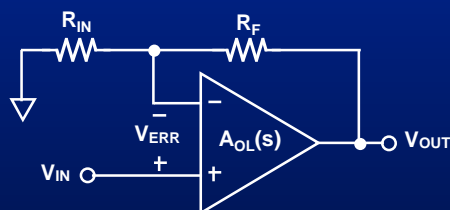


# High Speed Amplifiers

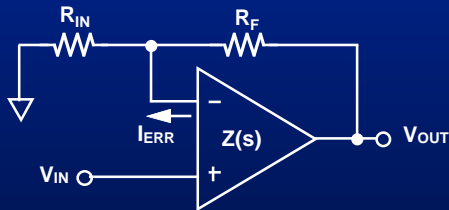
- **Current Feedback Amplifiers vs Voltage Feedback Amplifiers .....6.2**
- **High Speed Amplifier Applications**
  - **Integrator .....6.18**
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Contributing Author: Bonnie Baker

## WHICH IS THE BEST FOR MY APPLICATION?



Voltage Feedback Amplifier



Current Feedback Amplifier

6.2

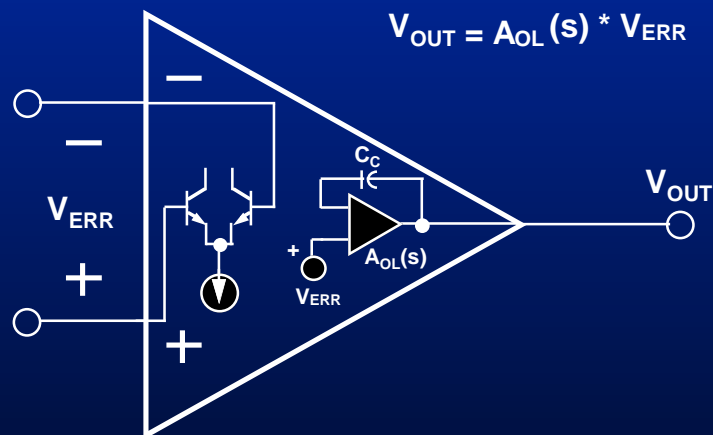


A commonly asked question, when it comes to matters of current feedback amplifiers is, “Why do I need one for my application?” One quick response to that question could be, “You can adjust the gain with the external resistors without adjusting the bandwidth”. A regular user of voltage feedback amplifiers would have a difficult time with this statement because he has been ingrained with the fact that the closed-loop gain is directly related to the closed-loop bandwidth (Gain Bandwidth Product). For most designers, the bandwidth specification is one of the key criterion determining which op amp is right for the application.

Current feedback amplifiers are also rumored to have higher bandwidths, faster slew rates and lower distortion than their voltage feedback amplifier cousins. On the other hand, misconceptions of what the current feedback amplifier is leads designers to believe that the amps are difficult to design with because of the mismatched inputs and the transimpedance open-loop gain.

This tutorial is designed to address the common misconceptions about the current feedback amplifier and prove its worthiness in a variety of applications.

## VOLTAGE FEEDBACK AMPLIFIER BLOCK DIAGRAM



6.3

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The voltage feedback amplifier is the most prolific amplifier on the market. Dependent on the characteristics of the specific amplifier, they are used in high speed as well as precision applications. Since the preferred frame of reference for most analog designers is the voltage feedback amplifier, the comparison and analysis begins with the topology shown above. This simplified block diagram illustrates many of the key characteristics of the voltage feedback amplifier. Starting with the input segment of the diagram, the inputs to the voltage feedback amplifier are evenly matched. As a consequence, input bias currents are very close to being the same magnitude and the difference between the two input bias currents are usually small. Additionally, the input impedances of the amplifier are close to equal and relatively high. For high speed voltage feedback amplifiers, input bias currents of  $1\mu\text{A}$  to  $100\mu\text{A}$  is not uncommon.

A small error voltage at the input of the amplifier is gained by the open-loop gain,  $A_{OL}(s)$ , which is usually fairly high. It is common for this gain to be in the range of 60dB, however, some high speed voltage feedback amplifiers can have open-loop gains of 90dB or better. The open-loop gain is frequency dependent and starts rolling off at a relatively low frequency. The resulting open-loop voltage output is the product of the open-loop gain times the input voltage error.

The compensation capacitor,  $C_C$ , contributes to the stability of the amplifier as well as dominates the slew rate performance.

## VOLTAGE FEEDBACK AMPLIFIER SPECIFICATIONS

Voltage Feedback  
Stable for  $G = > 2$

### GBWP = > 400MHz

PART	SR V/ $\mu$ s	A <sub>OL</sub> dB
OPA643	1000	95
OPA621	350	55
OPA675	350	65
OPA676	350	65
OPA651	300	47

### GBWP < 400MHz

PART	SR V/ $\mu$ s	A <sub>OL</sub> dB
OPA654	750	94
OPA641	650	57
OPA678	350	50
OPA637	100	112

Voltage Feedback  
Unity Gain Stable

### GBWP = > 200MHz

PART	SR V/ $\mu$ s	A <sub>OL</sub> dB
OPA642	380	95
OPA640	350	57
OPA655	290	58
OPA646	180	51
OPA620	175	55

### GBWP < 200MHz

PART	SR V/ $\mu$ s	A <sub>OL</sub> dB
OPA628	310	90
OPA650	240	45
OPA2650	240	45
OPA4650	240	45
OPA671	100	74



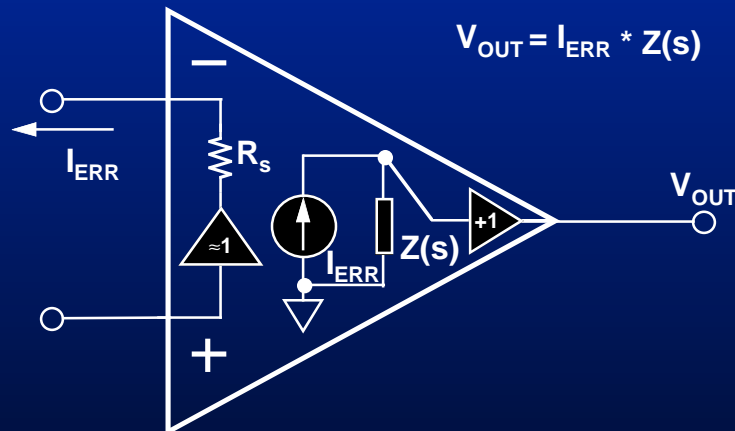
6.4

Some voltage feedback amplifiers are designed to be stable at closed-loop gains of  $\Rightarrow +2V/V$  or internally compensated to be stable at closed-loop gains of  $\Rightarrow +1V/V$ . Voltage feedback amplifiers that are stable at gains of  $+2V/V$  or higher, typically have wider bandwidths, higher slew rates, and lower input voltage noise.

The Gain Bandwidth Product (GBWP) of an amplifier is defined as the frequency at which the open-loop gain of the amplifier would be 0dB if the gain curve were extrapolated 20dB/decade from the dominant pole. This figure of merit does not account for poles and zeros in the amplifier's transfer function at higher frequency. Consequently, it is possible to operate the amplifier with a wider bandwidth than predicted by the GBWP because of the decrease of the phase margin. The trade-off for this increased bandwidth is gain peaking and in some instances instability. The amplifiers listed above have been separated into GBWP classes to assist in the product selection process. The term GBWP has no meaning when discussing the bandwidth performance of a current feedback amplifier (CFB), as will be shown later.

\* NOTE: Bold indicates **NEW PRODUCT**.

## CURRENT FEEDBACK AMPLIFIER BLOCK DIAGRAM



6.5

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The current feedback amplifier's block diagram illustrates how this amplifier differs from the voltage feedback amplifier. The inputs to the current feedback amplifier are not matched, consequently the input bias currents are different along with the input impedances. Typically, the current feedback amplifier's input bias current is in the micro ampere region. The ratio between the input bias currents is dependent on the current feedback input stage topology and can vary from +/- 2X to 5X. The inverting input has a higher input bias current magnitude and very low input resistance (ideally zero) as compared to the non-inverting input. On the other hand, the non-inverting input is buffered, has a high impedance so the magnitude of the input bias current is lower than the inverting input bias current. The buffer's gain is approximately +1V/V and its bandwidth is significantly wider than the bandwidth of the remaining internal stages of the amplifier.

A small error current from the inverting input of the current feedback amplifier is gained by the open-loop transimpedance of the amplifier,  $Z(s)$ , which is usually fairly high. The resulting open-loop output voltage of the current feedback amplifier is the product of the open-loop transimpedance ( $Z(s)$ ) times the input current error ( $I_{ERR}$ ).

## CURRENT FEEDBACK AMPLIFIER SPECIFICATIONS

<u>PART</u>	<u>BW @ G = +2</u> (MHz)	<u>SR</u> (V/ $\mu$ s)
OPA658	650	1700
OPA648	600	1200
OPA2658	500	1700
OPA4658	450	1700
OPA644	300	2500
OPA623	290	2100
OPA603	160	1000

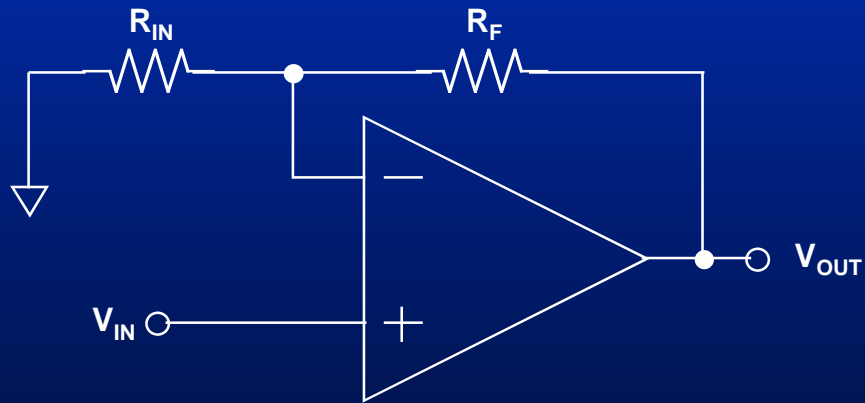
6.6



Current feedback amplifiers are usually optimized and specified to operate at a specific closed-loop gain, which is typically  $+2V/V$ . This approach to current feedback design is used to offer the best amplifier for a majority of the applications that the amplifier will be used in, but it can sometimes be misleading. Current feedback amplifiers can be used over a wide range of gains. All current feedback amplifiers are unity gain stable.

As a family, current feedback amplifiers are touted as having a higher slew rate than the voltage feedback amplifiers. This generalization is true for large and fast input signals. The tail current to the current feedback amplifier's inverting input is mirrored to the tail current of the internal leg that supplies current to the compensation capacitor controlling the slew rate. With high input  $dV/dt$  signals the delta current to the compensation capacitor is increased, consequently the slew rate is higher than if the input signal is small.

## TYPICAL NON-INVERTING AMPLIFIER



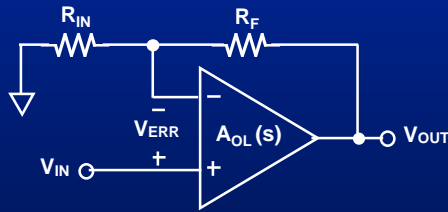
$$\frac{V_{OUT}}{V_{IN}} = 1 + R_F / R_{IN}$$

6.7

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Although the internal topology of the voltage feedback amplifier and current feedback amplifier seem quite different, both amplifier types are suitable for this typical non-inverting amplifier configuration. The low frequency gain,  $V_{OUT}/V_{IN}$ , is set by the resistors,  $R_F$  and  $R_{IN}$ . The feedback mechanism for the voltage feedback amplifier is  $V_{ERR}$ , which is very small, in conjunction with the high open-loop gain,  $A_{OL}$  and the feedback and input resistors. The feedback mechanism for the current feedback amplifier is  $I_{ERR}$ , which is also very small, in conjunction with the high transimpedance gain,  $Z$ , and the feedback resistor. In both cases, the error signal is small, the gain component is large and negative feedback is used to control the system.

## NON-INVERTING AMPLIFIER WITH A VOLTAGE FEEDBACK AMPLIFIER



$$A_{OL}(s) = \frac{G_{DC}}{1 + r_o C_C s}, \quad s = j\omega$$

$$V_{OUT} = A_{OL}(s) * V_{ERR}$$

$$\frac{V_{IN} - V_{ERR}}{R_{IN}} = \frac{V_{OUT} - (V_{IN} - V_{ERR})}{R_F}$$



6.8

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The voltage feedback amplifier can be analyzed across the frequency spectrum in the non-inverting circuit show above. By using the simple model discussed earlier, two equations quickly come out of the calculation. These calculations assume there are no contributions to the frequency behavior of the circuit from the input bias currents of the amplifier. Since this analysis assumes the amplifier is operating in its linear region and this is a small signal analysis, this is a good assumption.

The open-loop gain of the amplifier is modeled as a single pole system. The single pole in the  $A_{OL}(s)$  equation represents the dominant pole. Typically, this pole occurs in the 100s of kHz region. This formula is not an accurate representation of the open-loop gain over the entire frequency spectrum, however, it is adequate for purposes of this discussion. The DC open-loop gain is symbolized with the variable,  $G_{DC}$ . The element,  $r_o$  represents the effective impedance of the open-loop gain equation.  $C_C$  and  $r_o$  are used to set the frequency of the dominant pole.  $C_C$  in conjunction with internal bias currents also dominates the slew rate response of the voltage feedback amplifier.

Rigorous calculation of the transfer function reveals characteristics and limitations of the voltage feedback amplifier in this closed-loop system.



## CALCULATION CONCLUSIONS

$$\frac{V_{OUT}(s)}{V_{IN}(s)} = \frac{(1 + R_F / R_{IN})}{1 + \underbrace{(1 + R_F / R_{IN}) / A_{OL}(s)}_{G_N}}$$

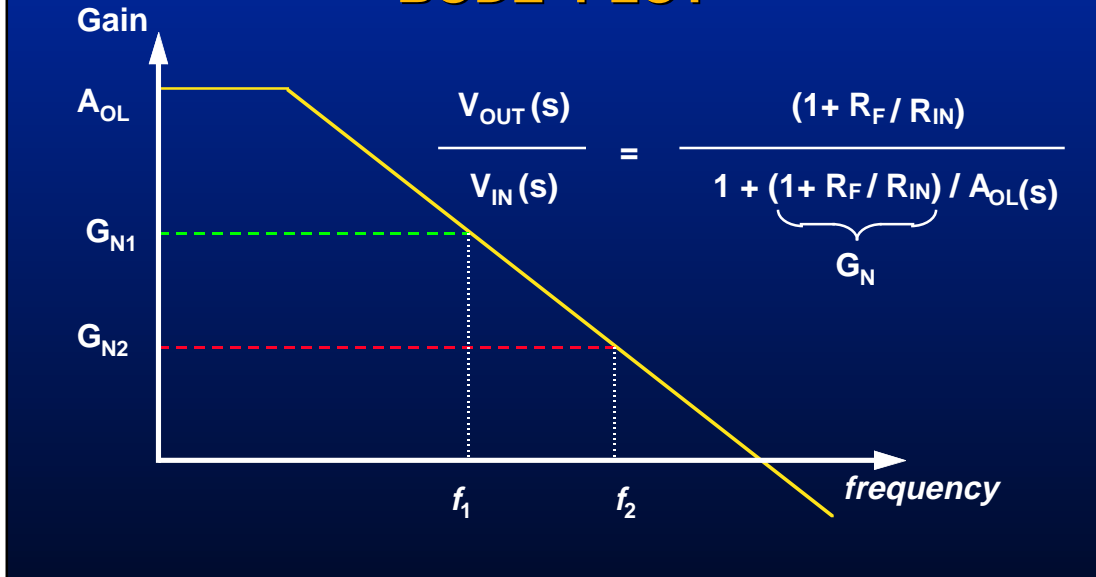
- DC Gain
- Frequency Behavior

6.9



As expected, the calculation proves that the closed-loop DC gain is equal to  $1 + R_F / R_{IN}$ . At low frequencies the open-loop gain of the amplifier is sufficiently high to allow for ignoring the gain error. As frequency increases,  $A_{OL}(s)$  begins to decrease and finally becomes the dominant controlling factor in the gain of the circuit. The calculation of the intersection of the open-loop gain,  $A_{OL}(s)$  and the noise gain ( $G_N$ ),  $(1 + R_F / R_{IN})$  gives a close approximation to the bandwidth of the closed-loop amplifier circuit. Gain peaking, which is caused by the phase response of the amplifier and the feedback circuit, can increase the bandwidth of the closed-loop system at the expense of increased instability. Careful examination of the denominator of this equation points out the dependence of the closed-loop bandwidth on the ratio of input ( $R_{IN}$ ) and feedback resistors ( $R_F$ ) in the circuit.

## VOLTAGE FEEDBACK AMPLIFIER BODE PLOT



The transfer function of the non-inverting amplifier is shown graphically above. The open-loop gain plot of the amplifier assumes a single pole system, which is not completely realistic. Even though this is the case, the generalization of closed-loop gain vs closed-loop bandwidth shown here is still true. As the closed-loop gain increases, the closed-loop bandwidth decreases. The circuit designer needs to take this characteristic under consideration when selecting the right amplifier for his application.

To guarantee stability the phase margin must be greater than 45 degrees. If the open-loop gain curve represented a single pole system, the phase margin would be 90 degrees or higher. Voltage feedback amplifiers and current feedback amplifiers have high frequency poles and zeros in its open-loop transfer function. In the case of voltage feedback amplifiers, increases in closed-loop bandwidth usually decreases the phase margin. A decrease in phase margin can cause gain peaking, giving a higher than expected closed-loop frequency response as well as an overshoot and ringing with the step function response.

## VOLTAGE FEEDBACK AMPLIFIER

- **BENEFITS**
  - Matched inputs
  - DC accuracy
- **DISADVANTAGES**
  - Bandwidth is tied to desired gain

$$\frac{V_{OUT}(s)}{V_{IN}(s)} = \frac{GAIN}{1 + (1 + R_F / R_{IN}) / A_{OL}(s)}$$

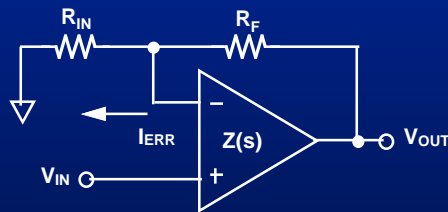
6.11



Matched inputs may or may not be a benefit when using voltage feedback amplifiers in high speed applications. The high impedance can be a saving grace at times when line termination is otherwise difficult. In addition, offset voltages and offset currents are relatively low compared to the current feedback amplifier topology. These offsets are gained by the closed-loop network. These characteristics may or may not be a benefit, recalling that this class of amplifier is typically used in high speed applications and offsets may or may not be an issue because of ac coupling techniques used in the circuit. High speed applications typically use low value resistors because of the bandwidth limitations due to parasitics encountered when high value resistors are used. Micro amps of bias current times k ohms of resistance will give millivolts of offset.

A possible disadvantage of the voltage feedback amplifier is the intimate relationship between the bandwidth and closed-loop gain. Additionally, harmonic distortion is typically not as good as the current feedback amplifier over a wide range of gains.

## CURRENT FEEDBACK AMPLIFIER NON-INVERTING GAIN



$$Z(s) = \frac{R_T}{1 + R_T C_T s}, \quad s = j\omega$$

$$\begin{aligned} V_{OUT} &= I_{ERR} * Z(s) \\ \frac{V_{IN}}{R_{IN}} &= \frac{V_{OUT} - V_{IN}}{R_F} + I_{ERR} \end{aligned}$$



6.12

The current feedback amplifier can also be used in the non-inverting circuit shown above. By using the simple model discussed earlier, two equations quickly come out of the calculation. These calculations assume there are no contributions to the frequency behavior of the circuit from the input offset voltage or buffer stage of the amplifier. Since this analysis assumes the amplifier is operating in its linear region and this is a small signal analysis, these are good assumptions.

The open-loop transimpedance of the amplifier is modeled as a single pole system. The single pole in the equation on the slide represents the dominant pole. Typically, this pole occurs in the 100s of kilohertz region. This formula is not an accurate representation of the open-loop transimpedance over the entire frequency spectrum, however, it is adequate for purposes of this discussion. The DC open-loop transimpedance is symbolized with the variable,  $R_T$ .  $C_T$  and  $R_T$  are used to derive the frequency of the dominant pole.  $C_T$  in conjunction with the transient inverting error current also dominates the slew rate response of the current feedback amplifier.

Rigorous calculation of the transfer function reveals characteristics and limitations of the current feedback amplifier in this closed-loop system.

## CALCULATION CONCLUSIONS

$$\frac{V_{OUT}(s)}{V_{IN}(s)} = \frac{(1 + R_F / R_{IN})}{1 + (R_F) / Z(s)}$$

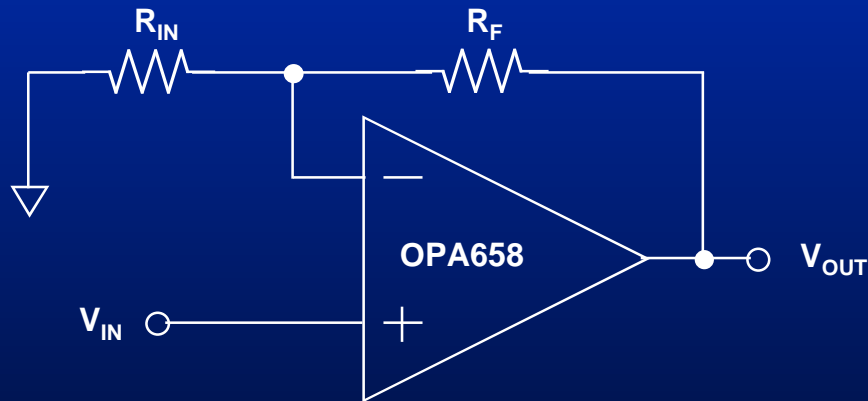
- DC Gain
- Frequency Behavior

6.13



The DC gain of this circuit is the same regardless of whether a current feedback or voltage feedback amplifier is used. The bandwidth of closed-loop response, when a current feedback amplifier is used, is dependent on feedback resistor,  $R_F$ , in conjunction with the transimpedance of the amplifier. The resistor,  $R_{IN}$ , does not effect the bandwidth of the circuit as it would if a voltage feedback amplifier was used in the circuit. This fundamental difference in the closed-loop response between the two amplifier topologies allows for each to have an advantage or disadvantage, as the case may be, dependent on the circuit topology selected.

## DESIGN PROBLEM NON-INVERTING



- Define required bandwidth
- Select your amplifier and  $R_F$
- Define the required closed-loop gain
- Select  $R_{IN}$

6.14

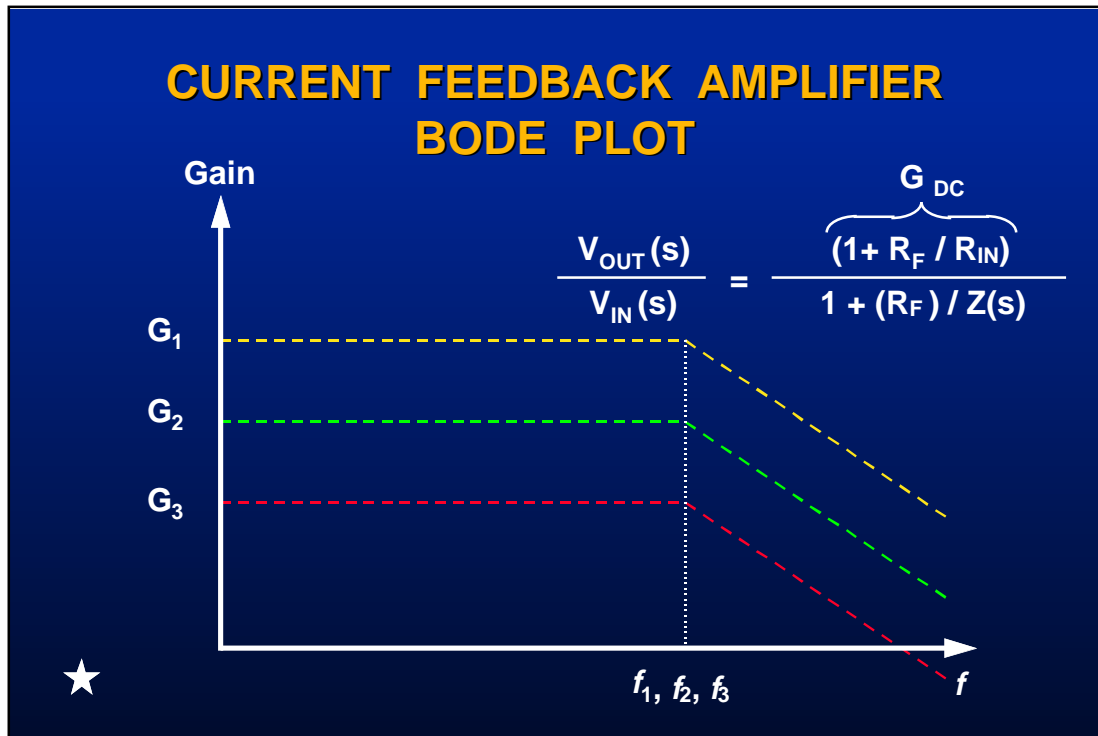


A circuit design problem using a current feedback amplifier uses a simple, straight forward process. Initially, the required closed-loop bandwidth must be determined as required by the application. From that specification, an appropriate current feedback amplifier can be selected. In the current feedback specification sheet, the appropriate feedback resistor is suggested.

The closed-loop gain is then determined as dictated by the application. The appropriate input resistor,  $R_{IN}$ , is selected according to the closed-loop gain requirements. In the event the closed-loop gain is changed, a new  $R_{IN}$  can be selected, leaving  $R_F$  constant.

The circuit design problem is a bit more complex when using a voltage feedback amplifier. The required closed-loop bandwidth and gain must be known from the beginning. The appropriate amplifier can then be selected by estimating the closed-loop bandwidth vs gain from the typical performance curves given in the specification sheet. In the event that the gain needs to be adjusted, it is possible that complicated compensation techniques may be required or that another amplifier, with different bandwidth characteristics, will be needed.

## CURRENT FEEDBACK AMPLIFIER BODE PLOT



Ideally, the closed-loop frequency bandwidth is independent of changes in  $R_{IN}$  so it is possible to adjust the closed-loop gain without changing the bandwidth. This conclusion is reached using assumptions that take into account first order effects of a current feedback amplifier in a closed-loop system. Ideally, the current feedback amplifier can be viewed as a single pole transimpedance system with infinite impedance at the non-inverting input and zero impedance at the inverting input. Additionally, the buffer gain between the inverting and non-inverting input is  $+1V/V$  with zero offset voltage. When these assumptions are used, it is easy to derive the relationship between the feedback resistor,  $R_F$ , the input resistor,  $R_{IN}$ , and the closed-loop bandwidth performance as is discussed in the previous sections.

When these assumptions are re-examined, it can be shown that second order effects have a small impact on the closed-loop bandwidth. In the equation below, alpha represents the gain of the input buffer, which is typically  $+0.996V/V$  as opposed to  $+1V/V$ .  $R_S$  represents the non-zero output impedance of the input buffer, which ranges from 10 to 40 $\Omega$  depending on the particular amplifier used.

$$\frac{V_{OUT}(s)}{V_{IN}(s)} = \frac{\alpha(1+R_F/R_{IN})}{1 + (R_F + R_S(1+R_F/R_{IN}))/Z(s)}$$

From the formula above, it is easy to see the limitations on the current feedback amplifier's frequency response performance. Because of the effects of  $R_S$ , the closed-loop bandwidth does vary slightly with changes in  $R_{IN}$ . In addition, the low frequency gain is attenuated by alpha.

## CURRENT FEEDBACK AMPLIFIERS

$$\frac{V_{OUT}(s)}{V_{IN}(s)} = \frac{GAIN}{1 + (R_F) / Z(s)}$$

- **BENEFITS**
  - Ease of Design
  - Dominant pole is higher/lower distortion
  - Bandwidth is dependent on  $R_F$
- **DISADVANTAGES**
  - DC Bias Current
  - Current Noise

6.16

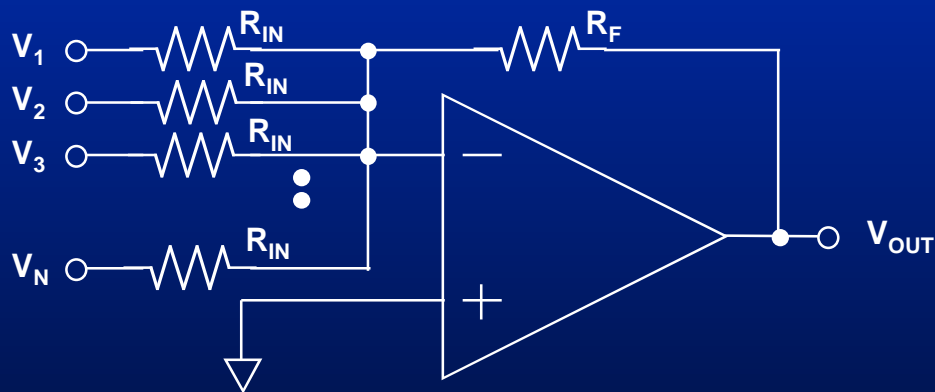


The current feedback amplifier offers more ease in the design process than the voltage feedback amplifier. The dominant pole of the current feedback open-loop transimpedance gain is higher in frequency than the voltage feedback open-loop gain dominant pole. Consequently, current feedback amplifiers have lower gain distortion as the signal increases in frequency. The bandwidth of the current feedback amplifier in a closed-loop configuration is dependent and adjustable with the feedback element.

Some of the disadvantages of the current feedback amplifier are, the DC input bias currents are mismatched and the current noise of the inverting input is higher than the non-inverting input.



## DESIGN PROBLEM INVERTING



### Current Feedback

$$\frac{V_{OUT}(s)}{V_{IN}(s)} = \frac{-R_F/R_{IN}}{1 + (R_F) / Z(s)}$$

### Voltage Feedback

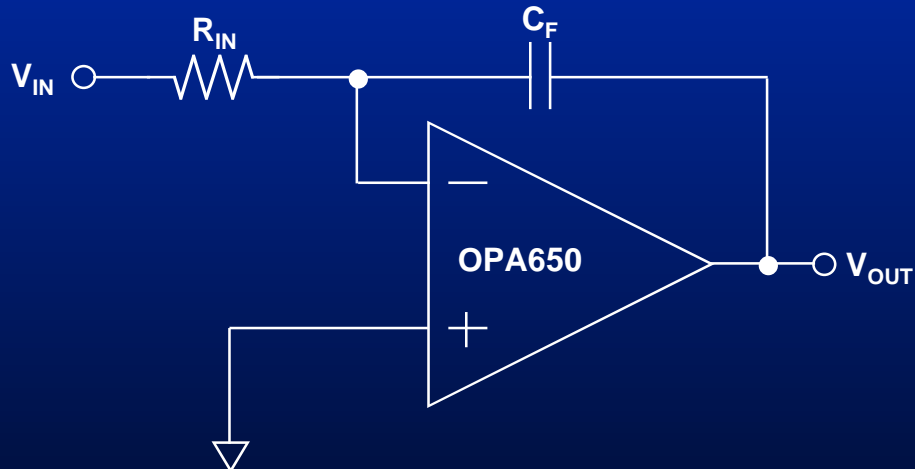
$$\frac{V_{OUT}(s)}{V_{IN}(s)} = \frac{-R_F/R_{IN}}{1 + (1+nR_F/R_{IN}) / A_{OL}(s)}$$

Following similar methods of calculation, the gain and frequency response of an inverting amplifier circuit can be derived. Note that the bandwidth of the circuit with a voltage feedback amplifier changes with changes in input resistance. Also note that the closed-loop bandwidth is smaller than expected. Since current feedback amplifiers depend on  $R_F$  to set the closed-loop frequency response, it is possible to change gain without changing bandwidth.

This amplifier circuit implements a simple summing function. Both the current feedback and voltage feedback amplifier will work in this circuit, however, if the number of inputs are changed on the fly, the configuration with the voltage feedback amplifier will change signal bandwidths. In the event that multiplexed inputs are required, the effective input resistance changes according to the number of signals multiplexed in at any particular time. This is easily illustrated by the transfer function of the circuit with the voltage feedback amplifier. In this case,  $R_{IN}$  and  $n$  are in the numerator of the ratio that determines the frequency response of the voltage feedback gain formula.

The current feedback amplifier performs best in this situation because of its relative immunity to changes in the effective non-inverting gain.

## INTEGRATOR AMPLIFIER



**BEST WITH VOLTAGE FEEDBACK AMP**

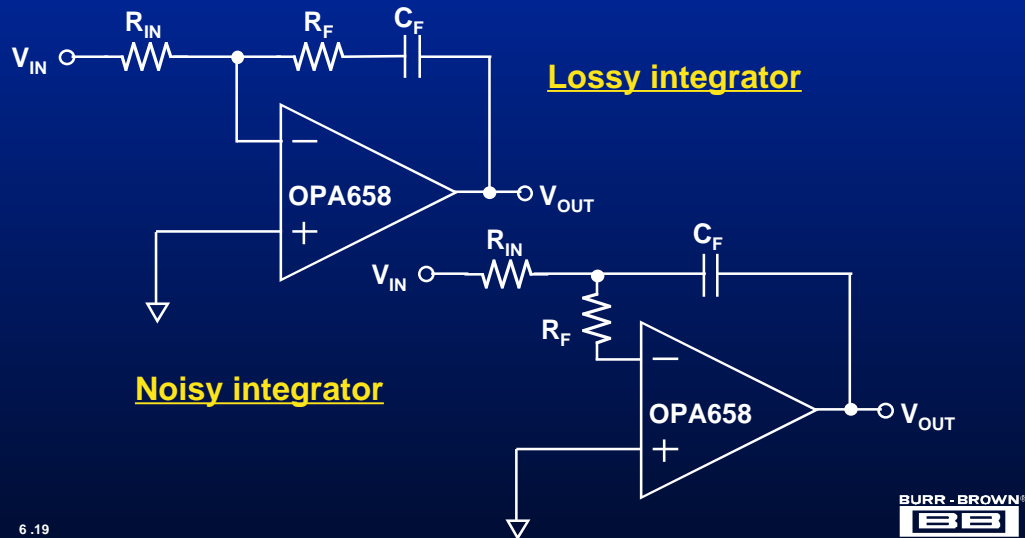


6.18

Integrators are a natural for many circuit applications, but using the current feedback amplifier in this configuration would be a mistake. The voltage feedback is the best choice when you consider that the feedback element,  $R_F$ , does not exist. With current feedback amplifiers, a capacitive element alone (without any resistance in series) in the feedback loop will make the circuit unstable.

As for voltage feedback amplifiers, unity gain stability is a requirement. The closed-loop gain of this circuit at higher frequencies is dominated by the capacitors and equal to  $(1 + C_{IN} / C_F)$ , where  $C_{IN}$  is equal to the parallel combination of the amplifier input capacitance and the input resistor parasitic capacitance and  $C_F$  is equal to the capacitance in the feedback loop of the circuit. To insure stability with these amplifiers, the high frequency gain equation,  $(1 + C_{IN} / C_F)$ , must be equal to or greater than the specified stable gain of the amplifier used.

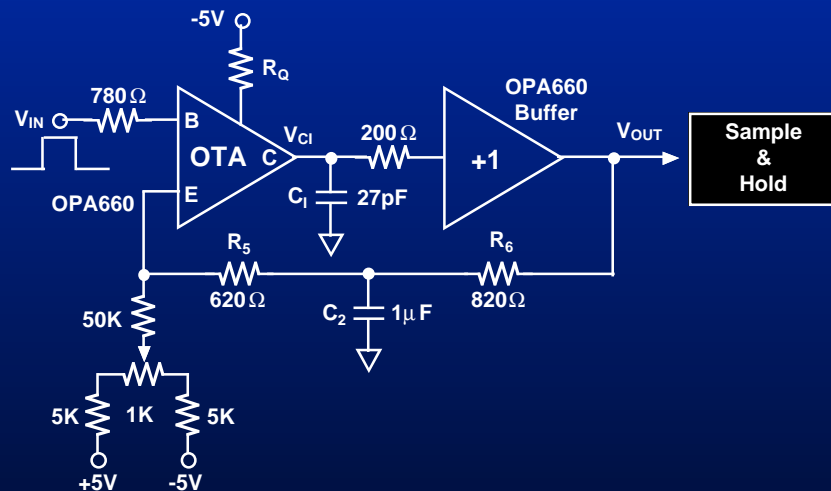
## INTEGRATORS USING CURRENT FEEDBACK AMPLIFIER



It is possible to use current feedback amplifiers as integrators as long as the required feedback resistance is in the circuit. In the first diagram the required feedback resistor is placed in the feedback loop in series with the integrating capacitor. The appropriate feedback resistor for this circuit is the recommended manufacturer's value. Although this circuit will perform the integration function, there is some degradation of signal bandwidth.

In the second diagram, the required feedback resistor is placed in series with the inverting input of the amplifier. The current feedback amplifier loop requirements are fulfilled with the position of  $R_F$ . Although, stability is achieved with  $R_F$ , a relatively high voltage noise source is introduced into the circuit. Typically, the current noise from the inverting input of the current feedback amplifier is significantly higher than the non-inverting input of the same amplifier as well as higher than most voltage feedback amplifiers. This current noise is multiplied by the resistor,  $R_F$ , and then multiplied by the closed-loop noise gain of the circuit.

## NANOSECOND INTEGRATOR



6.20



A transconductance op amp can be configured as a “nanosecond integrator” as illustrated here with the OPA660. This circuit can process incoming pulses that have an amplitude of up to +/-2.5V and as short as 8ns in duration. The OPA660 will respond to a 2ns risetime.

A transconductance amplifier (OTA), like the OPA660, is a voltage-controlled current source, which is particularly useful when the load is a capacitor. The relationship between the voltage across the load capacitor,  $C_1$ , and the OTA output current is:  $V_C = I * t / C_1$ . The transfer function of the integrator is:

$$V_{OUT} = \frac{g_m}{C_1} \int V_{BE} dt$$

where  $C_1$  = integration capacitor,  $V_{BE}$  = the voltage across the input terminals of the OPA660 (B and E) and  $g_m$  is the transimpedance of OPA660, adjustable with an external resistor,  $R_Q$ .

The output voltage is equal to the time integral of the input voltage. Two constants influence the output voltage; the transconductance ( $g_m$ ) of the amplifier and the external capacitor,  $C_1$ . The transconductance, which is essentially the gain of the OTA can be varied over a wide range. This allows some flexibility to the voltage level of the pulses and the selection of the integration capacitor,  $C_1$ . The equation above shows that the capacitor has a reciprocal affect to the output voltage, consequently, the smaller the capacitor the higher the voltage,  $V_{OUT}$ . The integrated signal has a DC feedback path to the emitter pin through the low-pass filter,  $R_5$ ,  $R_6$  and  $C_2$ . This counteracts the effects of the bias currents of the OTA and the buffer integrating on  $C_1$ . The output offset voltage can be adjusted to zero using the potentiometer,  $R_8$ .

## NANOSECOND INTEGRATOR TEST RESULTS

Channel 1  
Input  
2V/DIV

Channel 2  
Output  
2V/DIV



10ns/DIV



6.21

The test results of the previous circuit is shown here. The OTA charges the integration capacitor,  $C_I$ , linearly according to the equation:

$$V_{C_I} = (V_{BE} * g_m * t) / C_I$$

With:

$V_{C_I}$  = voltage across capacitor  $C_I$

$V_{BE}$  = base - emitter voltage of the OPA660 OTA section

$g_m$  = transconductance of the OTA

$t$  = time

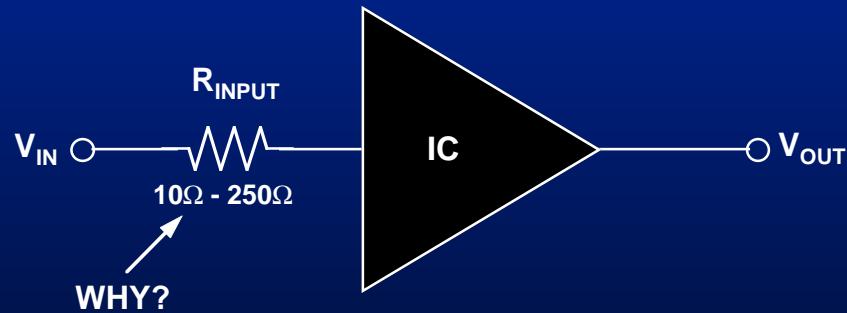
$C_I$  = integration capacitor

The voltage across the capacitor increases linearly as predicted by the equation, as shown with channel 2 in the diagram above. At the end of the input pulse, the voltage can be sampled by a sample/hold amplifier. The delay between the input pulse and the charging of the capacitor is approximately 250ps. This corresponds to the group delay time of the OPA660. The group delay time can be calculated taking the frequency where the open-loop gain has reduced to 0dB and calculated using the equation:

$$t_d = 1/2 \pi f_{0dB}$$

To avoid integration error, the signal delay time of the op amp should be less than 1/20th of the pulse width.

## SERIES INPUT RESISTORS FOR HIGH SPEED AMPLIFIERS



IC: OPA622, OPA623, OPA660, OPA2662, BUF600, BUF601, MPC10X, SHC615

6.22

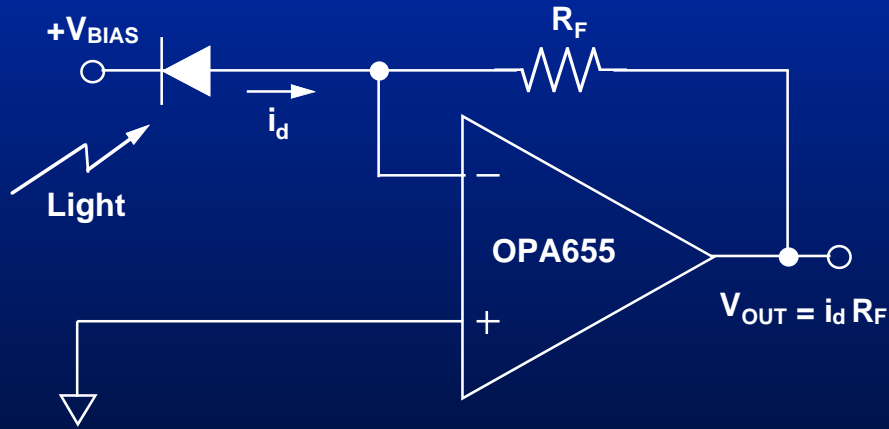


Sometimes high-speed amplifiers need a series input resistor, because package parasitics become more and more apparent at higher signal frequencies. Package parasitics are mainly due to the leadframe pins, bondwire and the IC die itself. The pins and bondwire can be modeled as high frequency inductors, with small capacitors between each. The die adds parasitic capacitance from the bondpad on the die to the die substrate.

All together, these parasitics can form resonant circuits, with high Q values and resonant frequencies in the range of 700MHz to 1GHz. Most problems that are created by these parasitics occur at the high impedance input of the IC. Even if the overall bandwidth of the IC is much less than the resonant frequency, the transistors in the input stage can still be affected. An indication of problems associated with the parasitics is higher than expected gain peaking of the amplifier. A series input resistor will help prevent excessive gain peaking problems or even oscillation by dampening the parasitic LC circuit. Typical values for this resistor are between  $10\Omega$  to  $250\Omega$ . The value can vary widely because of different PC-board parasitics that will add to this problem.

One rule, however, exists: the smaller the package the less its parasitics and the smaller the associated effects. Therefore, designers should choose SOIC packages over the DIP packages whenever possible.

## APPLICATION TRANSIMPEDANCE



**BEST WITH VOLTAGE FEEDBACK AMP**

6.23

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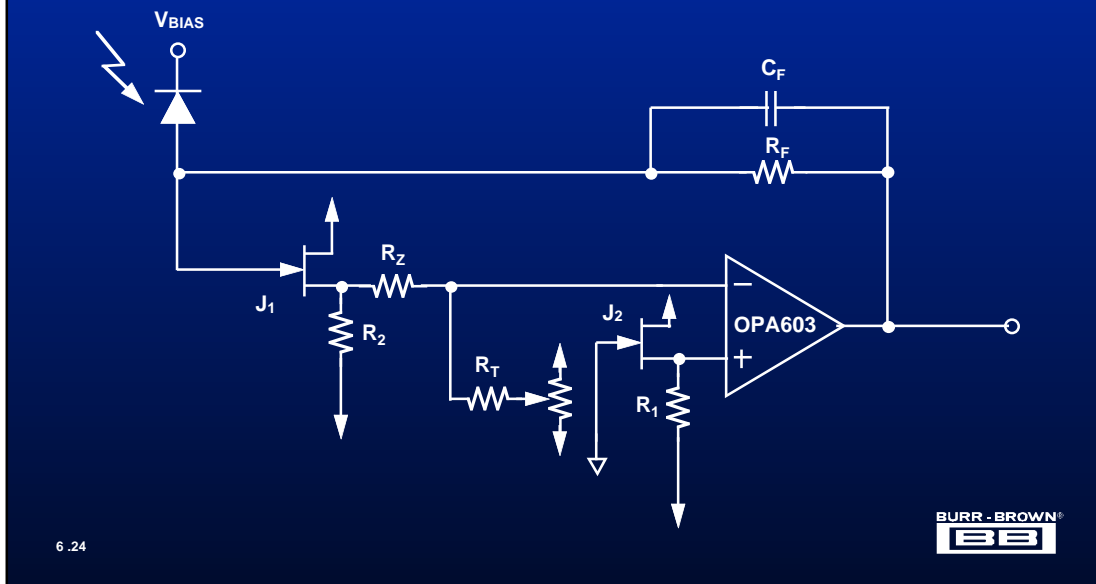
Photodiode preamp circuits are used for a wide variety of applications all involving sensing light and converting that information to a useful voltage. The photodiode that is sensing the light can be configured with or without a bias voltage. If speed and response time are important, the photodiode is typically configured with a reverse bias voltage to lower the junction capacitance, as shown in the figure.

Voltage feedback amplifiers are a natural for transimpedance amplifier circuits. Typically, the photodiode is selected for its responsivity and physical dimensions. The gain is then adjusted by changing  $R_F$ . If this was done with the current feedback amplifier in the circuit, the bandwidth would change with the gain adjustments, which could be inefficient.

On the other hand, current feedback amplifiers can be used in this circuit. It is erroneous to conclude that the current feedback amplifier is not appropriate because of its input bias current. The difference between the DC input bias currents between the voltage feedback amplifiers and the current feedback amplifiers are not that great. For instance, the input bias currents of the OPA642 voltage feedback amplifier are typically  $18\mu\text{A}$  and the input bias currents of the OPA644 current feedback amplifier are typically  $2\mu\text{A}$  for the non-inverting input and  $20\mu\text{A}$  for the inverting input. The unity gain bandwidths of both amplifiers are close to the same,  $450\text{MHz}$  and  $300\text{MHz}$ , respectively. Additionally,  $I_{ERR}$  is wrongly perceived as a detrimental current to this type of application.  $I_{ERR}$  with the current feedback amplifier, like  $V_{ERR}$  with the voltage feedback amplifier is relatively small and generally does not interfere with the overall operation of the circuit.

If input bias currents in the micro ampere range are too large or cause unacceptable offset errors in the circuit, alternative circuits can be implemented. One topology, would use discrete FET transistors and a high speed amplifier. Another design approach would use a voltage feedback amplifier with a FET input like the OPA655. The OPA655 FET input voltage feedback amplifier is in a class of its own having a typical unity gain bandwidth of  $400\text{MHz}$  and input bias currents of  $5\text{pA}$ .

## DISCRETE FET INPUT FOR HIGH SPEED AMPLIFIER



The high input bias currents of current feedback amplifiers can be buffered with JFETs to give the desirable combination of constant bandwidth with minimum input loading. In this circuit the JFETs are configured as source followers and do not need to be biased to zero volts  $V_{GS}$ . The only real trick to this circuit is the compensation resistor,  $R_Z$ . In manufacturer's data sheets, the optimum value for the feedback resistor ( $R_{FB}$ ) is recommended in order to concurrently achieve wideband and stable performance. For this circuit, the summation of  $R_Z$  plus the JFET transconductance should equal  $R_{FB}$ . The feedback resistor,  $R_F$ , is then selected to optimize the dynamic response of the photodiode.

As shown in the figure, a JFET buffered OPA603 (current feedback amplifier) is configured as a photodiode transimpedance amplifier. The 2N5911 ( $J_1$  and  $J_2$ ) input transistors are resistively biased since there is no common-mode swing. The compensation resistor,  $R_Z$ , is selected in order to achieve 55 degrees of phase margin in unity-gain. A feedback capacitor helps to eliminate the peaking that would normally result from the feedback pole created by  $R_F$  and the parasitic capacitance at the inverting input. With the values listed below, the circuit has a 2MHz bandwidth.

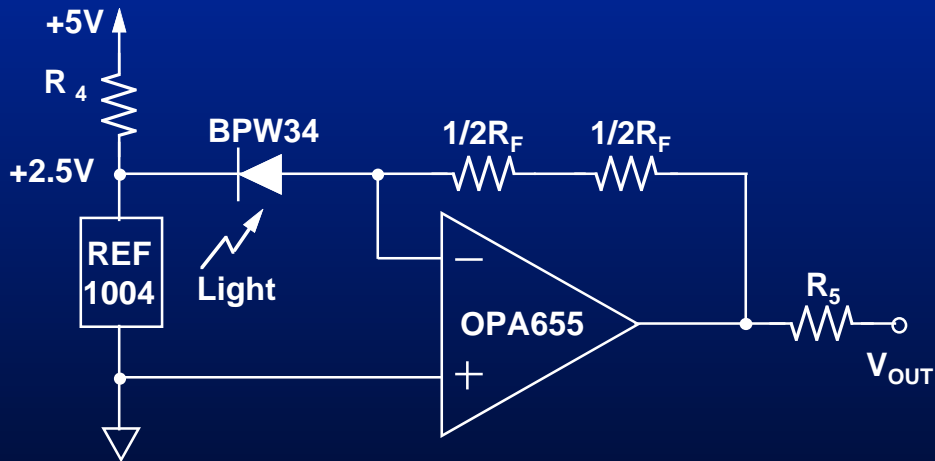
Special attention should be paid to the circuit layout. As with all current feedback amplifiers the inverting input should be kept as low capacitance as possible. Ground planes should be removed in the vicinity of the inverting input. The junction of  $R_Z$  and  $R_T$  should be soldered as close to the amplifier pin as possible and the resistor lead lengths should be kept as short as possible. These practices should be observed for the feedback network, also, if the frequency of operation is expected to be in the MHz region.

$R_1 = 6K\Omega$	$R_2 = 6K\Omega$
$R_Z = 1.42K\Omega$	$R_F = 100K\Omega$
$C_F = 1pF$	$C_{\text{photodiode}} = 10pF$
$R_{\text{photodiode}} = 100M\Omega$	$J_1 = J_2 = 2N5911$

$R_T$  is used to adjust the output offset to zero and is usually in the  $M\Omega$  range.



## JFET AMPLIFIER IN TRANSIMPEDANCE CIRCUIT



6.25



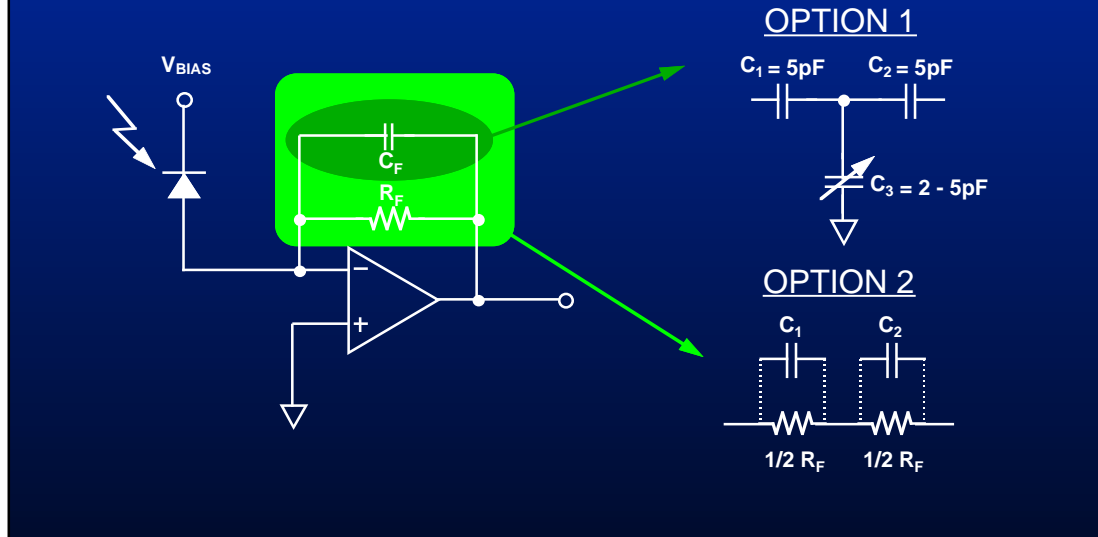
The complete circuit implementation for a transimpedance amplifier using the OPA655 is shown above. In case the transimpedance circuit exhibits gain peaking, it is very difficult to implement the appropriate compensation capacitor,  $C_F$ , at the risk of lessening the signal bandwidth. If a single feedback resistor equal to or higher than  $0.5\text{M}\Omega$  is used, the feedback capacitor may need to have a value well below one  $\text{pF}$ . A split feedback resistor allows the feedback capacitance due to parasitics around the amplifier to have manageable values of a few tenths of a  $\text{pF}$ .

The reference circuit, such as the REF1004-2.5 which is a 2.5 volt reference, can be added to the circuit to provide a stable reverse bias voltage across the sensor diode. The OPA655 is also used as the cable driver, using  $R_5$  for the  $50\Omega$  termination resistor.

The tested configurations and their results are shown below.

BPW34 Capacitance	$R_F$	$f_{-3\text{dB}}$
38pF ( $V_{\text{BIAS}}=1\text{V}$ )	2 x 309K $\Omega$	1.7MHz
38pF ( $V_{\text{BIAS}}=1\text{V}$ )	2 x 249K $\Omega$	1.9MHz
27pF ( $V_{\text{BIAS}}=2.5\text{V}$ )	2 x 274K $\Omega$	1.9MHz
20pF ( $V_{\text{BIAS}}=5\text{V}$ )	2 x 274K $\Omega$	2.0MHz
16pF ( $V_{\text{BIAS}}=10\text{V}$ )	2 x 274K $\Omega$	2.2MHz
16pF ( $V_{\text{BIAS}}=10\text{V}$ )	2 x 309K $\Omega$	1.88MHz
16pF ( $V_{\text{BIAS}}=10\text{V}$ )	1 x 549K $\Omega$	2.18MHz

## REDUCING CIRCUIT FEEDBACK CAPACITANCE

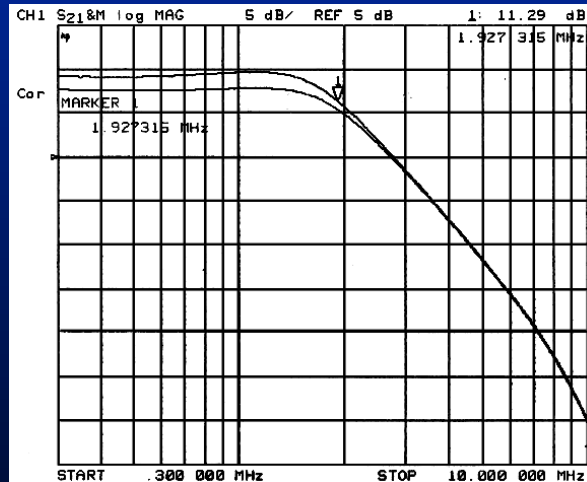


The final transimpedance amplifier solution should be sufficiently stable with a wide enough bandwidth to accommodate the speed of the input signal. The variables in this design problem are the photodiode, the op amp and the op amp's feedback network.

For high speed applications, it is difficult to achieve optimum results due to the pole set by the feedback capacitor,  $C_F$ , and the feedback resistor,  $R_F$ . In order to increase the pole frequency of the feedback loop and increase the bandwidth response,  $C_F$  must be designed at a low value, typically less than 2pF. Since this is uncommonly low, two circuit options are recommended to achieve this performance. Option 1, the T-network uses two 5pF capacitors and a trim capacitor to design the low value capacitor. The effective capacitance of this circuit is equal to  $(C_1 C_2) / (C_1 + C_2 + C_3)$ .

If an inexpensive sub-pico farad capacitance is required, option 2 is recommended. With this series resistor topology, the resistance adds and the discrete and/or parasitic capacitances divide. The effective resistance of this network is  $1/2R_F + 1/2R_F = R_F$  and the effective capacitance of this circuit is  $(C_1 C_2) / (C_1 + C_2)$ . This option can be implemented with one capacitor, such as  $C_1$ , and no discrete capacitor for  $C_2$ . In this situation,  $C_2$  would be replaced by parasitic capacitance of the resistor and PCB, which could easily be as low as 0.2pF by using standard RN55D resistors and careful layout techniques. As a consequence, the circuit can be designed with feedback capacitance lower than the capacitance achievable with a single resistor and no discrete capacitor.

## JFET TRANSIMPEDANCE AMPLIFIER FREQUENCY PERFORMANCE



OPA655

6.27



The small signal bandwidth of this transimpedance amplifier is illustrated in this diagram. Tests were performed using the HP Network Analyzer, 8753A. In both cases the feedback resistor,  $R_F$ , is  $2 \times 274\text{K}\Omega$ . With trace #1, the photodiode was reverse biased with 2.5V, causing a parasitic capacitance across the photodiode of 27pF. The -3dB bandwidth of this trace is measured at 1.927315MHz. The effective capacitance in the feedback loop is  $\sim 0.151\text{pF}$ . Notice the small amount of gain peaking of approximately 1dB.

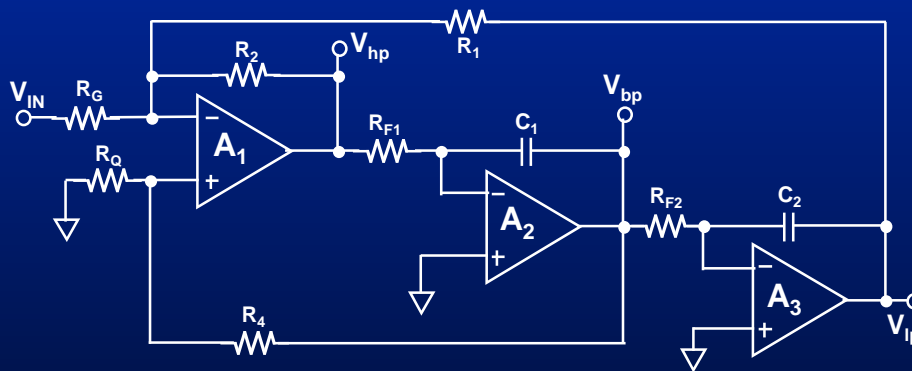
In trace #2, the photodiode was reverse biased with 5V, causing a parasitic capacitance across the photodiode of 20pF. The signal bandwidth is slightly increased to 2MHz.

To achieve this performance, care should be taken to remove the ground plane from areas where the inverting input and feedback resistors are.

The OPA655 high speed voltage feedback amplifier has a FET input stage to ensure low input bias currents resulting in low DC errors and very low noise making the device a good choice for high speed integrators and transimpedance stages.

Transimpedance amplifiers are used in a variety of applications, ranging from precision measurements, such as medical blood analyzer circuits, to high speed designs, such as fiber optic receiver circuits. At the risk of over generalization, precision circuits typically require amplifiers with low offset voltage, input bias current, input capacitance, voltage noise and current noise. With high speed designs, the amplifier's slew rate, bandwidth, input capacitance and the circuit's parasitic capacitance are critical to achieve high speed performance. The combination of precision and speed becomes challenging because of the limited selection of amplifiers available on the market.

## STATE VARIABLE FILTER



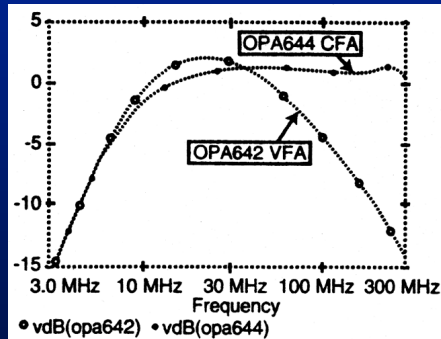
NEEDS BOTH TYPES

6.28

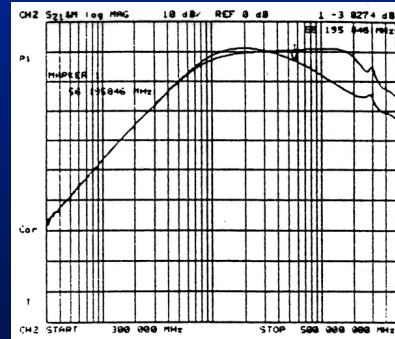
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The state variable filter is an excellent choice of topology if low pass, band pass and high pass filters are needed for concurrent outputs. From previous discussion,  $A_2$  and  $A_3$  are configured as integrators and should be voltage feedback amplifiers.  $A_1$  is the wild card in this circuit. Since  $R_G$  adjusts the gain of the circuit, a current feedback amplifier is more suitable rendering a wider overall bandwidth to the circuit. This becomes critical with the high pass filter.

## STATE VARIABLE FILTER FREQUENCY RESPONSE



SIMULATED



MEASURED

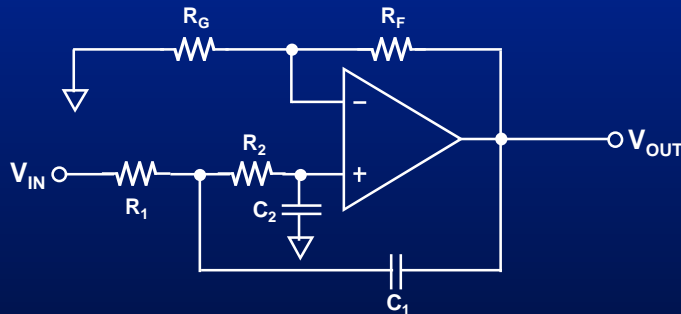
PERFORMANCE RESULTS

6.29



The results for the frequency performance of the state variable filter high pass output is shown in this slide. The graph on the left shows the Spice simulation of the circuit performance. Of the two curves, the bottom most curve shows the performance of the circuit with a voltage feedback amplifier, OPA642, used for all three amplifiers,  $A_1$ ,  $A_2$ , and  $A_3$ . The top most curve shows the performance of the circuit using a current feedback amplifier, OPA644, in the  $A_1$  position of the circuit and two voltage feedback amplifiers, OPA642, for  $A_2$  and  $A_3$ . The graph on the right shows the actual performance of the two circuits. Note the attenuation of both plot responses around 200MHz. This behavior is caused by layout parasitics.

## SALLEN - KEY 2nd - ORDER LOW PASS



BEST WITH CURRENT FEEDBACK AMP

6.30

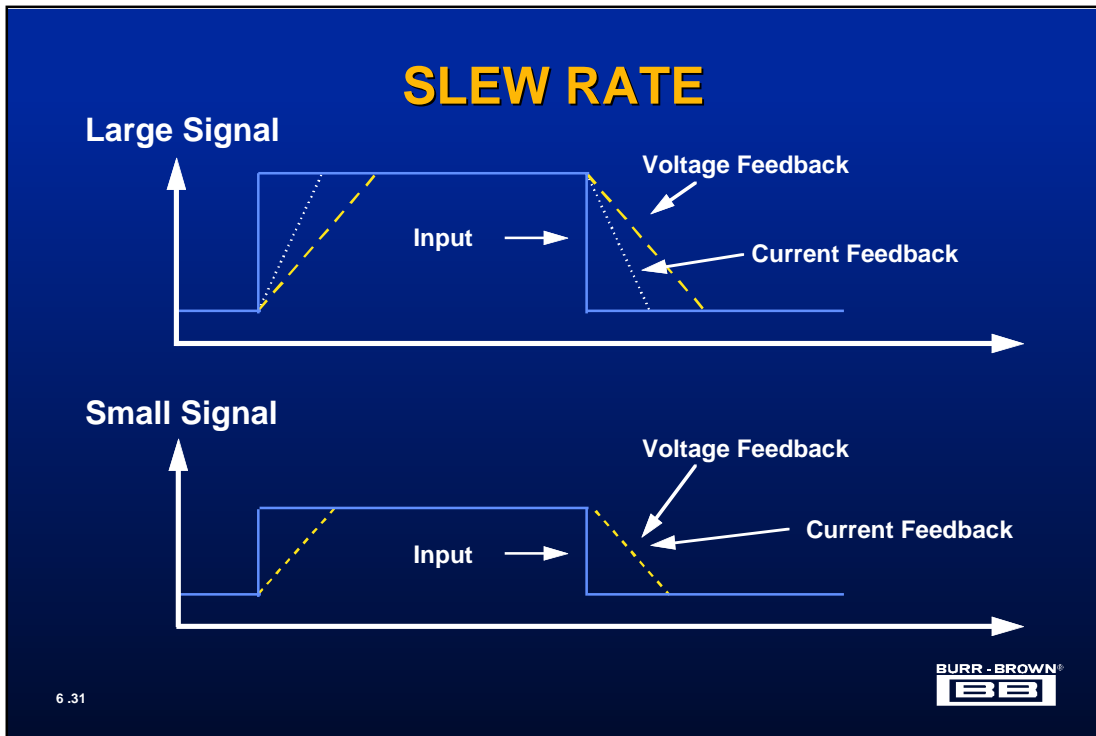


This 2nd-order Sallen-Key low pass filter is distinguished from other filter topologies by its use of a non-inverting gain and a passive RC positive feedback network. The Q of the circuit (for  $R_1 = R_2$  and  $C_1 = C_2$ ) is equal to  $Q = 1/(3 - K)$ , where K is the closed-loop DC gain ( $1 + R_F / R_G$ ). When compared to the State-Variable filter configuration, this filter is more sensitive to component tolerances and gain accuracy is dependent on the ratio of  $R_F : R_G$ . The useable Q range is confined to  $Q < 5$ .

If the filter is configured for unity gain (ie.  $R_G$  is open and  $R_3$  is shorted), a current feedback or voltage feedback amplifier can be used for the application. Additionally, if a large gain is implemented in the circuit the voltage feedback amplifier bandwidth will show a reduction with increasing gains making the current feedback topology a better choice. The transfer function for this circuit using an ideal amplifier is:

$$\frac{V_{OUT}}{V_{IN}} = \frac{G / R_1 R_2 C_1 C_2}{s^2 + s (1/R_1 C_1 + 1/R_2 C_1 + (1 - G)/R_2 C_2) + 1 / R_1 R_2 C_1 C_2}$$

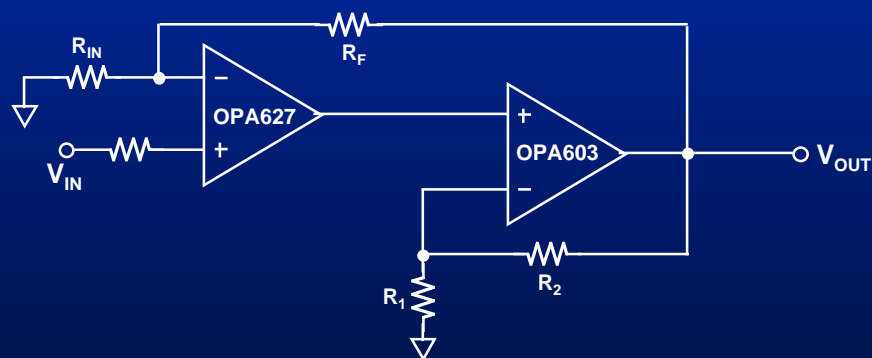
where,  $G = 1 + R_F / R_G$



Voltage feedback amplifiers have slew rate limiting built into the internal circuitry. The slew rate is dominated by the tail currents of the bias strings and the size of the compensation capacitor. The slew rate is independent of the input swing or edge rate.

Current feedback amplifiers do not have the same slew rate limiting constraints as the voltage feedback amplifier. A major advantage of a current feedback amplifier is the absence of slew rate limiting. This is accomplished with the unique topology of the input stage of the current feedback amplifier. As the input signal to the amplifier changes current becomes available to charge the internal compensation capacitance. This current is proportional to the initial voltage imbalance and divided by  $R_{IN} \parallel R_F$ . As the size of the initial voltage lessens, the slew rate is reduced as well.

## COMPOSITE AMPLIFIER



USES BOTH TYPES

6.32



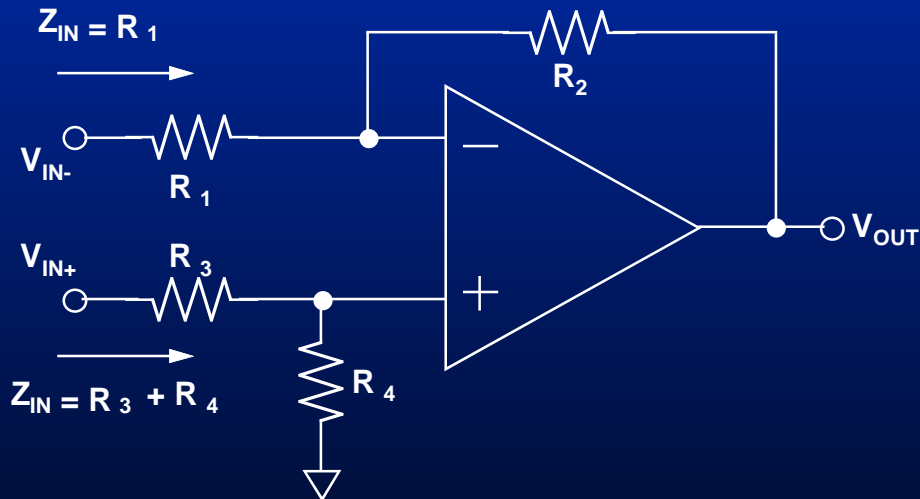
A composite amplifier uses a precision voltage feedback amplifier, OPA627, and the OPA603 current feedback amplifier to utilize the best qualities of both amplifiers. Precision voltage feedback op amps such as OPA627 have excellent performance with precision specifications where the closed-loop gain is low compared to the open-loop gain of the amplifier. However, starting at relatively low frequencies, the loop gain rolls-off at 20dB/decade as the signal frequency increases. This can produce significant errors at higher frequencies where the loop gain can be very low.

Current feedback amplifiers, such as the OPA603, have good dynamic performance at both low and high gains. This is because the feedback component ( $R_2$ ) sets the bandwidth and the input resistor ( $R_1$ ) then sets the closed-loop gain. Unfortunately, the DC ( $V_{OS}$ ,  $dV_{OS}/dT$ , CMR, etc.) performance of current feedback op amps is poor compared to the precision voltage feedback amp.

The combination of the two amplifier types in this circuit render some impressive results. For example, an overall gain of 20V/V and a signal bandwidth of 30MHz can be obtained by placing the OPA603 in a gain of 12V/V ( $R_2 = 1.02K\Omega$ ). Slew rate and full-power response of the OPA627 are boosted in the composite amplifier. Since the OPA603 adds gain at the output of the OPA627, the slew rate of the OPA627 is increased by the gain of the OPA603. The resulting slew rate is 730V/ $\mu$ s and settling time to 0.01% is 520ns. Care must be taken when selecting the feedback resistor for the OPA603. Excessive phase shift through that amplifier will cause instability. This composite amplifier takes advantage of the more precise OPA627 (voltage feedback amplifier) and the faster slewing OPA603 (current feedback amplifier).



## HIGH SPEED DIFFERENCE AMPLIFIER

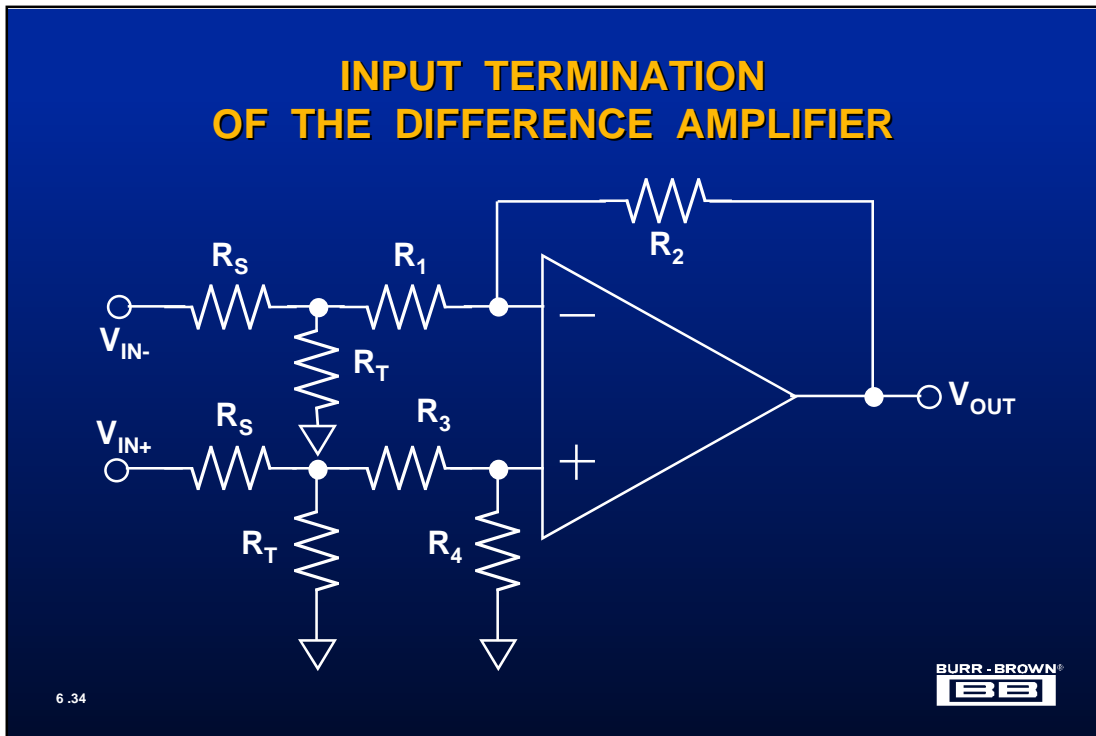


6.33

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At first pass, one would say the voltage feedback amplifier would be the only amplifier type appropriate for the difference amplifier configuration. First impressions may indicate that the DC errors of the bias currents of the current feedback amplifier would cause significant limitations. Actually, that may not be the overriding concern for this high speed application where the resistors in the circuit are low value and DC errors are typically ac coupled out of the signal. This circuit will work with both the voltage feedback amplifier and current feedback amplifier, with one major limitation; input signal termination when either amplifier is used. In both cases, the input resistance looking into the difference amplifier is equal to  $R_1$  for the inverting input and  $R_3 + R_4$  for the non-inverting input.  $R_1$  should equal  $R_3 + R_4$  in order to match input loads when true independent sources are to be differentiated. In the case of the current feedback amplifier, a further restriction is placed on the values of the resistors in the circuit.  $R_2$  must be equal to the manufacturer's recommended resistor value to insure good dynamic performance. Although the voltage feedback amplifier feedback resistor is not as restricted in value as the current feedback amplifier, proper termination would require that the feedback resistor be  $50\Omega$  assuming a gain of one. Depending on the amplifier output drive capability and the output load, this may or may not be an issue.

## INPUT TERMINATION OF THE DIFFERENCE AMPLIFIER



A resistor network can be added in order to overcome the problems with low impedance sources and circuit matching. In this design exercise, the source resistance,  $R_S$  is usually known and set prior to the design of the difference amplifier. Additionally,  $R_2$ , for the current feedback amplifier is suggested by the manufacturer in order to achieve optimum bandwidth performance. In order to properly terminate the source resistance,  $R_S$ , should be equal to the parallel combination of  $R_1$  and  $R_T$ . This termination requirement attenuates the signal by 50%. Consequently, the gain of the inverting signal should then be  $+2V/V$ . This is achieved by making  $R_2$  twice that of  $R_1$ .

The non-inverting input is terminated in a similar fashion, causing a 50% attenuation of the signal. This is made possible by setting the source resistance equal to the parallel combination of  $R_T$  and  $(R_3 + R_4)$ . Additionally,  $(R_3 + R_4)$  is set to equal  $R_1$  in order make the termination resistors,  $R_T$ , equivalent for the inverting as well as the non-inverting inputs. Further attenuation of the signal is done with the voltage divider formed by  $R_3$  and  $R_4$ . The non-inverting signal is then gained by  $(1 + R_2 / R_{IN})$ , where  $R_{IN}$  is the impedance looking from the inverting input of the amplifier back towards the source,  $R_1 + R_T \parallel R_S$ . The complete set of design equations are listed below:

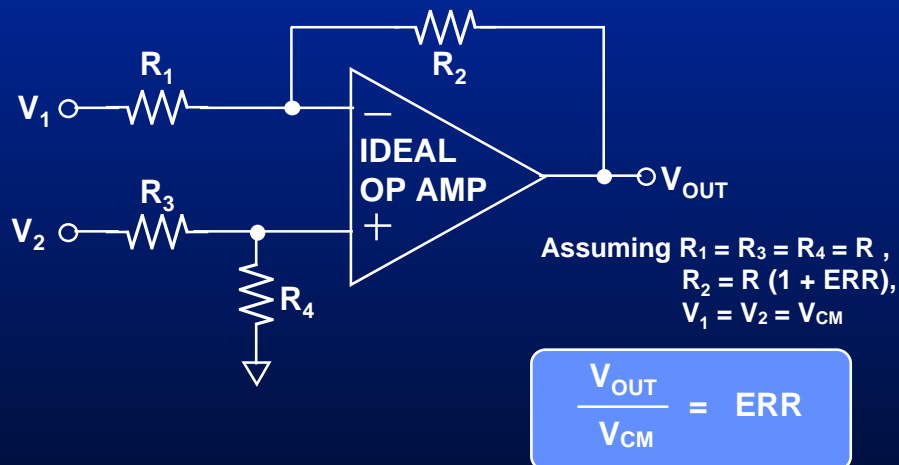
$$R_S = R_1 \parallel R_T$$

$$R_1 = R_2 / 2$$

$$R_2 / (R_1 \parallel (R_T + R_S)) + 1 = (R_3 + R_4) / 2 R_4$$

$$(R_3 + R_4) = R_1$$

## COMMON-MODE REJECTION OF DIFFERENCE AMPLIFIER

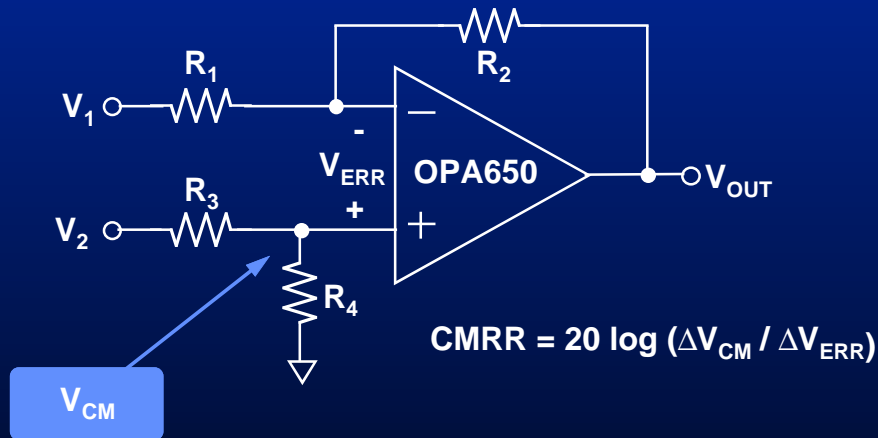


The common-mode rejection of the difference amplifier is equally dominated by the common-mode rejection of the amplifier and the common-mode rejection of the network around the amplifier. For precision circuits the common-mode rejection of the amplifier is usually ignored because precision amplifiers typically have CMRR performance in the 90 to 120 dB range. Evaluation of the resistive network reveals that the discrete designs are limited in their common-mode rejection performance by the resistive network. Given the assumptions above, the results of the calculations of common-mode rejection ratio of the network is in the table below:

ERR	CMRR
<b>0.001%</b>	<b>100dB</b>
<b>0.0032%</b>	<b>90dB</b>
<b>0.01%</b>	<b>80dB</b>
<b>1%</b>	<b>40dB</b>

With high speed amplifiers, CMRR figures range in the 50dB to 60dB performance area. Consequently, it is necessary to consider the amplifier performance as well as the resistor performance. This evaluation also holds true for  $R_1 = R_2$  and  $R_3 = R_4$ , where  $R_1$  does not necessarily equal  $R_3$ .

## COMMON-MODE REJECTION WITH A VOLTAGE FEEDBACK AMPLIFIER

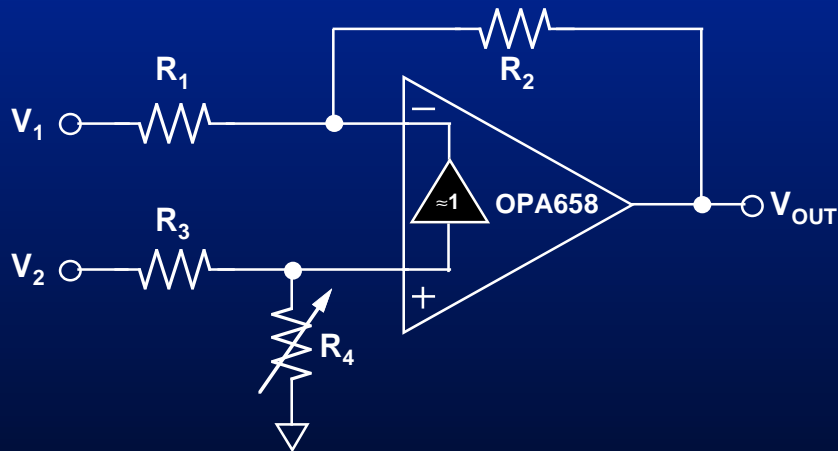


6.36



For the voltage feedback amplifier, the common-mode rejection of the device is determined by the changes in the error voltage,  $V_{ERR}$ , with changes in the common-mode voltage at the input of the amplifier,  $V_{CM}$ . Very little can be done about this limitation of the voltage feedback amplifier in the difference amplifier configuration. If the CMRR of the amplifier is the limiting portion of the circuit, usually, another voltage feedback amplifier, with higher CMRR performance is selected.

## DIFFERENCE AMPLIFIER WITH CURRENT FEEDBACK AMPLIFIER



6.37



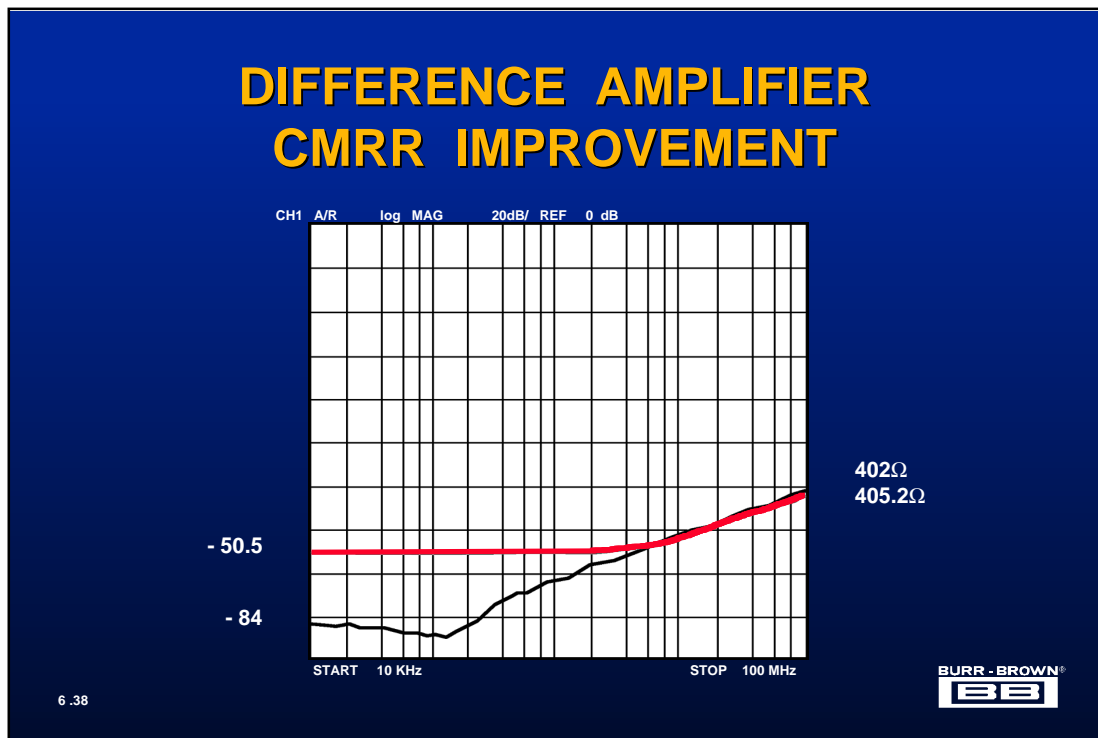
As with the voltage feedback amplifier, the current feedback amplifier can dominate the CMRR performance of the difference amplifier. If all four resistors are perfectly matched, the common-mode rejection of the difference amplifier is usually limited to 50 to 60dB, depending on the amplifier selected. The current feedback amplifier differs from the voltage feedback amplifier in that the common-mode rejection of this circuit can be substantially improved at lower frequencies. This is achieved by adjusting the gain of the non-inverting signal in such a way to accommodate for the gain error,  $\alpha$ , in the buffer portion of the input stage of the current feedback amplifier.

The transfer function of the difference amplifier, with the buffer gain included in the evaluation is:

$$V_{OUT} / V_{CM} = \alpha R_4(1+R_2/R_1) / ((R_3 + R_4) - R_2/R_1)$$

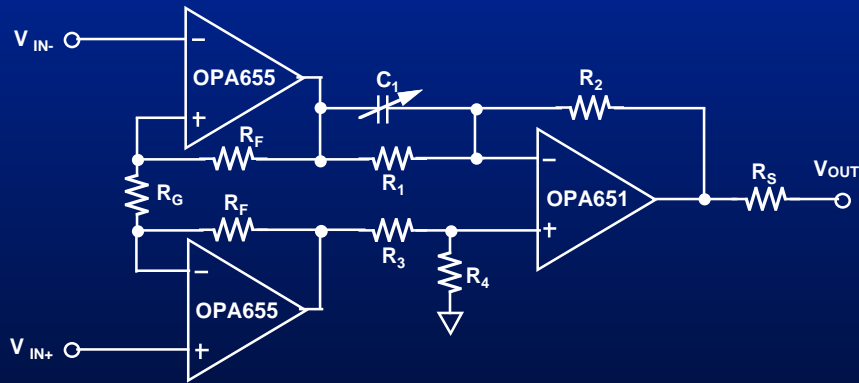
As the calculation illustrates,  $R_4$  can most effectively counter the gain error of alpha. If alpha is equal to one, the common-mode rejection is dependent on the resistive network. Typical common-mode rejection ratio specifications for the current feedback amplifier, OPA658, measure at 50dB, given  $R_1 / R_2 = R_3 / R_4$ . In theory, it is possible to adjust the common-mode rejection of the difference amplifier that uses a current feedback amplifier to 20 to 30dB better, by adjusting the  $R_4$  resistor.

## DIFFERENCE AMPLIFIER CMRR IMPROVEMENT



In order to prove the theory, the OPA658 current feedback amplifier was used with a matched resistive network or  $R_1$  through  $R_4$  to be  $402\Omega$ . The common-mode rejection performance of the amplifier is measured and shown in the top most curve on the diagram above.  $R_4$  is then adjusted to maximize the common-mode rejection performance. Note that the low frequency CMRR is improved while high frequency rejection is unaffected by the variance in  $R_4$ . In this test example, the adjusted resistor,  $R_4$ , is changed to be  $405.2\Omega$ . This adjusted resistance in the circuit counteracts the attenuation in the input stage caused by the input buffer gain,  $\alpha$ .

## HIGH SPEED INSTRUMENTATION AMPLIFIER



By-pass capacitors not shown

6.39



The difference amplifier's low input impedance can be a difficult problem to design around. A solution to impedance matching problems in high speed circuits is to implement a high speed instrumentation amplifier using a three op amp topology. This is achieved with OPA655s as input amplifiers and the OPA651 for the difference amplifier. The DC gain equation for this topology is  $(1 + 2R_F / R_G)$ . The common-mode rejection capability of the circuit is dominated by the common-mode rejection of the difference amplifier which is usually dominated by the resistors,  $R_1$  through  $R_4$ . At high frequencies, the roll off of this circuit is affected by the two input amplifiers as well as the output amplifiers. As the gain increases on the input stage by adjusting  $R_G$ , the circuit bandwidth limitation is dominated by the input amplifiers, OPA655. All amplifiers in this circuit are voltage feedback amplifiers. The linear common-mode range of the input op amps range is approximately  $\pm 2.4V$  with  $\pm 5V$  supplies. As the output voltage increases, the linear input range will be limited by the output voltage swing of the input amplifiers.

This circuit was configured in a gain of  $+2V/V$ . The component values for this circuit are listed below:

$$R_1 = 300\Omega$$

$$R_2 = 300\Omega$$

$$R_3 = 150\Omega$$

$$R_4 = 150\Omega$$

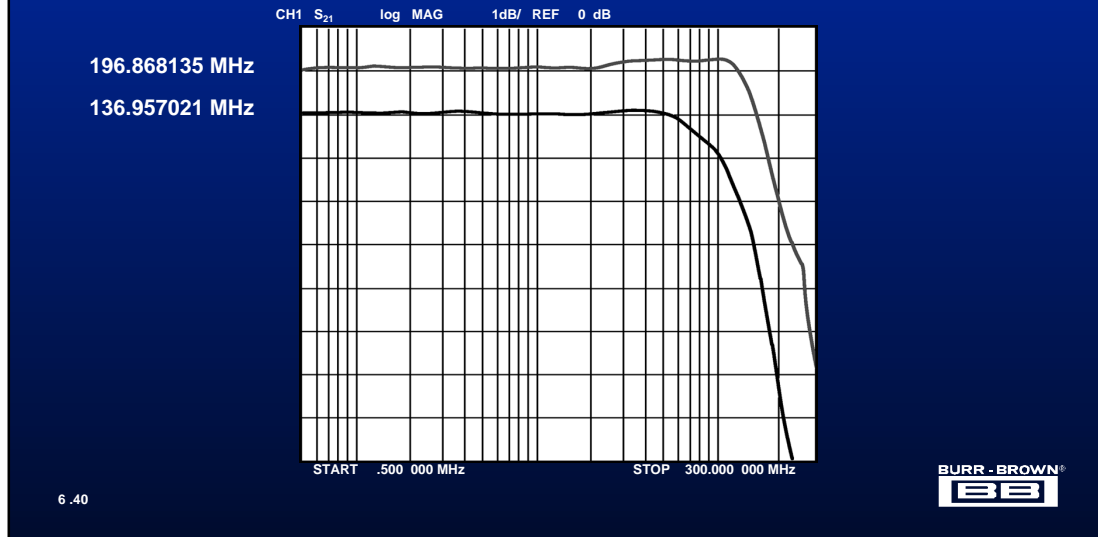
$$R_F = 100\Omega$$

$$R_G = 200\Omega$$

$$R_S = 50\Omega$$

$$C_1 = 1.5 \text{ to } 6\text{pF}$$

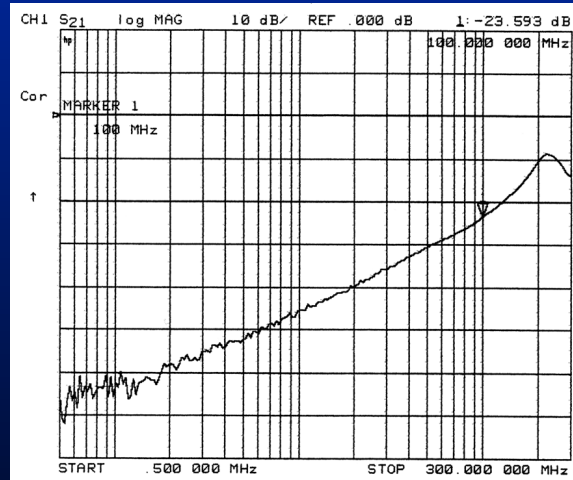
## HIGH SPEED IA FREQUENCY RESPONSE



The frequency response of this circuit is illustrated above. The bottom most plot represents the performance of the circuit in a gain of +1.5V/V, where  $R_G$  is left open. The top most plot represents the performance of the circuit in a gain of +3.5V/V. The change in the bandwidth of the circuit is a result of the effects of the gain bandwidth product of the OPA655.



# HIGH SPEED IA COMMON-MODE REJECTION RATIO

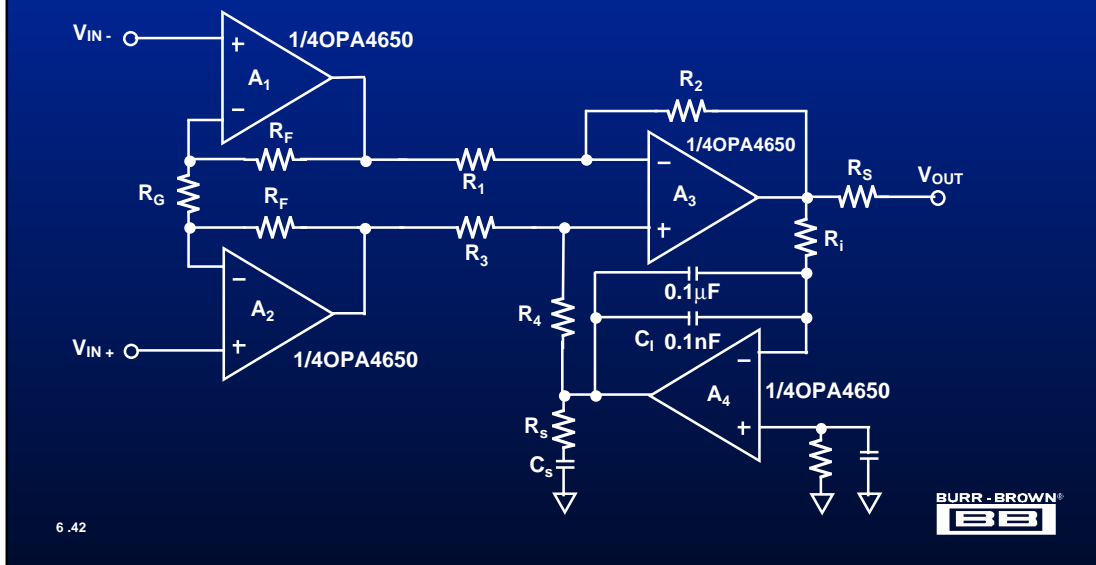


6.41



The common-mode rejection of the instrumentation amplifier in a gain of 0dB is shown in this plot. At 100MHz the common-mode rejection is measured to be -23dB.

## INSTRUMENTATION AMPLIFIER WITH DC - CORRECTION



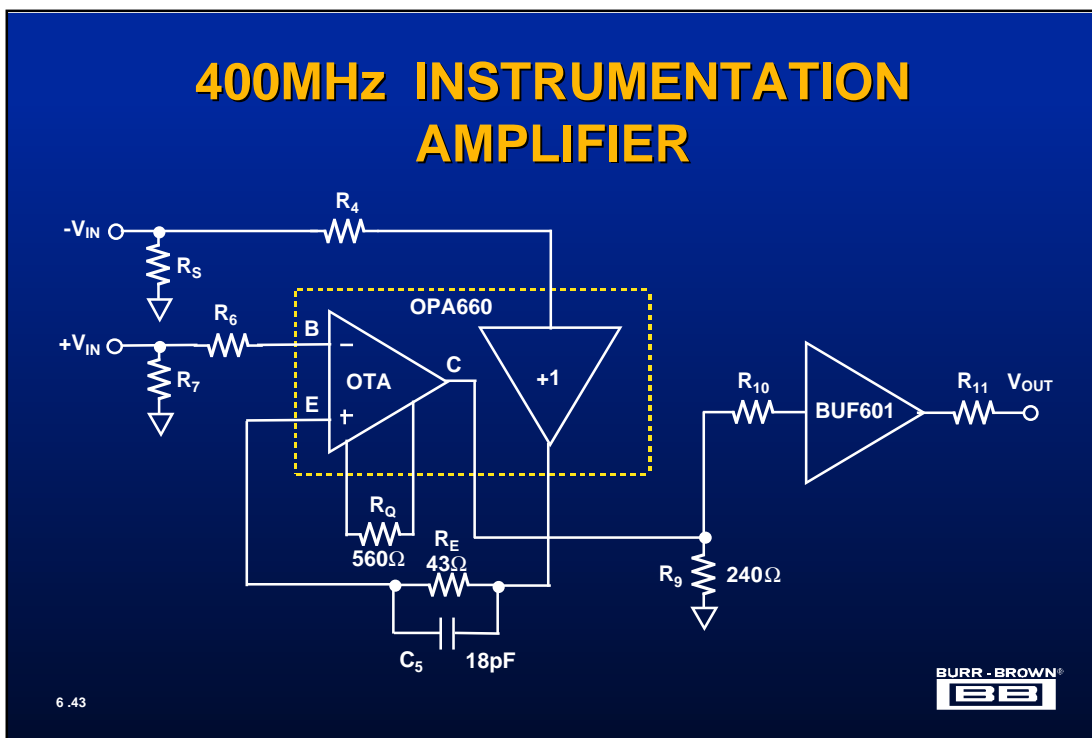
This circuit shows the high speed voltage feedback amplifier in a standard three op amp instrumentation amplifier configuration with a DC-correction feedback circuitry. In a gain of two, the instrumentation amplifier has a bandwidth of approximately 120MHz. The gain is set by the resistors,  $R_F$  and  $R_G$ , where the gain of the circuit is equal to  $1 + 2 R_F / R_G$ . Amplifier  $A_3$  is configured as a difference amplifier with all four resistors ( $R_1 - R_4$ ) equal to  $402\Omega$ . The optimal feedback resistor value for the op amp, OPA4650, illustrated in this figure is  $402\Omega$ .  $R_1 - R_4$  resistors should be well matched to achieve a good common-mode rejection.

Wide band op amps are known to have only moderate DC specifications, like input bias currents in the  $\mu A$  range, or an offset voltage of several mV. The maximum input offset voltage of the OPA4650 is  $\pm 5.5mV$ . To increase the DC accuracy of this circuit the remaining fourth amplifier of the OPA4650 quad amplifier ( $A_4$ ) is used for a DC correction loop. Configured as an integrator, the amplifier  $A_4$  senses the DC component at the output of  $A_2$  and corrects it with a voltage of opposite polarity into the non-inverting node of  $A_3$ . The time constant of the integrator is set by  $R_1$  and  $C_1$ , with a transfer function of  $V_{OUT} = 1/R_1 C_1 \int (V_{IN} dt)$ . If higher values of integration capacitance is used it is recommended that another capacitor of smaller values, like  $0.1\mu F$  be put in parallel. This will counteract the inductive behavior of the  $0.1nF$  capacitor at higher frequencies.

One could argue that the integrator could be built using a low-speed, precision op amp. This may not be a good choice in terms of good common mode rejection over frequency. The 'cold' side of the resistor  $R_4$  should see a constant low impedance throughout the entire frequency range. The output impedance of the OPA4650 is only  $0.08\Omega$  at  $0.1MHz$ . In case the output impedance of the amplifier shows too much inductive behavior, an RC series circuit can be placed from the output of  $A_4$  to ground.

To compensate for the error voltage caused by  $I_B \cdot R_1$ , a resistor of similar value can be placed at the non-inverting input of  $A_4$ . The disadvantage is that it adds more voltage noise to the circuit, however, a capacitor in parallel will shunt most of this noise.

## 400MHz INSTRUMENTATION AMPLIFIER

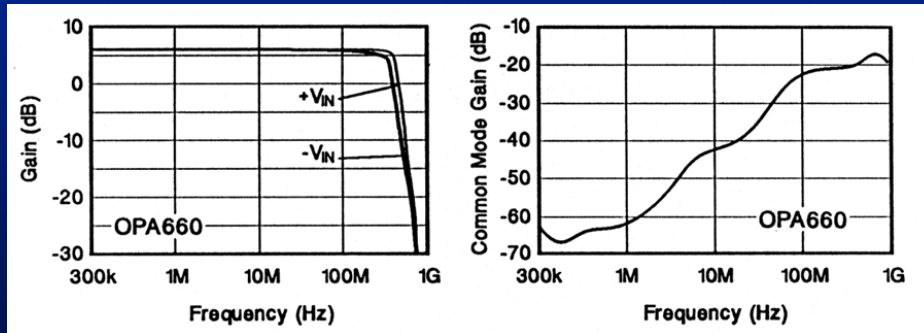


An example of the versatility of the transconductance amplifier, OPA660 is shown here. The OTA segment of the OPA660 operates similar to a transistor. The B input is high impedance, the E input is low impedance and follows the voltage changes of the B input. One differentiation between the OTA and an ideal transistor is that the B and E inputs are nearly the same voltage, with a small offset difference. When the E input sinks or sources current the C output also sinks or sources current. This 400MHz differential amplifier can be difficult to design, involving extensive and complex hardware. The OPA660 makes it easy with stable operation and -60dB of common-mode rejection at 1MHz. The OPA660 is configured in an open-loop structure (no feedback loop) with two identical, high-impedance inputs. When a differential voltage is applied at the input, current flows through  $R_E$ . That current is mirrored to the high impedance collector output (C) of the OTA producing a corresponding output voltage across  $R_9$ . The OTA is set for a gain of +4 according to the gain equation:

$$G = \frac{R_9}{R_E + 2/g_m}$$

where  $g_m$  is the transconductance of the OTA. The  $g_m$  depends on the quiescent current which is controlled by  $R_Q$ .  $R_Q$  is chosen for a +/-20mA quiescent current making the  $g_m = 120\text{mA/V}$ . The buffer amplifier, BUF601, is used to drive low-impedance loads and to decouple the output of the OTA from the load. Both inputs and the output are terminated for 50Ω systems. This can be adapted to other termination requirements by replacing  $R_3$ ,  $R_7$ , and  $R_{11}$ . The resistors  $R_4$ ,  $R_6$  and  $R_{10}$  are placed in series with the high-impedance inputs in order to reduce gain peaking. Capacitor  $C_5$  in parallel with  $R_E$  compensates the parasitic capacitance present at the OTA's collector (C), consequently expanding the circuits achievable bandwidth.

## 400MHz INSTRUMENTATION AMPLIFIER PERFORMANCE

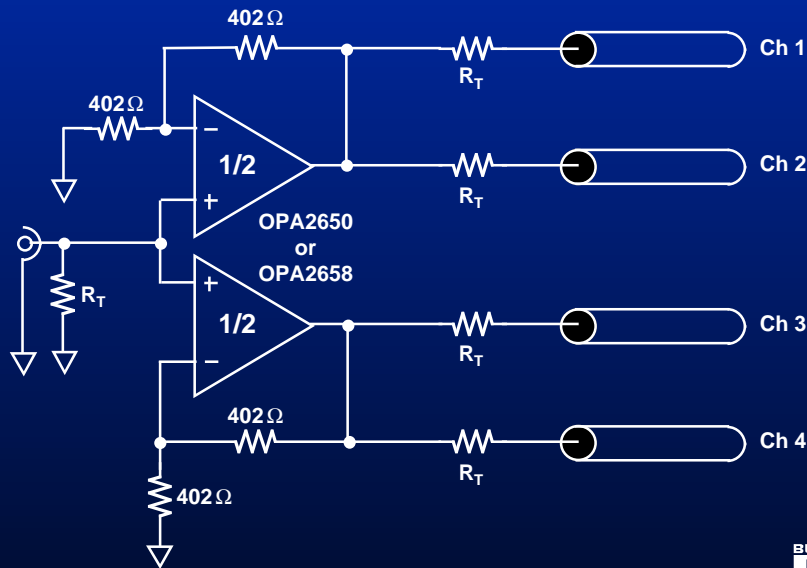


6.44



As shown the bandwidth of this circuit is approximate 400MHz. Without the capacitor  $C_5$ , the -3dB frequency roll-off of the circuit would be 85MHz. The common-mode gain vs frequency is shown in the second plot. The common-mode rejection is approximately -63dB at low frequencies and -20dB at 400MHz. In addition, the low noise voltage density of  $7.7\text{nV}\sqrt{\text{Hz}}$  makes it possible to process very small signals.

## DRIVING MULTIPLE VIDEO LOADS



6.45

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Video distribution amplifiers are used to drive the same video signal into multiple loads. In this application a dual high-speed amplifier, like the voltage feedback OPA2650 or the current feedback amplifier OPA2658, can be used in this circuit. The standard video amplitude is 1V peak-to-peak, with the video portion of the signal swinging between 0V and +0.7V and the synchronization portion of the signal swinging between -0.3V and 0V. Consequently, most video driver applications require larger positive output voltage swing capability.

Considering a complementary transistor (NPN and PNP) push-pull output stage built on a high-speed process, the PNP transistors are known to be the weak part when comparing the ability to sink and source current. This means that high-speed op amps tend to have higher output currents for sourcing than for sinking. This performance characteristic is used to an advantage with video driver applications. The OPA2650 (dual voltage feedback amplifier) has sourcing current capability of +75mA and sinking capability of -65mA at room temperature. The OPA2658 (dual current feedback amplifier) can source +90mA and sink -60mA at room temperature.

## COMPARISON OF CFB vs VFB

	CFB	VFB
Closed-Loop Bandwidth	Relatively unaffected by closed-loop gain	Varies according to GBW relation
Feedback Network	AC performance set by resistor values	AC performance independent of resistor values
Slew Rate	Set by input edge rate	Fixed internally
Dominant Noise Source	Inverting input current noise	Input voltage noise
Input Impedance	Asymmetrical	Symmetrical
Bias Current	DC errors difficult to cancel	DC errors canceled by matched input impedance

To summarize, the closed-loop bandwidth of the current feedback amplifier is relatively unaffected by the closed-loop gain as long as  $R_F$  is not changed significantly. There is an input error resistance,  $R_S$ , of the low impedance inverting input consequently requiring some adjustments of  $R_F$  to obtain near ideal results. On the other hand, the closed-loop bandwidth when using a voltage feedback amplifier varies with gain approximately 20dB/decade.

From another perspective, the current feedback amplifier's ac performance is set by the feedback resistor value. It is possible to restrict the bandwidth of the circuit by using a higher  $R_F$  which would improve the dynamic response without adjusting the gain. With the voltage feedback amplifier in the circuit, the dynamic response is dependent on the ratio of the input resistor,  $R_{IN}$ , and the feedback resistor,  $R_F$ , and independent of the values.

Typically the slew rate of the current feedback amplifier is better than the voltage feedback amplifier. With a qualifier, the slew rate of the current feedback amplifier is set by the input edge rate. In the case of the voltage feedback amplifier, the slew rate is set internally and independent of the input signal edge rate.

The dominant noise source for the current feedback amplifier is the inverting input current. With the voltage feedback amplifier, current noise is very low and voltage noise dominates the performance.

The errors due to input impedance and input bias current mismatch on the current feedback inputs go hand in hand. These mismatches can cause DC errors that can be difficult to reduce.

# PRODUCT SELECTION TREES

Voltage Feedback  
Stable for  $G = > 2$

GBWP = > 400MHz  
 OPA643 (SR = 1000V/ $\mu$ sec)  
 OPA621 (SR = 350V/ $\mu$ sec)  
 OPA675 (SR = 350V/ $\mu$ sec)  
 OPA676 (SR = 350V/ $\mu$ sec)  
 OPA651 (SR = 300V/ $\mu$ sec)

GBWP < 400MHz  
 OPA654 (SR = 750V/ $\mu$ sec)  
 OPA641 (SR = 650V/ $\mu$ sec)  
 OPA678 (SR = 350V/ $\mu$ sec)  
 OPA637 (SR = 100V/ $\mu$ sec)

Voltage Feedback  
Unity Gain Stable

GBWP = > 200MHz  
 OPA642 (SR = 380V/ $\mu$ sec)  
 OPA640 (SR = 350V/ $\mu$ sec)  
 OPA655 (SR = 300V/ $\mu$ sec)  
 OPA646 (SR = 180V/ $\mu$ sec)  
 OPA620 (SR = 175V/ $\mu$ sec)

GBWP < 200MHz  
 OPA628 (SR = 310V/ $\mu$ sec)  
 OPA650 (SR = 240V/ $\mu$ sec)  
 OPA2650 (SR = 240V/ $\mu$ sec, dual)  
 OPA4650 (SR = 240V/ $\mu$ sec, quad)  
 OPA671 (SR = 100V/ $\mu$ sec)

Current Feedback  
SR > 1000V/ $\mu$ sec  
(BW @  $G = +2$ )

OPA658 (650MHz)  
 OPA648 (600MHz)  
 OPA2658 (500MHz, dual)  
 OPA4658 (450MHz, quad)  
 OPA644 (300MHz)  
 OPA623 (290MHz)  
 OPA603 (160MHz)

6.47



High speed amplifiers can be found with voltage feedback and current feedback topologies with a variety of bandwidths and slew rates. The tables above summarize the Burr-Brown product offering. The voltage feedback amplifiers are separated into two general categories, the first for amplifiers that are stable in a gain of +2 or high and the second for amplifiers capable of unity gain stability. Each of the two voltage feedback categories are further subdivided into gain bandwidth product groupings. This separation along with the slew rate capability of the amplifier is useful for first order product selection.

The current feedback amplifiers are listed in the third column and ordered according to small signal bandwidth in a closed-loop gain of +2V/V. Slew rate is not specified because of its strong dependency on the input signal characteristics.

## PRODUCT SELECTION GUIDE

Circuit Requirements	CFB	VFB
Fast Slew	Yes	Maybe
Wideband	Yes	Maybe
High Gain	Yes	Maybe
Low DC Errors	No	Yes
Low Distortion	Yes	Maybe
Low Noise Transimpedance	No	Yes
Ease of Design	Yes !	Maybe

6.48



This table briefly summarizes some general rules of thumb for the selection of the proper amplifier for the application. If the circuit requires a fast slewing amplifier, particularly for large signals the current feedback amplifier will typically slew faster than the voltage feedback amplifier. However, recently, voltage feedback amplifiers have been introduced with quite good slew performance and very good bandwidth.

If the application requires an amplifier that has high closed-loop gain, the voltage feedback amplifier would be a more appropriate amplifier for the socket. Current feedback amplifiers are optimized for one gain, typically  $+2V/V$ . It is possible to use the amplifiers in higher gains at the expense of losing gain accuracy.

Current feedback amplifiers, typically, have lower harmonic distortion across the closed-loop bandwidth. Applications, such as video, require good dynamic performance making the current feedback amplifier many times the preferred amplifier.