

Current Feedback Op Amp Applications Circuit Guide

National Semiconductor
 OA-07
 David Potson
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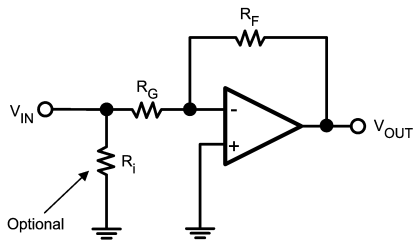


Introduction

No two high-speed applications are the same – or at least it seems that way. Nonetheless, while every system has its particular requirements, many of the design techniques are common among different designs. This application note illustrates design techniques utilizing current-feedback op amps and the practical circuits where they are used. The circuits should work well with any Comlinear op amp if appropriate adjustments are made for different feedback resistance values.

Inverting Gain

As with voltage-feedback op amps, the ratio of the feedback resistor to the gain-setting resistor determines the voltage gain in current-feedback op amp circuits. With current feedback, however, dynamic performance is largely independent of the voltage gain. (See application note AN300-1 for a technical discussion of current-feedback.) Also, the optimum feedback resistor value for a current-feedback op amp is indicated in the datasheet.

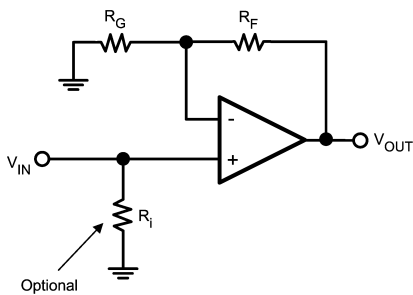


$$V_{OUT} = \frac{R_F}{R_G} V_{IN}$$

For input impedance of 50Ω,
 select $R_I \parallel R_G$ equal to 50Ω

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Non-Inverting Gain



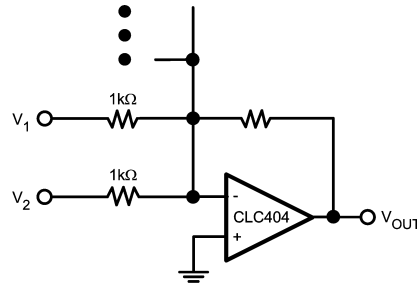
$$V_{OUT} = \left(1 + \frac{R_F}{R_G}\right) V_{IN}$$

R_I set the input impedance.

01278102

Summing Amplifier

Current-feedback op amps are the natural choice in summing applications since the bandwidth and other key specs are relatively unaffected by high gain setting. (The parallel combination of all the input resistors yields a small effective gain-setting resistance and hence a large effective gain setting.)

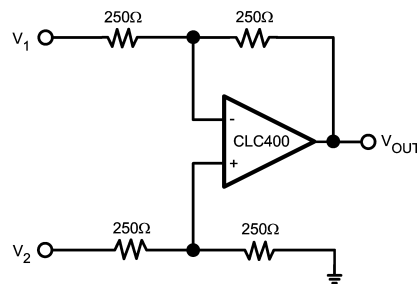


$$V_{OUT} = -(V_1 + V_2 \dots)$$

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Differential Amplifier

Be sure to obey common-mode input voltage limits shown in the op amp datasheet. If large, saturating input signals are expected, use an overdrive-protected op amp and appropriate protection circuitry.

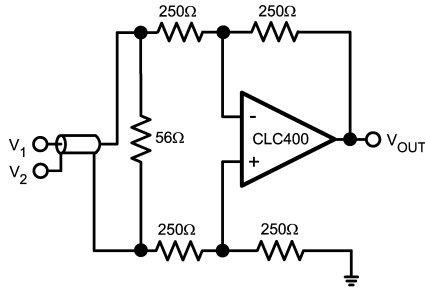


$$V_{OUT} = (V_2 - V_1)$$

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Differential Line Receiver

This circuit provides good common mode rejection and 50Ω termination for signals which need to be transmitted through coaxial lines.



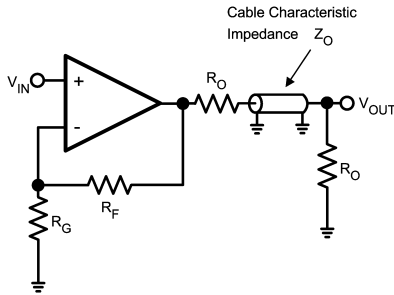
$$V_{OUT} = (V_2 - V_1)$$

Differential input resistance is 50Ω

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Coaxial Cable Driver

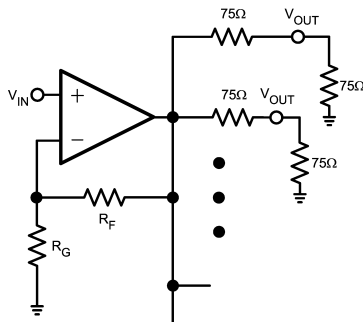
Proper transmission line driving techniques are important when high-speed signals have to travel more than a few inches. The back-matching and terminating resistor, R_O , are chosen to match the characteristic impedance of the coaxial line. If the load is well-matched to the transmission line impedance, the back-matching resistor may be omitted for greater voltage swing. (Remember that back matching creates a voltage divider which attenuates the output signal by 50%.)



$$R_O = Z_O$$

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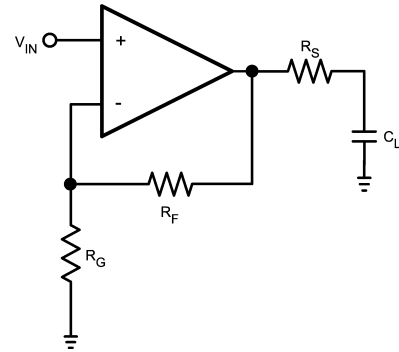
Distribution Amplifier



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Driving Capacitive Loads

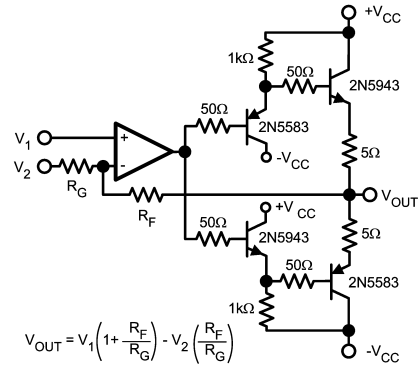
The damping resistor, R_S , reduces pulse-response overshoot and frequency-response peaking caused by the load capacitance. The value of R_S may be found on some of the op amp datasheet or may be found experimentally.



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Output Current Booster

This circuit provides up to 400mA of output current. Since the output buffer circuit introduces additional phase lag, the feedback resistor, R_F , may have to be increased above the datasheet recommendation to decrease loop gain and thus improve stability. The gain-setting resistor, R_G , is then chosen for the desired gain.

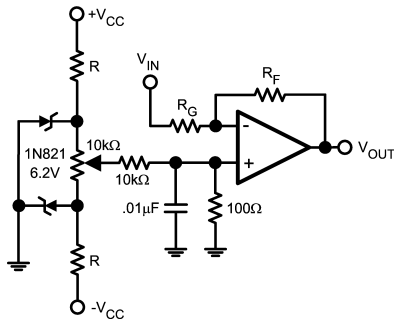


$$V_{OUT} = V_1 \left(1 + \frac{R_F}{R_G} \right) - V_2 \left(\frac{R_F}{R_G} \right)$$

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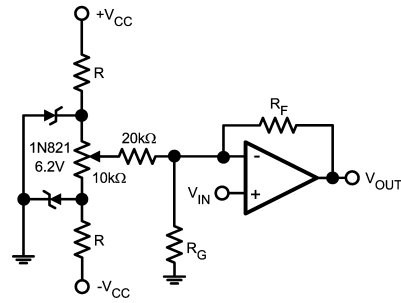
Simple Offset Adjustment

The Zener diode biasing resistor, R, should be chosen to provide a diode current of 7.5mA.



Inverting

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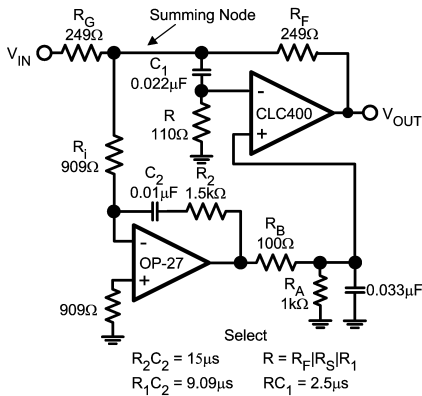


Non-Inverting

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Composite Amplifier for Low Offset and Drift (#1)

This composite circuit provides both high speed and good DC performance and unlike most composite circuits, it provides good settling performance (17ns to 0.1%). In operation, the OP-27 op amp drives its output such that the summing node is driven to 0V (which is the normal case for an inverting gain circuits). Thus, the circuit output takes on the high performance DC characteristics of the OP-27. At high frequencies, the high-speed op amp takes over to provide good AC performance.

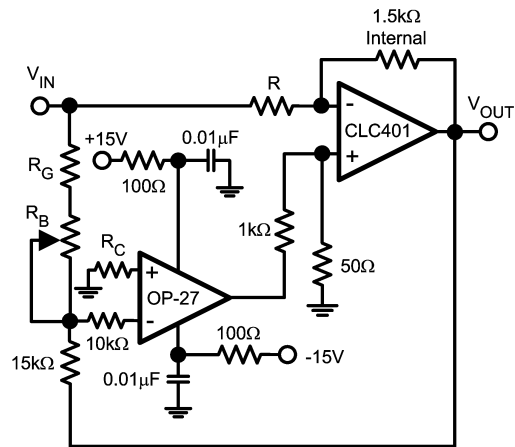


$$R_2 C_2 = 15 \mu s \quad R = R_F | R_G | R_1$$

$$R_1 C_2 = 9.09 \mu s \quad R C_1 = 2.5 \mu s$$

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input and output. R_B is adjusted for minimum output voltage at the OP-27 when V_{OUT} is a 70kHz square wave of $10V_{PP}$ centered at 0V.



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$$V_{OUT} = - \left(\frac{1500}{R} \right) V_{IN}$$

$$R_A \cong 9.5R$$

$$R_B \cong 0.5R$$

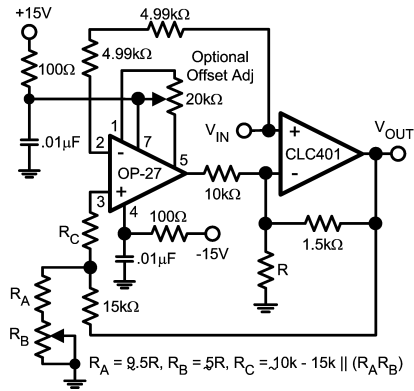
$$R_C \cong 10K - 15K \parallel (R_A + R_B)$$

Composite Amplifier for Low Offset and Drift (#2)

This composite circuit is useful with those products which have the feedback resistor connected internally to both the

Non-Inverting Composite Amplifier

As with the previous circuit, R_B is chosen for minimum output voltage at the OP-27 when V_{OUT} is a 70kHz square wave of $10V_{PP}$ center at 0V.

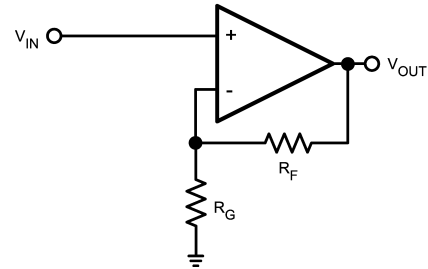


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Reducing Bandwidth

Bandwidth and loop stability is controlled by R_F . Increasing R_F reduces bandwidth according to the approximate relationship:

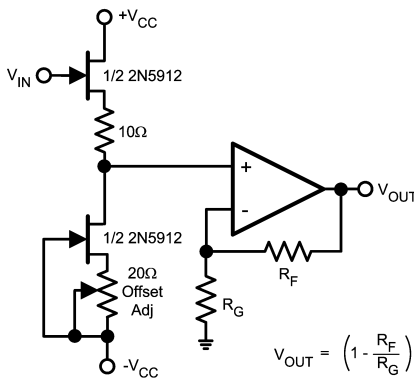
$$\frac{BW_2}{BW_1} \approx \frac{R_{F1}}{R_{F2}}$$



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FET-Input Circuit

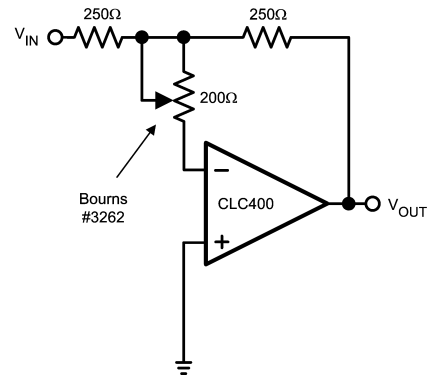
A FET-input circuit is useful when a greater input impedance is desired or when bias currents or noise current need to be reduced.



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Adjustable Bandwidth

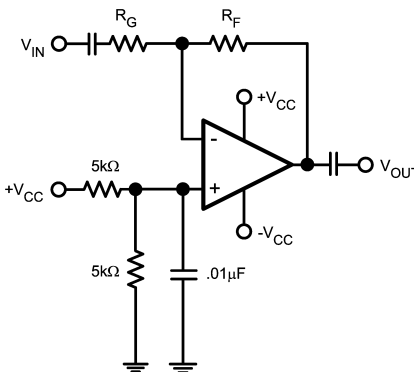
By increasing the inverting input impedance (which is normally very low) of a current feedback op amp, the bandwidth of the op amp can be reduced. The bandwidth of the circuit below can be varied over a range of 60MHz to 160MHz.



01278119

AC-coupled Amplifier (with Single-Supply Biasing)

The voltage divider circuit at the non-inverting input biases the op amp input and output at the supply midpoint. For those op amps having a bias pin, these pins should also connect to the supply midpoint bias circuit.



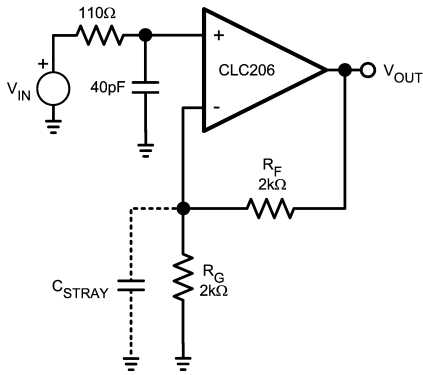
01278117

Reducing Frequency-Response Peaking

(due to stray capacitance in parallel with R_G)

The low-pass filter at the non-inverting input cancels the frequency-response zero caused by C_{STRAY} . At low non-inverting gains, the CLC231 or CLC400 will provide a flatter frequency response without the need for the low-pass filter (because they can be used with lower feedback resistor values).

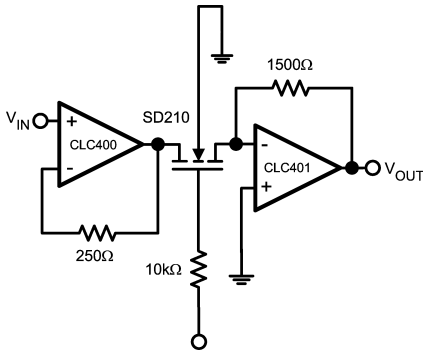
Reducing Frequency-Response Peaking (Continued)



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Adjustable Gain Using a FET

This circuit provides a 26dB adjustment range and a gain flatness of 1dB from DC to 50MHz. An SD210 FET provides low on-resistance with minimal capacitive loading.

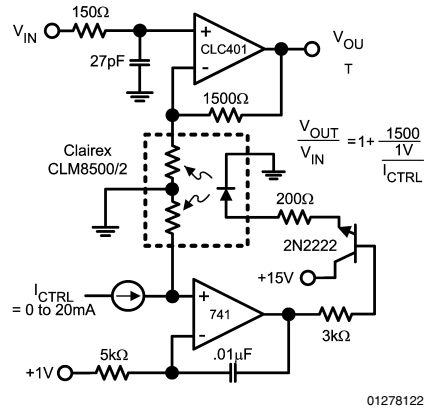


$V_{CONTROL} = 0 \text{ TO } +5V$

01278121

Adjustable Gain Using a Photoresistor

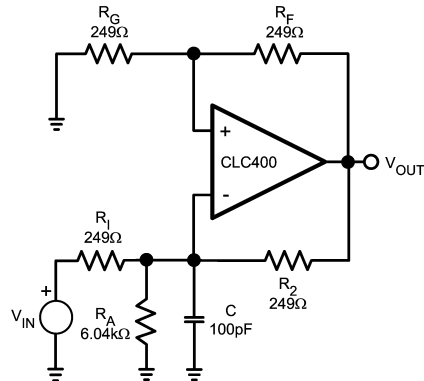
This circuit provides a 12dB adjustment range and a gain flatness of 1dB from DC to 20MHz. The 741 circuit improves temperature stability and repeatability of the photoresistor circuit.



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Integrator (#1)

With current-feedback op amps, it is important to keep large capacitance values out of the inverting feedback loop in order to maintain stability.



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For stable operation,

$$\frac{R_2}{R_1 \parallel R_A} \geq \frac{R_F}{R_G}$$

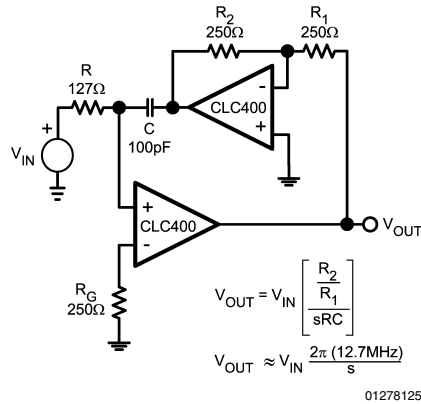
All resistors are 1%

$$V_{OUT} = V_{IN} \left[1 + \frac{R_F}{R_G} \right] \frac{1}{sR_1C}$$

$$V_{OUT} \sim V_{IN} \frac{2\pi (12.8\text{MHz})}{s}$$

Integrator (#2)

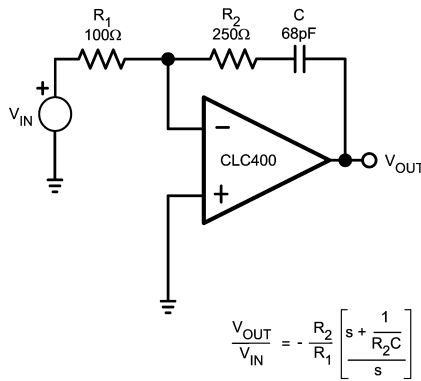
This integrator provides higher DC gain than #1. For values shown, the DC gain is 55dB. Higher values can be obtained by reducing R_G , however, the ratio of R_G to R_1 should remain constant for adequate loop stability. Much of the output noise is directly proportional to R_1 so that the higher DC gain is obtained at the expense of higher noise. This circuit does not have the stability problem that is related to resistor matching as does integrator #1.



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Integrator with Zero

In this circuit, feedback capacitance is acceptable because the op amp relies upon the value of the feedback resistor for stability.

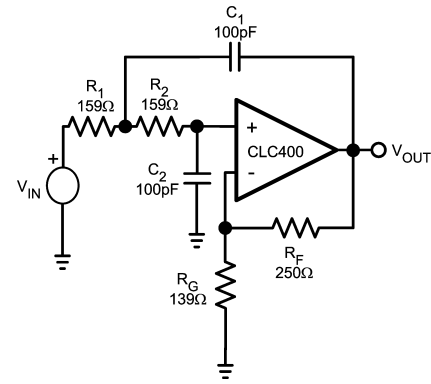


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Active Filter Circuits

The following five circuits illustrate how current feedback op amps provide high-performance active filter functions. The 'KRC' realization is used (see references at the end of the app note) since it does not require reactive elements in the (negative) feedback path, which would compromise stability. When the filter cutoff frequency is small relative to the amplifier bandwidth, the transfer functions shown will provide good accuracy. However, as with any active filter circuit, the group delay through the op amp becomes significant for cutoff frequencies greater than about 10% of the op amp bandwidth. For such designs, computer analysis tools and an iterative design approach is helpful.

Low-Pass Filter (10MHz, Q = 5)



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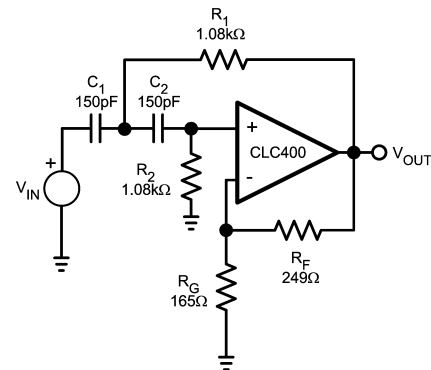
$$\frac{V_{OUT}}{V_{IN}} = \frac{s \frac{K_0}{R_1 C_1}}{s^2 + s \left(\frac{1}{R_1 C_1} + \frac{1}{R_3 C_2} + \frac{1}{R_3 C_1} + \frac{1 - K_0}{R_2 C_1} \right) + \frac{R_1 + R_2}{R_1 R_2 R_3 C_1 C_2}}$$

$$K_0 = 1 + \frac{R_F}{R_G} \quad \omega_0^2 = \frac{1}{R_1 R_2 C_1 C_2} \quad \left| \quad R_1 = R_3 = R, \quad C_1 = C_2 = C: \right.$$

$$Q = \frac{1}{\sqrt{\frac{R_1 C_2}{R_2 C_1} + \sqrt{\frac{R_2 C_2}{R_1 C_1} + (1 - K_0) \sqrt{\frac{R_1 C_1}{R_2 C_2}}}} \quad \omega_0 = \frac{\sqrt{2}}{RC}$$

$$Q = \frac{\sqrt{2}}{4 - K_0}$$

High-Pass Filter (1MHz, Q = 2)



01278129

$$\frac{V_{OUT}}{V_{IN}} = \frac{K_0 s^2}{s^2 + s \left(\frac{1}{R_2} \left(\frac{1}{C_1} + \frac{1}{C_2} \right) + \frac{1 - K_0}{R_1 C_1} \right) + \frac{R_1 + R_2 C_1 C_2}{R_1 R_2 C_1 C_2}}$$

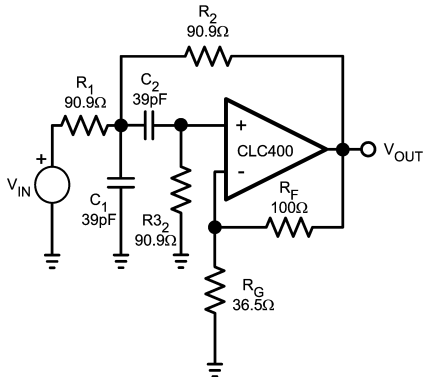
$$K_0 = 1 + \frac{R_F}{R_G} \quad \omega_0^2 = \frac{1}{R_1 R_2 C_1 C_2} \quad \left| \quad R_1 = R_3 = R, \quad C_1 = C_2 = C: \right.$$

$$Q = \frac{1}{\sqrt{\frac{R_1}{R_2} \left(\frac{C_2}{C_1} + \sqrt{\frac{C_1}{C_2}} \right) + \sqrt{\frac{R_2 C_2}{R_1 C_1} (1 - K_0)}}} \quad \omega_0 = \frac{1}{RC}$$

$$Q = \frac{1}{3 - K_0}$$

Band-Pass Filter (40MHz, Q = 4)

The component values shown are 'predistorted' from the nominal design values to account for the 1.6ns op amp group delay, which is significant relative to the filter cutoff frequency.



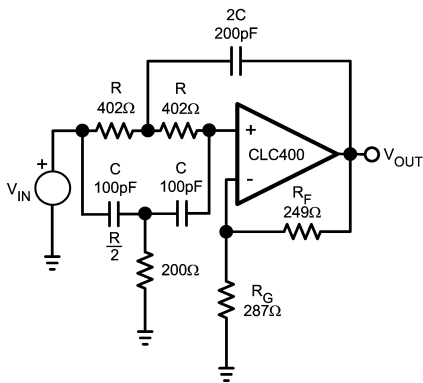
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$$\frac{V_{OUT}}{V_{IN}} = \frac{s K_0}{s^2 + s \left(\frac{1}{R_1 C_1} + \frac{1}{R_3 C_2} + \frac{1}{R_3 C_1} + \frac{1-K_0}{R_2 C_1} \right) + \frac{R_1 + R_2}{R_1 R_2 R_3 C_1 C_2}}$$

$$K_0 = 1 + \frac{R_F}{R_G} \quad \omega_0^2 = \frac{R_1 + R_2}{R_1 R_2 R_3 C_1 C_2} \quad \left| \begin{array}{l} R_1 = R_3 = R, \quad C_1 = C_2 = C: \\ \omega_0 = \frac{\sqrt{2}}{RC} \\ Q = \frac{\sqrt{2}}{4 - K_0} \end{array} \right.$$

$$Q = \frac{\sqrt{\frac{R_2 C_1 (R_1 + R_2)}{R_1 R_3 C_2}}}{1 + \frac{R_1}{R_2} + \frac{R_2}{R_3} \left(1 + \frac{C_1}{C_2} \right) - K_0}$$

Band-Stop Filter (4MHz, Q = 4)



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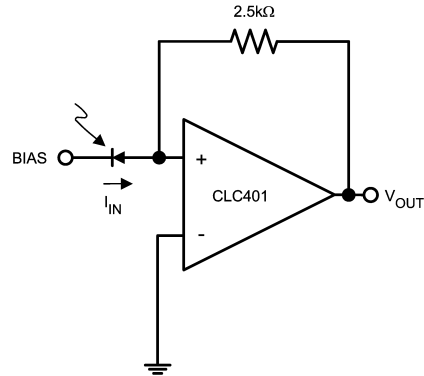
$$\frac{V_{OUT}}{V_{IN}} = \frac{K_0 \left(s^2 + \frac{1}{R^2 C^2} \right)}{s^2 + 2 \frac{s}{RC} (2 - K_0) + \frac{1}{R^2 C^2}}$$

$$K_0 = 1 + \frac{R_F}{R_G} \quad \omega_{0POLE}^2 = \omega_{0ZERO}^2 = \frac{1}{R^2 C^2}$$

$$Q_{POLE} = \frac{1}{2(2 - K_0)}$$

Photodiode Amplifier

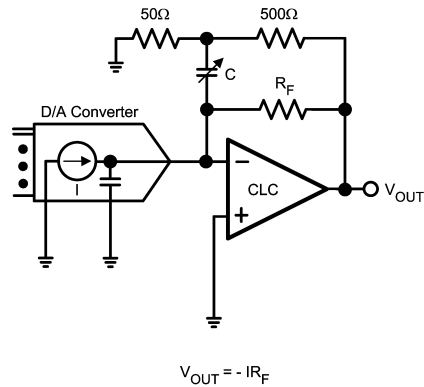
The circuit below provides a transimpedance gain of $-2.5k\Omega$ to convert the photodiode current into a voltage.



01278135

D/A Converted Buffer Amplifier

Most high-speed, current-output D/A converters provide the best performance when driving little or no load impedance. The circuit below meets this requirement while also providing a transimpedance gain which converts the D/A output current into a Voltage. The variable capacitor in the feedback loop should be adjusted for desired pulse response to compensate for the D/A output capacitance, which otherwise caused frequency-response peaking or instability. The 50Ω and 500Ω resistors reduce the effective value of the feedback capacitance so that a reasonable value capacitor may be used.



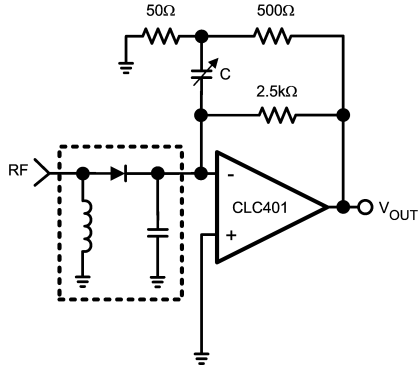
$$V_{OUT} = -I R_F$$

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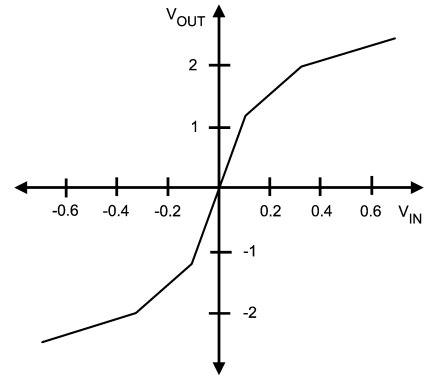
For example, with the CLC401, $R_F = 2.5k\Omega$ and a D/A output capacitance of 20pF, $C \cong 5pF$.

Tunnel Diode Detector Amplifier

See the D/A converter buffer circuit for circuit highlights.



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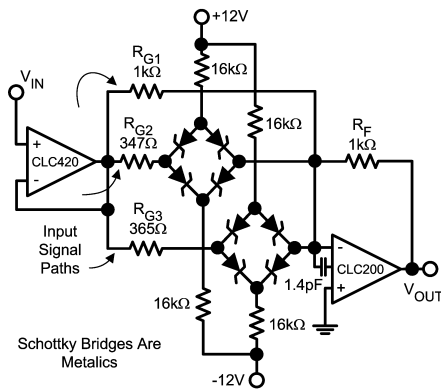
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Non-Linear Transfer Functions

Current-feedback op amps are particularly useful in non-linear transfer function circuits. Since bandwidth and other key specifications are independent of gain—the dynamic performance is relatively independent of signal level.

In analyzing the circuit, it is useful to identify the three input signal paths that contribute to the output voltage (the 1kΩ resistor and the two diode bridges). Each of these paths terminates at the inverting input—a point that is at virtual ground. Due to feedback, the current through the feedback resistor is equal to the sum of these input currents. The output voltage, therefore, is the product of the feedback resistor and the sum of the input current.

The individual input currents are equal to the input voltage divided by the respective gain-setting resistor. However, in the signal paths containing the bridges, the current follows this linear relationship until it limits at 12V/16kΩ. This is what leads to the non-linear gain. A more accurate analysis requires that the diode bulk and dynamic resistance be included.



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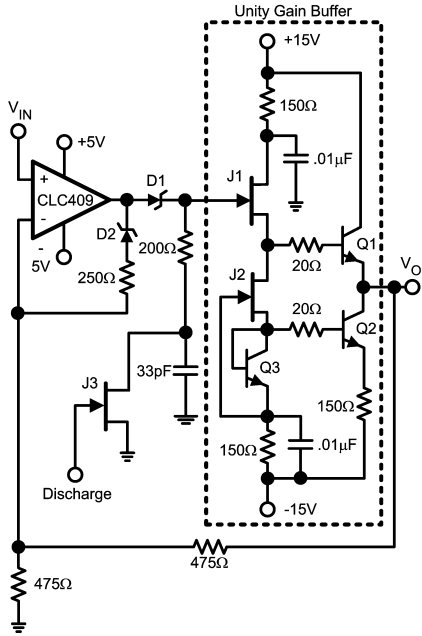
Peak Detector

The circuit shown in the next column can capture 50ns pulses over a 20mV to 1.5V input range. The circuit consists of three basic blocks: the op amp and diodes, the storage capacitor and discharge circuit, and the unity-gain buffer. The peak detecting action is caused by the conduction or non-conduction of the two diodes in the feedback loop.

When $V_{OUT} > V_{IN}$, the op amp output swings in the negative direction until D2 conducts and the feedback path is completed and the op amp does not saturate.

When $V_{OUT} < V_{IN}$, the op amp swings in the positive direction causing D1 to conduct and the storage capacitor voltage to be charge through the 200Ω isolating resistor. This action continues until $V_{OUT} > V_{IN}$ and equilibrium is established.

Peak Detector (Continued)



01278140

50ns pulses over a
20mV to 1.5V input range.

D1, 2 1N5711
Q1, 2, 3 MPSH10
J1, 2 2N5911
J3 2N4391

References:

A. Budak, *Passive and Active Network Analysis and Synthesis*, Houghton Mifflin Company, Boston, 1974.

IEEE Transactions on Circuits and Systems, Volume CAS-27, 'Optimum Configurations for Single Amplifier Biquadratic Filters,' number 12, pages 1155-1163, December 1980.

Note: The circuits included in this application note have been tested with National Semiconductor parts that may have been obsoleted and/or replaced with newer products. Please refer to the CLC to LMH conversion table to find the appropriate replacement part for the obsolete device.

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