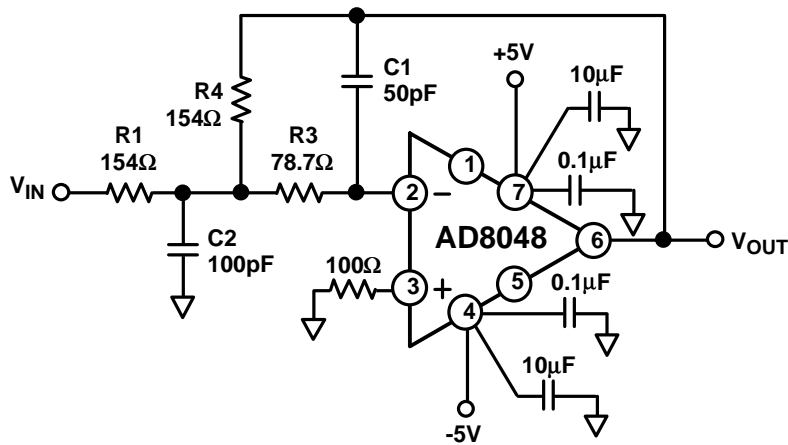
EITHER CFB OR VFB OP AMPS CAN BE USED IN THE SALLEN-KEY FILTER CONFIGURATION



1.22

VFB op amps, on the other hand, make very flexible active filters. A multiple feedback 20MHz lowpass filter using the AD8048 is shown in Figure 1.23.

MULTIPLE FEEDBACK 20MHz LOWPASS FILTER USING THE AD8048 VFB OP AMP



1.23

In general, the amplifier should have a bandwidth which is at least ten times the bandwidth of the filter if problems due to phase shift of the amplifier are to be avoided. (The AD8048 has a bandwidth of over 200MHz in this configuration). The filter is designed as follows:

Choose:

$$F_0 = \text{Cutoff Frequency} = 20\text{MHz}$$

$$\alpha = \text{Damping Ratio} = 1/Q = 2$$

$$H = \text{Absolute Value of Circuit Gain}$$

$$= |-R_4/R_1| = 1$$

$$k = 2\pi F_0 C_1$$

$$C_2 = \frac{4C_1(H+1)}{\alpha^2} = 100\text{pF}, \text{ for } C_1 = 50\text{pF}$$

$$R_1 = \frac{\alpha}{2Hk} = 159.2\Omega, \text{ use } 154\Omega$$

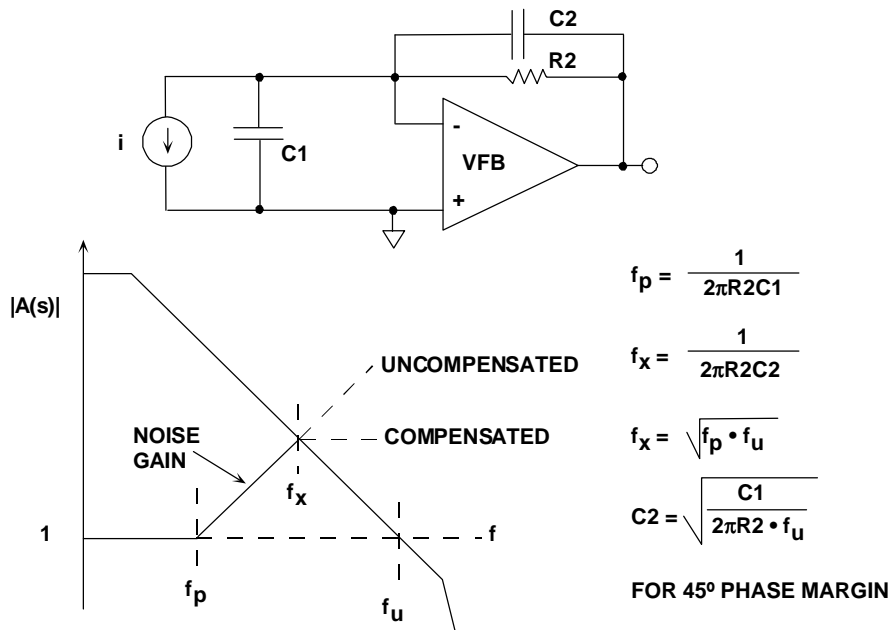
$$R_3 = \frac{\alpha}{2k(H+1)} = 79.6\Omega, \text{ use } 78.7\Omega$$

$$R_4 = H \cdot R_1 = 159.2\Omega, \text{ use } 154\Omega$$

HIGH SPEED CURRENT-TO-VOLTAGE CONVERTERS, AND THE EFFECTS OF INVERTING INPUT CAPACITANCE

Fast op amps are useful as current-to-voltage converters in such applications as high speed photodiode preamplifiers and current-output DAC buffers. A typical application using a VFB op amp as an I/V converter is shown in Figure 1.24.

COMPENSATING FOR INPUT CAPACITANCE IN A CURRENT-TO-VOLTAGE CONVERTER USING VFB OP AMP



1.24

The net input capacitance, C_1 , forms a pole at a frequency f_p in the noise gain transfer function as shown in the Bode plot, and is given by:

$$f_p = \frac{1}{2\pi R_2 C_1}.$$

If left uncompensated, the phase shift at the frequency of intersection, f_x , will cause instability and oscillation. Introducing a zero at f_x by adding feedback capacitor C_2 stabilizes the circuit and yields a phase margin of about 45 degrees. The location of the zero is given by:

$$f_x = \frac{1}{2\pi R_2 C_2}.$$

Although the addition of C_2 actually decreases the pole frequency slightly, this effect is negligible if $C_2 \ll C_1$. The frequency f_x is the geometric mean of f_p and the unity-gain bandwidth frequency of the op amp, f_u ,

$$f_x = \sqrt{f_p \cdot f_u}.$$

These equations can be solved for C_2 :

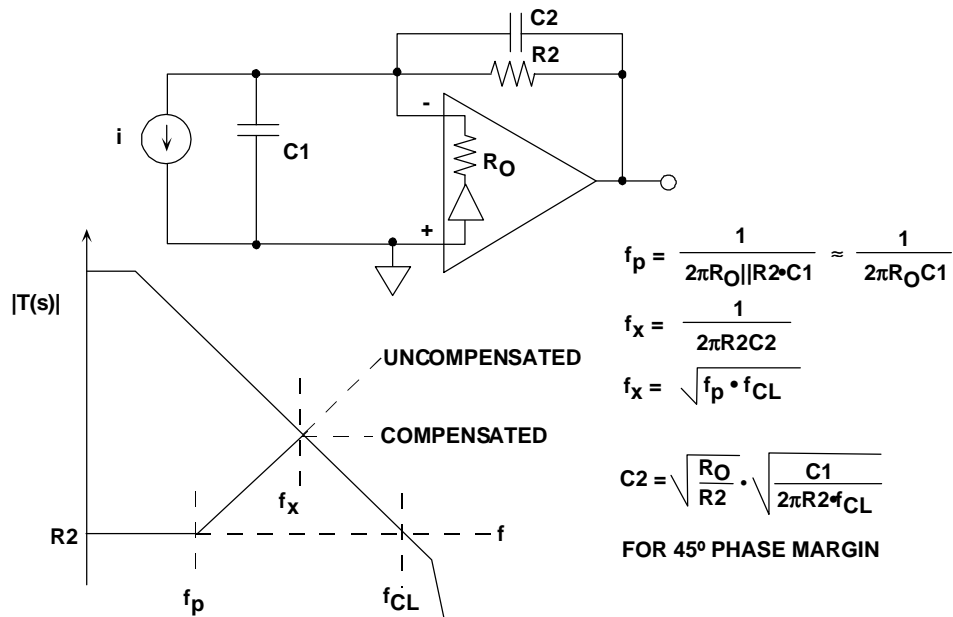
$$C_2 = \sqrt{\frac{C_1}{2\pi R_2 \cdot f_u}}.$$

This value of C2 will yield a phase margin of about 45 degrees. Increasing the capacitor by a factor of 2 increases the phase margin to about 65 degrees (see References 4 and 5).

In practice, the optimum value of C2 may be optimized experimentally by varying it slightly to optimize the output pulse response.

A similar analysis can be applied to a CFB op amp as shown in Figure 1.25. In this case, however, the low inverting input impedance, Ro, greatly reduces the sensitivity to input capacitance. In fact, an ideal CFB with zero input impedance would be totally insensitive to any amount of input capacitance!

COMPENSATING FOR INPUT CAPACITANCE IN A CURRENT-TO-VOLTAGE CONVERTER USING CFB OP AMP



1.25

The pole caused by C1 occurs at a frequency f_p :

$$f_p = \frac{1}{2\pi(R_O || R_2)C_1} \approx \frac{1}{2\pi R_O C_1}$$

This pole frequency will be generally be much higher than the case for a VFB op amp, and the pole can be ignored completely if it occurs at a frequency greater than the closed-loop bandwidth of the op amp.

We next introduce a compensating zero at the frequency f_x by inserting the capacitor C2:

$$f_x = \frac{1}{2\pi R_2 C_2}$$

As in the case for VFB, f_x is the geometric mean of f_p and f_{cl} :

$$f_x = \sqrt{f_p \cdot f_u} .$$

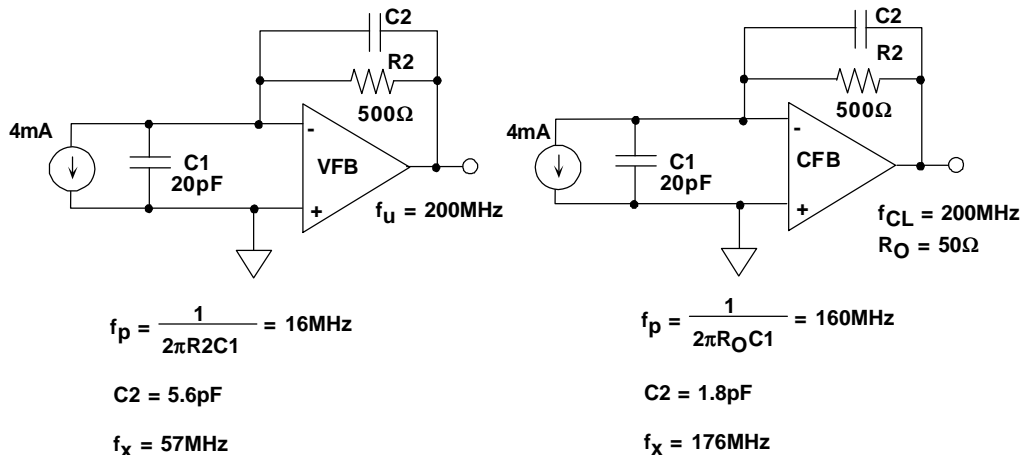
Solving the equations for C2 and rearranging it yields:

$$C2 = \sqrt{\frac{R_o}{R2}} \cdot \sqrt{\frac{C1}{2\pi R2 \cdot f_{cl}}} .$$

There is a significant advantage in using a CFB op amp in this configuration as can be seen by comparing the similar equation for C2 required for a VFB op amp. If the unity-gain bandwidth product of the VFB is equal to the closed-loop bandwidth of the CFB (at the optimum R2), then the size of the CFB compensation capacitor, C2, is reduced by a factor of $\sqrt{R2 / R_o}$.

A comparison in an actual application is shown in Figure 1.26. The full scale output current of the DAC is 4mA, the net capacitance at the inverting input of the op amp is 20pF, and the feedback resistor is 500Ω. In the case of the VFB op amp, the pole due to C1 occurs at 16MHz. A compensating capacitor of 5.6pF is required for 45 degrees of phase margin, and the signal bandwidth is 57MHz.

LOW INVERTING INPUT IMPEDANCE OF CFB OP AMP MAKES IT RELATIVELY INSENSITIVE TO INPUT CAPACITANCE WHEN USED AS A CURRENT-TO-VOLTAGE CONVERTER



For the CFB op amp, however, because of the low inverting input impedance ($R_o = 50\Omega$), the pole occurs at 160MHz, the required compensation capacitor is about 1.8pF, and the corresponding signal bandwidth is 176MHz. In actual practice, the

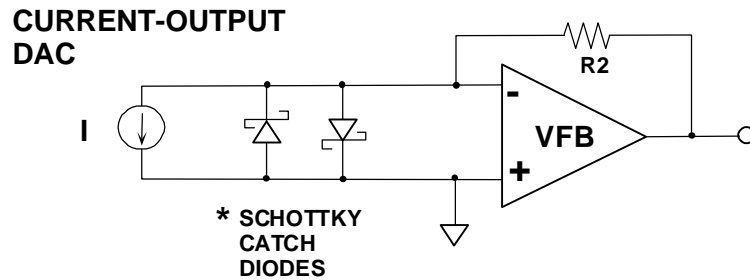
pole frequency is so close to the closed-loop bandwidth of the op amp that it could probably be left uncompensated.

It should be noted that a CFB op amp's relative insensitivity to inverting input capacitance is when it is used in the inverting mode. In the non-inverting mode, even a few picofarads of stray capacitance on the inverting input can cause significant gain-peaking and potential instability.

Another advantage of the low inverting input impedance of the CFB op amp is when it is used as an I/V converter to buffer the output of a high speed current output DAC. When a step function current (or DAC switching glitch) is applied to the inverting input of a VFB op amp, it can produce a large voltage transient until the signal can propagate through the op amp to its output and negative feedback is regained. Back-to-back Schottky diodes are often used to limit this voltage swing as shown in Figure 1.27. These diodes must be low capacitance, small geometry devices because their capacitance adds to the total input capacitance.

A CFB op amp, on the other hand, presents a low impedance (R_o) to fast switching currents even before the feedback loop is closed, thereby limiting the voltage excursion without the requirement of the external diodes. This greatly improves the settling time of the I/V converter.

LOW INVERTING INPUT IMPEDANCE OF CFB OP AMP HELPS REDUCE AMPLITUDE OF FAST DAC TRANSIENTS



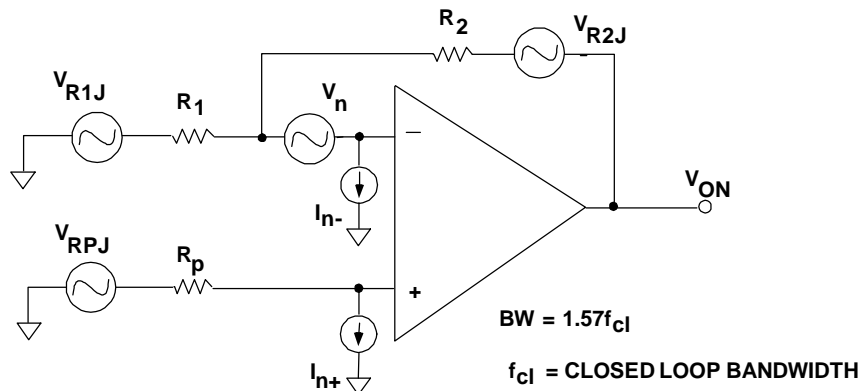
* NOT REQUIRED FOR CFB OP AMP
BECAUSE OF LOW INVERTING INPUT IMPEDANCE

Op amp noise has two components: low frequency noise whose spectral density is inversely proportional to the square root of the frequency and white noise at medium and high frequencies. The low-frequency noise is known as 1/f noise (the noise *power* obeys a 1/f law - the noise voltage or noise current is proportional to 1/√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 the 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 high-speed op amps.

In most applications of high speed op amps, it is the total output rms noise that is generally of interest. Because of the high bandwidths, the chief contributor to the output rms noise is the white noise, and that of the 1/f noise is negligible.

In order to better understand the effects of noise in high speed op amps, we use the classical noise model shown in Figure 1.28. This diagram identifies all possible white noise sources, including the external noise in the source and the feedback resistors. The equation allows you to calculate the total output rms noise over the closed-loop bandwidth of the amplifier. This formula works quite well when the frequency response of the op amp is relatively flat. If there is more than a few dB of high frequency peaking, however, the actual noise will be greater than the predicted because the contribution over the last octave before the 3db cutoff frequency will dominate. In most applications, the op amp feedback network is designed so that the bandwidth is relatively flat, and the formula provides a good estimate. Note that BW in the equation is the equivalent noise bandwidth which, for a single-pole system, is obtained by multiplying the closed-loop bandwidth by 1.57.

OP AMP NOISE MODEL FOR A FIRST-ORDER CIRCUIT WITH RESISTIVE FEEDBACK



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

Figure 1.29 shows a table which indicates how the individual noise contributors are referred to the output. After calculating the individual noise spectral densities in this table, they can be squared, added, and then the square root of the sum of the squares yields the RSS value of the output noise spectral density since all the sources are uncorrelated. This value is multiplied by the square root of the noise bandwidth (noise bandwidth = closed-loop bandwidth multiplied by a correction factor of 1.57) to obtain the final value for the output rms noise.

REFERRING ALL NOISE SOURCES TO THE OUTPUT

NOISE SOURCE EXPRESSED AS A VOLTAGE	MULTIPLY BY THIS FACTOR TO REFER TO THE OP AMP OUTPUT
Johnson Noise in R_p : $\sqrt{4kTR_p}$	Noise Gain = $1 + \frac{R_2}{R_1}$
Non-Inverting Input Current Noise Flowing in R_p : $I_{n+}R_p$	Noise Gain = $1 + \frac{R_2}{R_1}$
Input Voltage Noise: V_n	Noise Gain = $1 + \frac{R_2}{R_1}$
Johnson Noise in R_1 : $\sqrt{4kTR_1}$	$-R_2/R_1$ (Gain from input of R_1 to Output)
Johnson Noise in R_2 : $\sqrt{4kTR_2}$	1
Inverting Input Current Noise Flowing in R_2 : $I_{n-}R_2$	1



1.29

Typical high speed op amps with bandwidths greater than 150MHz or so, and bipolar input stages have input voltage noises ranging from about 2 to 20nV/ $\sqrt{\text{Hz}}$. To put voltage noise in perspective, let's look at the Johnson noise spectral density of a resistor:

$$v_n = \sqrt{4kTR \cdot BW},$$

where k is Boltzmann's constant, T is the absolute temperature, R is the resistor value, and BW is the equivalent noise bandwidth of interest. (The equivalent noise bandwidth of a single-pole system is 1.57 times the 3dB frequency). Using the formula, a 100 Ω resistor has a noise density of 1.3nV/ $\sqrt{\text{Hz}}$, and a 1000 Ω resistor about 4nV/ $\sqrt{\text{Hz}}$ (values are at room temperature: 27°C, or 300K).

The base-emitter in a bipolar transistor has an equivalent noise voltage source which is due to the "shot noise" of the collector current flowing in the transistor's

(noiseless) incremental emitter resistance, r_e . The current noise is proportional to the square root of the collector current, I_c . The emitter resistance, on the other hand, is inversely proportional to the collector current, so *the shot-noise voltage is inversely proportional to the square root of the collector current*. (Reference 5, Section 9).

Voltage noise in FET-input op amps tends to be larger than for bipolar ones, but current noise is extremely low (generally only a few tens of $\text{fA}/\sqrt{\text{Hz}}$) because of the low input bias currents. However, FET-inputs are not generally required for op amp applications requiring bandwidths greater than 100MHz.

Op amps also have input current noise on each input. For high-speed FET-input op amps, the gate currents are so low that input current noise is almost always negligible (measured in $\text{fA}/\sqrt{\text{Hz}}$).

For a VFB op amp, the inverting and non-inverting input current noise are typically equal, and almost always uncorrelated. Typical values for wideband VFB op amps range from $0.5\text{pA}/\sqrt{\text{Hz}}$ to $5\text{pA}/\sqrt{\text{Hz}}$. The input current noise of a bipolar input stage is increased when input bias-current cancellation generators are added, because their current noise is not correlated, and therefore adds (in an RSS manner) to the intrinsic current noise of the bipolar stage.

The input voltage noise in CFB op amps tends to be lower than for VFB op amps having the same approximate bandwidth. This is because the input stage in a CFB op amp is usually operated at a higher current, thereby reducing the emitter resistance and hence the voltage noise. Typical values for CFB op amps range from about 1 to $5\text{nV}/\sqrt{\text{Hz}}$.

The input current noise of CFB op amps tends to be larger than for VFB op amps because of the generally higher bias current levels. The inverting and non-inverting current noise of a CFB is usually different because of the unique input architecture, and are specified separately. In most cases, the inverting input current noise is the larger of the two. Typical input current noise for CFB op amps ranges from 5 to $40\text{pA}/\sqrt{\text{Hz}}$.

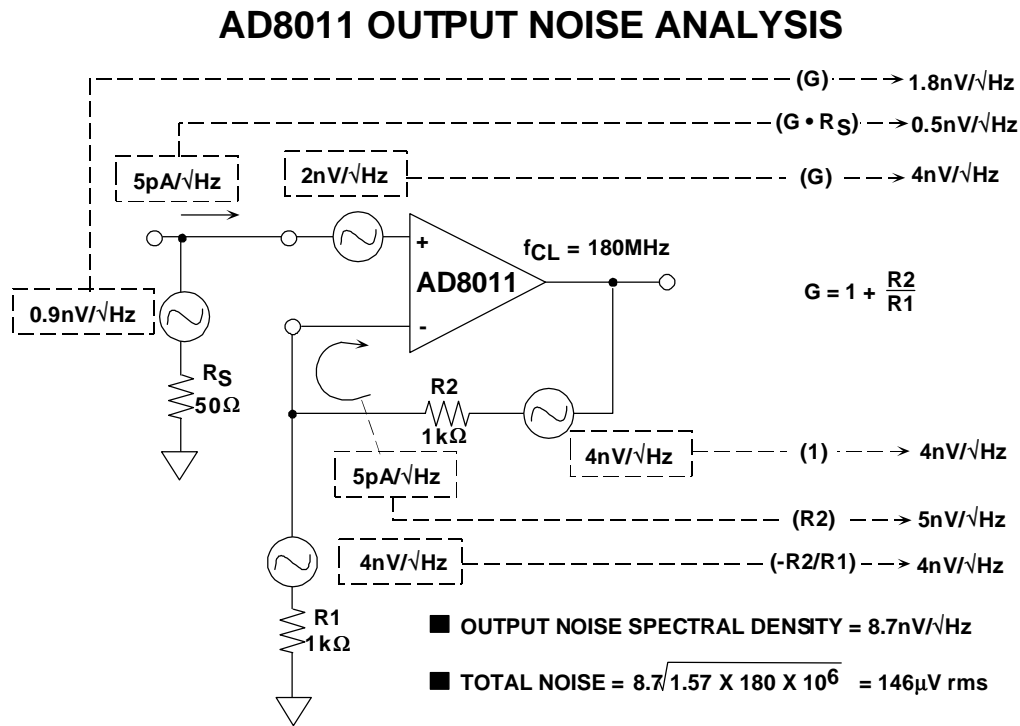
The general principle of noise calculation is that uncorrelated noise sources add in a root-sum-squares manner, which means that if a noise source has a contribution to the output noise of a system which is less than 20% of the amplitude of the noise from other noise source in the system, then its contribution 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. This can simplify the calculations using the formula, assuming the correct decisions are made regarding the sources to be included and those to be neglected.

The sources which dominate the output noise are highly dependent on the closed-loop gain of the op amp. Notice that for high values of closed loop gain, the op amp voltage noise will tend be the chief contributor to the output noise. At low gains, the effects of the input current noise must also be considered, and may dominate, especially in the case of a CFB op amp.

Feedforward/feedback resistors in high speed op amp circuits may range from less than 100Ω to more than $1\text{k}\Omega$, so it is difficult to generalize about their contribution

to the total output noise without knowing the specific values and the closed loop gain. The best way to make the calculations is to write a simple computer program which performs the calculations automatically and include all noise sources. In most high speed applications, the source impedance noise can be neglected for source impedances of 100Ω or less.

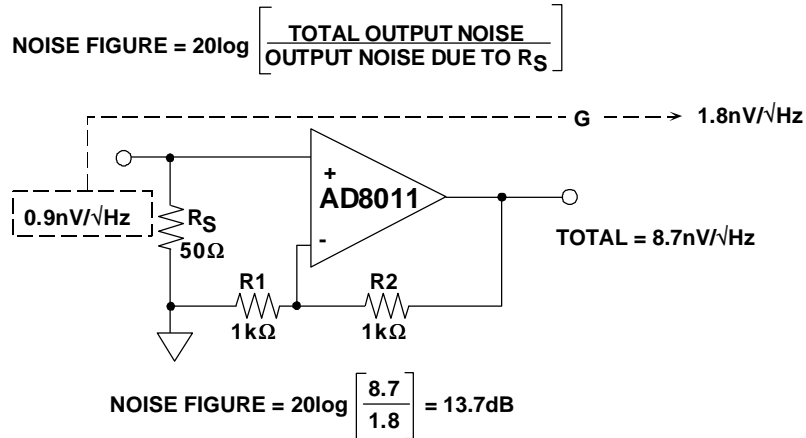
Figure 1.30 shows an example calculation of total output noise for the AD8011 (300MHz, 1mA) CFB op amp. All six possible sources are included in the calculation. The appropriate multiplying factors which reflect the sources to the output are also shown on the diagram. For $G=2$, the close-loop bandwidth of the AD8011 is 180MHz. The correction factor of 1.57 in the final calculation converts this single-pole bandwidth into the circuits equivalent noise bandwidth.



1.30

In communications applications, it is common to specify the *noise figure* (NF) of an amplifier. Figure 1.31 shows the definition. NF is the ratio of the total integrated output noise from all sources to the total output noise which would result if the op amp were "noiseless" (this noise would be that of the source resistance multiplied by the gain of the op amp using the closed-loop bandwidth of the op amp to make the calculation). Noise figure is expressed in dB. The value of the source resistance must be specified, and in most RF systems, it is 50Ω . Noise figure is useful in communications receiver design, since it can be used to measure the decrease in signal-to-noise ratio. For instance, an amplifier with a noise figure of 10dB following a stage with a signal-to-noise ratio of 50dB reduces the signal-to-noise ratio to 40dB.

NOISE FIGURE OF AN OP AMP



1.31

The ratio is commonly expressed in dB and is useful in signal chain analysis. In the previous example, the total output voltage noise was $8.7\text{nV}/\sqrt{\text{Hz}}$. Integrated over the closed loop bandwidth of the op amp (180MHz), this yielded an output noise of $146\mu\text{V rms}$. The noise of the 50Ω source resistance is $0.9\text{nV}/\sqrt{\text{Hz}}$. If the op amp were noiseless (with noiseless feedback resistors), this noise would appear at the output multiplied by the noise gain ($G=2$) of the op amp, or $1.8\text{nV}/\sqrt{\text{Hz}}$. The total output rms noise just due to the source resistor integrated over the same bandwidth is $30.3\mu\text{V rms}$. The noise figure is calculated as:

$$\text{NF} = 20 \log_{10} \left(\frac{146}{30.3} \right) = 13.7\text{dB}.$$

The same result can be obtained by working with spectral densities, since the bandwidths used for the integration are the same and cancel each other in the equation.

$$\text{NF} = 20 \log_{10} \left(\frac{8.7}{1.8} \right) = 13.7\text{dB}.$$

HIGH SPEED OP AMP NOISE SUMMARY

- Voltage Feedback Op Amps:
 - ◆ Voltage Noise: 2 to $20\text{nV}/\sqrt{\text{Hz}}$
 - ◆ Current Noise: 0.5 to $5\text{pA}/\sqrt{\text{Hz}}$
- Current Feedback Op Amps:

- ◆ Voltage Noise: 1 to 5nV/√Hz
- ◆ Current Noise: 5 to 40pA/√Hz

- Noise Contribution from Source Negligible if < 100Ω
- Voltage Noise Usually Dominates at High Gains
- Reflect Noise Sources to Output and Combine (RSS)
- Errors Will Result if there is Significant High Frequency Peaking

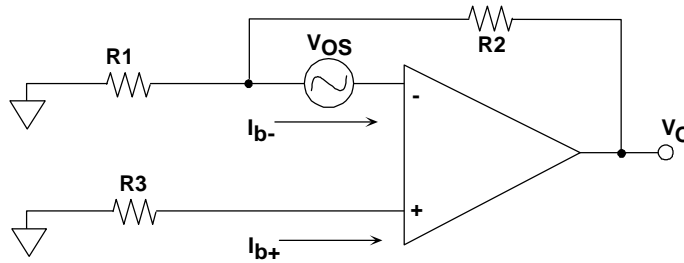


1.32

DC CHARACTERISTICS OF HIGH SPEED OP AMPS

High speed op amps are optimized for bandwidth and settling time, not for precision DC characteristics as found in lower frequency op amps such as the industry standard OP27. In spite of this, however, high speed op amps do have reasonably good DC performance. The model shown in Figure 1.33 shows how to reflect the input offset voltage and the offset currents to the output.

MODEL FOR CALCULATING TOTAL OP AMP OUTPUT VOLTAGE OFFSET



- $V_O = \pm V_{OS} \left[1 + \frac{R_2}{R_1} \right] + I_{b+} R_3 \left[1 + \frac{R_2}{R_1} \right] - I_{b-} R_2$

- IF $I_{b+} = I_{b-}$ AND $R_3 = R_1 || R_2$

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



1.33

Input offset voltages of high speed bipolar input op amps are rarely trimmed, since offset voltage matching of the input stage is excellent, typically ranging from 1 to 3mV, with offset temperature coefficients of 5 to 15 μ V/ $^{\circ}$ C.

Input bias currents on VFB op amps (with no input bias current compensation circuits) are approximately equal for (+) and (-) inputs, and can range from 1 to 5 μ A. The output offset voltage due to the input bias currents can be nulled by making the effective source resistance, R3, equal to the parallel combination of R1 and R2.

This scheme will not work, however, with bias-current compensated VFB op amps which have additional current generators on their inputs. In this case, the net input bias currents are not necessarily equal or of the same polarity. Op amps designed for rail-to-rail input operation (parallel PNP and NPN differential stages as described later in this section) have bias currents which are also a function of the common-mode input voltage. External bias current cancellation schemes are ineffective with these op amps also. It should be noted, however, that it is often desirable to match the source impedance seen by the (+) and (-) inputs of VFB op amps to minimize distortion.

CFB op amps generally have unequal and uncorrelated input bias currents because the (+) and (-) inputs have completely different architectures. For this reason, external bias current cancellation schemes are also ineffective. CFB input bias currents range from 5 to 15 μ A, being generally higher at the inverting input

OUTPUT OFFSET VOLTAGE SUMMARY

- **High Speed Bipolar Op Amp Input Offset Voltage:**
 - ◆ Ranges from 1 to 3mV for VFB and CFB
 - ◆ Offset TC Ranges from 5 to 15 μ V/ $^{\circ}$ C

- **High Speed Bipolar Op Amp Input Bias Current:**
 - ◆ For VFB Ranges from 1 to 5 μ A
 - ◆ For CFB Ranges from 5 to 15 μ A

- **Bias Current Cancellation Doesn't Work for:**
 - ◆ Bias Current Compensated Op Amps
 - ◆ Current Feedback Op Amps

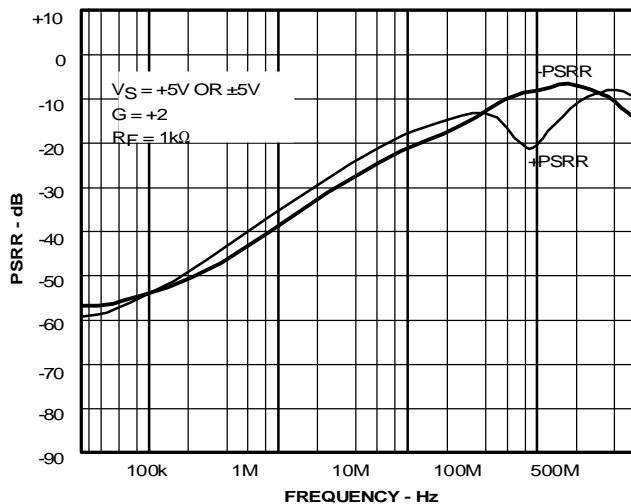


1.34

PSRR CHARACTERISTICS OF HIGH SPEED OP AMPS

As with most op amps, the power supply rejection ratio (PSRR) of high speed op amps falls off rapidly at higher frequencies. Figure 1.35 shows the PSRR for the AD8011 CFB 300MHz CFB op amp. Notice that at DC, the PSRR is nearly 60dB, however, at 10MHz, it falls to only 20dB, indicating the need for excellent external LF and HF decoupling. These numbers are fairly typical of most high speed VFB or CFB op amps, although the DC PSRR may range from 55 to 80dB depending on the op amp.

AD8011 POWER SUPPLY REJECTION RATIO



1.35

The power pins of op amps must be decoupled directly to a large-area ground plane with capacitors which have minimal lead length. It is generally recommended that a low-inductance ceramic surface mount capacitor ($0.01\mu\text{F}$ to $0.1\mu\text{F}$) be used for the high frequency noise. The lower frequency noise can be decoupled with low-inductance tantalum electrolytic capacitors (1 to $10\mu\text{F}$).

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