

$E_{noe3} = A_{v0}(e_{nif1}/f_{zfl})\sqrt{(f_{pfl}^3 - f_{zfl}^3)/3}$, where $f_{pfl} = 1/2\pi R_f' C_S$, and C_S is the stray feedback capacitance.

$$E_{noe4} = 0.$$

$E_{noe5} = A_{v0}(e_{nif1}/f_{c2})\sqrt{1/f_{i2}}$, where f_{c2} is the unity-gain crossover frequency of A_2 , and $f_{i2} = f_{pfl} = 1/2\pi R_f' C_S$.

$E_{noe} = \sqrt{E_{noe1}^2 + E_{noe2}^2 + E_{noe3}^2 + E_{noe5}^2}$, the total rms output noise resulting from the spectral noise voltage e_{nif1} of A_1 .

$E_{noR} = A_{v0}\sqrt{KTR_f'\pi f_{pfl}}$, where $f_{pfl} = 1/2\pi R_f' C_S$.

$$E_{noi} = A_{v0}R_f'\sqrt{q\pi f_{pfl}I_B}/2.$$

$E_{noA1} = \sqrt{E_{noR}^2 + E_{noi}^2 + E_{noe}^2}$, the total output noise due to the A_1 current-to-voltage converter.

$E_{noA2} = A_{v0}e_{nif2}\sqrt{\pi f_{i2}/2}$, the output noise due to the A_2 voltage amplifier. Here, $f_{i2} = f_{c2}/A_{v0}$ and f_{c2} is the unity-gain crossover frequency of A_2 .

$E_{no} = \sqrt{E_{noA1}^2 + E_{noA2}^2}$, the total rms output noise of the two-op-amp photodiode amplifier.

7.3 Adding Voltage Gain

For many photodiode applications, the two op amps per photodetector above represents a significant drawback. Large arrays of photodetectors sometimes require hundreds of photodiode amplifiers. For such cases, one op amp provides both the current-to-voltage conversion and the voltage gain with somewhat less bandwidth improvement. The obvious circuit for this solution introduces nonlinearity and degrades rather than improves bandwidth. However, a bootstrap modification to this circuit avoids the nonlinearity and restores much of the bandwidth improvement of the two-amplifier solution.

7.3.1 Voltage gain alternatives

Evaluating the obvious circuit first illustrates the limitations later removed by the bootstrap. To provide both current-to-voltage conversion and voltage gain, this circuit first moves the conversion resistance R_f' to the noninverting input of the op amp as in Fig. 7.8. There, the signal current supplied by D_1 still flows through R_f' and again produces the desired signal voltage $i_p R_f'$. Moving R_f' also makes available the amplifier's normal feedback path for the addition of voltage gain. There, the addition of R_1 and R_2 sets this gain, converting the op amp to

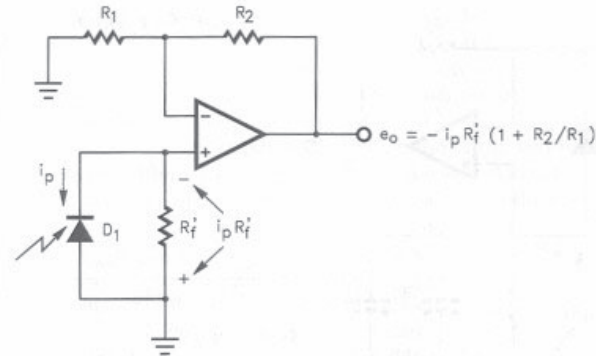


Figure 7.8 Using one op amp, the straightforward combination of current-to-voltage conversion and voltage gain impresses signal voltage on the diode, producing nonlinearity and restricting bandwidth.

a noninverting voltage amplifier. This amplifier provides the voltage gain $(1 + R_2/R_1)$ to the $i_p R_f'$ signal developed across R_f' . However, this signal voltage also falls across the photodiode, reintroducing two performance limitations described in Chap. 2. This voltage-mode photodiode operation produces response nonlinearity, and the signal voltage develops a current in the photodiode capacitance that shunts the photodiode output. The shunting bypass of R_f' produces a response pole at $1/2\pi R_f' C_D$ and, for most applications, dramatically reduces bandwidth when compared with the basic current-to-voltage converter.

Avoiding these limitations, Fig. 7.9 bootstraps the photodiode by connecting it directly between the op amp's inputs. This connection removes the primary signal swing $i_p R_f'$ from the diode, restoring the linear response and most of the bandwidth improvement. Here, the circuit's resistors perform generally the same functions as in the preceding circuit. Current from the photodiode still flows through R_f' , developing the same signal voltage, but that voltage no longer appears across the photodiode. The op amp feedback action reduces the voltage between the amplifier's inputs, and across the photodiode, to near zero. The bootstrap connection also supplies the diode current to the R_1, R_2 feedback network, altering the signal gain. However, the low resistances of this network produce little signal voltage in comparison with that produced by R_f' . Ignoring this added effect, the circuit again amplifies the voltage developed upon R_f' , increasing the output signal to $e_o = (1 + R_2/R_1)i_p R_f' = A_{v0}i_p R_f' = i_p R_T$. Here, $R_T = (1 + R_2/R_1)R_f'$ represents the equivalent transresistance of the circuit.

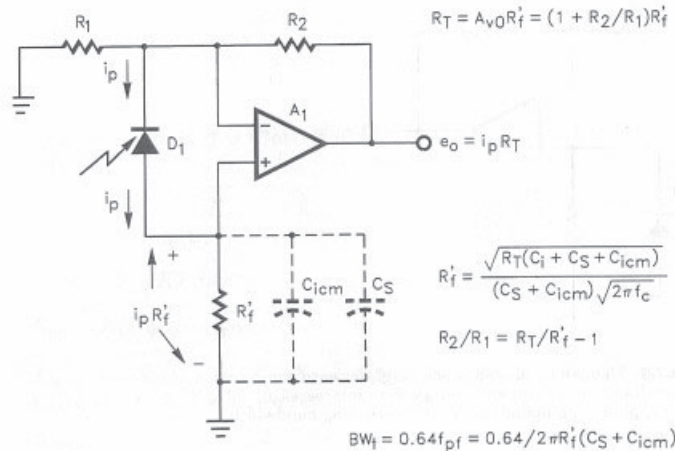


Figure 7.9 Bootstrapping the photodiode between the op amp inputs removes signal swing from the diode while retaining the signal gain of Fig. 7.8.

7.3.2 Optimizing the single-amplifier bandwidth-versus-noise compromise

This single-amplifier connection develops the same net transresistance as the preceding two-amplifier solution. However, the bandwidth improvement realized decreases here. A modified bandwidth limit results due to the move of the high-value conversion resistance R_f' . That resistor now connects to the noninverting input of the op amp where the common-mode input capacitance of the op amp C_{icm} also shunts R_f' . Previously, only the smaller stray capacitance, C_S of the two-amplifier circuit, shunted this resistor in its feedback connection. The increased shunting here rolls off bandwidth at $f_{pf} = 1/2\pi R_f' (C_S + C_{icm})$ rather than the preceding $f_{p\Omega} = 1/2\pi R_f' C_S$. Thus, compared with the two-amplifier case, the single-amplifier solution degrades the bandwidth improvement by the factor $1 + C_{icm}/C_S$.

For a given transresistance, $R_T = A_{v0} R_f'$, and increasing A_{v0} again reduces the required R_f' , moving this modified bandwidth limit to a higher frequency. However, increasing A_{v0} also reduces the frequency of the separate bandwidth limit imposed by the op amp. The amplifier imposes a bandwidth limit again marked by its $1/\beta$ intercept with the A_{OL} response. As described with the two-amplifier analyses, this limit occurs at $f_i = \beta f_c$ where β and f_c are the feedback factor and the unity-gain crossover frequency of the op amp. Increasing A_{v0} decreases β , reducing the frequency of the f_i bandwidth limit. In

compromise, increasing A_{v0} until the circuit's two bandwidth limits coincide, $f_{pf} = f_i = \beta f_c$, maximizes the final circuit bandwidth with neither limit imposing a dominant constraint. This choice also optimizes the circuit's noise-versus-bandwidth compromise as detailed in the discussion of Sec. 7.2.1.

Analysis quantifies the potential bandwidth improvement of the single-amplifier solution through comparison with the basic current-to-voltage converter. Adding voltage gain to reduce the basic R_f to R_f' initially increases bandwidth by reducing the sensitivity to parasitic capacitance. However, two effects described above limit this increase in the single-amplifier case. First, setting $f_i = f_{pf}$ produces a double rather than single-pole roll off for signal bandwidth, just as described for the two-amplifier case. This reduces signal bandwidth from $BW_i = f_{pf}$ to $BW_i = 0.64 f_{pf}$. Next, the added shunting of the amplifier's input capacitance tends to reduce the f_{pf} pole frequency, as seen by comparing the basic $f_{pf} = 1/2\pi R_f C_S$ with the new $f_{pf} = 1/2\pi R_f' (C_S + C_{icm})$. Noting that $R_f' = R_T/A_{v0}$, the new bandwidth limit becomes $BW_i = 0.64 A_{v0} / 2\pi R_T (C_S + C_{icm})$, where R_T represents the net transresistance of the circuit. For the same transresistance, the current-to-voltage converter alone requires a feedback resistance of $R_f = R_T$, making its $BW_i = 1/2\pi R_T C_S$ for comparison purposes. Dividing this BW_i expression into the preceding one shows that bandwidth changes by a factor of $0.64 A_{v0} / (1 + C_{icm}/C_S)$. In cases where $A_{v0} > 1.56(1 + C_{icm}/C_S)$, the single-amplifier solution improves bandwidth over that of the basic current-to-voltage converter.

To quantify A_{v0} the preceding $f_{pf} = f_i = \beta f_c$ optimum first requires defining the circuit's feedback factor β . However, moving the photodiode to its bootstrap position complicates this task. Positioned between the amplifier inputs, the diode provides an added, positive feedback path between the amplifier's output and noninverting input. Figure 7.10 models this connection for feedback analysis with $C_i = C_D + C_{id}$ combining the diode capacitance with the differential-input capacitance of the op amp. Through the feedback network, C_i couples a portion of output signal e_o to R_f' and the amplifier's noninverting input. This added coupling results in one feedback network that supplies feedback signals to both op amp inputs.

Normally, separate analyses of negative and positive feedback factors, β_- and β_+ , yields individual results for determination of the net feedback factor $\beta = \beta_- - \beta_+$. However, the single feedback network here simplifies the analysis given a reexamination of the feedback factor definition. Fundamentally, feedback factor is the fraction of the output signal fed back to the amplifier input. With two inputs, an op amp departs somewhat from this fundamental definition. The

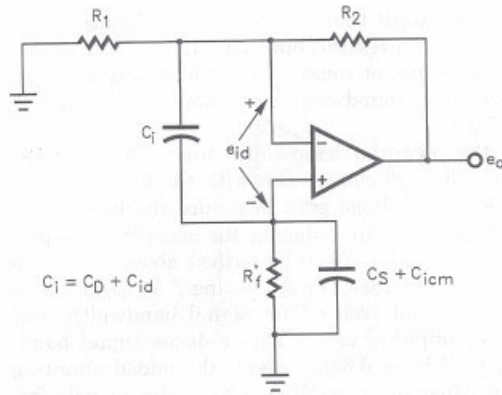


Figure 7.10 The photodiode bootstrap of Fig. 7.9 adds capacitance between the op amp inputs, coupling a positive feedback signal to R_f' . One feedback factor analysis combines the circuit's negative and positive feedback effects by solving for $\beta = e_{id}/e_o$.

differential nature of the two inputs subtracts the feedback signals supplied to the two, resulting in a net differential feedback signal e_{id} . Then, $e_{id}/e_o = \beta$, and one analysis combines the separate β_- and β_+ analyses that otherwise produce more complex intermediate results. In this analysis, the approximation $R_1 \ll Z_f'$ further simplifies the result, where Z_f' represents the parallel impedance combination of $C_S + C_{icm}$ and R_f' . The low value of $C_S + C_{icm}$ and the high value of R_f' support this approximation for high-gain photodiode amplifiers. Then, analyzing Fig. 7.10 for $e_{id}/e_o = \beta$ produces

$$\beta = \beta_0 \frac{1 + R_f'(C_S + C_{icm})s}{1 + R_f'(C_i + C_S + C_{icm})s}$$

where $\beta_0 = R_1/(R_1 + R_2) = 1/A_{v0} = R_f'/R_T$.

Manipulation of this β result and application of the $f_{pf} = f_i = \beta f_c$ condition defines R_f' and A_{v0} for the optimum. First, $f_i = \beta f_c$ defines the frequency of the $1/\beta$ intercept, and this occurs at a higher frequency where β equals its high-frequency value, β_H . There, the circuit's capacitances control the feedback and the preceding β expression reduces to

$$\beta_H = \frac{\beta_0(C_S + C_{icm})}{C_i + C_S + C_{icm}}$$

Substituting $\beta_0 = R_f'/R_T$, reduces this to

$$\beta_H = \frac{R_f'}{R_T} \frac{C_S + C_{icm}}{C_i + C_S + C_{icm}}$$

Then, equating $f_{pf} = 1/2\pi R_f'(C_S + C_{icm})$ and $f_i = \beta_H f_c$ yields the key design equation for the single-amplifier solution

$$R_f' = \frac{\sqrt{R_T(C_i + C_S + C_{icm})}}{(C_S + C_{icm})\sqrt{\omega_c}}$$

where $\omega_c = 1/2\pi f_c$.

Application of this equation defines the optimum A_{v0} and circuit resistances for a given application. First, the specific application defines the value of R_T required to produce the desired transresistance. Then, the equation defines the value of R_f' that optimizes the circuit's bandwidth-versus-noise compromise. Given R_T and R_f' , the relationship $A_{v0} = R_T/R_f'$ defines A_{v0} . Next, the relationship $A_{v0} = 1 + R_2/R_1$ defines the ratio $R_2/R_1 = R_T/R_f' - 1$. A second consideration converts this ratio into the final values for R_1 and R_2 . Together, these resistors present a load resistance of $R_L = R_1 + R_2$ to the op amp output. Picking a convenient resistance value for this load defines the individual resistors as $R_1 = (R_f'/R_T)R_L$ and $R_2 = (1 - R_f'/R_T)R_L$. These design choices for R_f' , R_1 , and R_2 produce a single-amplifier combination of current-to-voltage conversion and voltage gain with optimum bandwidth and minimized noise increase.

7.3.3 Noise analysis for the single-amplifier case

The noise performance of the single-amplifier solution of Fig. 7.8 repeats that of the preceding two-amplifier case with only a few changes. The added voltage gain again amplifies all noise components, and the $f_i = f_{pf}$ setting results in an e_{noe} curve identical in shape to the earlier Fig. 7.7. Adding the effects of the C_{icm} bypass adapts the earlier results to this case.

At first, it might seem that the bootstrap circuit removes the gain peaking effect upon the amplifier input noise voltage. Previously, this noise voltage produced a noise current with the diode's capacitance, and that current flowed through the large feedback resistance, producing the noise gain peaking. The bootstrap circuit here removes the large resistance from the feedback path but replaces it with the resistor R_f' at the amplifier's noninverting input. Noise current generated in C_i , by the noise voltage between the op amp inputs, still flows in the reconnected R_f' , producing much the same noise peaking effect. To

quantify this effect, consider the feedback factor expression β developed above. The op amp amplifies its input noise voltage by a gain of $1/\beta$ up to the gain roll off imposed by the amplifier's A_{OL} decline. From the previous β analysis, the bootstrap photodiode amplifier produces the noise gain

$$1/\beta = A_{v0} \frac{1 + R_f'(C_i + C_S + C_{icm})s}{1 + R_f'(C_S + C_{icm})s}$$

where $A_{v0} = 1/\beta_0 = 1 + R_2/R_1$. Thus, the noise gain displays the response zero, $f_{zf} = 1/2\pi R_f'(C_i + C_S + C_{icm})$, and the response pole, $f_{pf} = 1/2\pi R_f'(C_S + C_{icm})$, that characterize noise gain peaking. Together, the added voltage gain A_{v0} and the gain peaking effect reproduce the Fig. 7.7 e_{noe} curve with just the f_{zf} and f_{pf} differences here noted in the listing below. Then, calculation of the noise result for the Fig. 7.9 bootstrap amplifier follows the step-by-step procedure:

$E_{noe1} = A_{v0} e_{nif} \sqrt{f_f \ln(f_f/f_1)}$, where $A_{v0} = 1 + R_2/R_1$. Also, e_{nif} is the e_{ni} noise floor level, and f_f is the $1/f$ noise corner frequency of the op amp. As explained in Sec. 5.2.3, setting $f_1 \approx 0.01$ Hz avoids a theoretical contradiction and ensures analytical accuracy.

$E_{noe2} = A_{v0} e_{nif} \sqrt{f_{zf} - f_f}$, where $f_{zf} = 1/2\pi R_f'(C_i + C_S + C_{icm})$, where $C_i = C_D + C_{id}$, C_S is the stray capacitance, and C_{icm} is the common-mode input capacitance of the op amp. Numerical evaluation here can produce the square root of a negative number. Then, substitute zero for the result as explained in Sec. 5.3.1.

$E_{noe3} = A_{v0} (e_{nif}/f_{zf}) \sqrt{(f_{pf}^3 - f_{zf}^3)/3}$, where $f_{pf} = 1/2\pi R_f'(C_S + C_{icm})$.

$E_{noe4} = 0$.

$E_{noe5} = A_{v0} (e_{nif} f_{c2}) \sqrt{1/f_i}$, where f_c is the unity-gain crossover frequency of the amplifier.

$E_{noe} = \sqrt{E_{noe1}^2 + E_{noe2}^2 + E_{noe3}^2 + E_{noe5}^2}$, the total rms output noise resulting from the spectral noise voltage e_{ni} of the amplifier input.

$E_{noR} = A_{v0} \sqrt{KTR_f' \pi f_{pf}}$.

$E_{noi} = A_{v0} R_f' \sqrt{q \pi f_{pf} I_B - 1/2}$.

$E_{no} = \sqrt{E_{noR}^2 + E_{noi}^2 + E_{noe}^2}$, the total rms output noise of the bootstrap photodiode amplifier.

Compared with the two-amplifier case, the above results eliminate E_{noeA2} and display changes in C_i , f_{zf} , and f_{pf} . Noise voltage E_{noeA2} ,

the noise of A_2 in the previous results, no longer exists for this single-amplifier solution. Capacitance C_i becomes $C_i = C_D + C_{id}$, in this case, eliminating the previous C_{icm} component. For this circuit, the role of C_{icm} transfers to the bypass R_f 's signal voltage. This bypass alters f_{zf} and f_{pf} as described in the $1/\beta$ analysis where $f_{zf} = 1/2\pi R_f'(C_i + C_S + C_{icm})$ and $f_{pf} = 1/2\pi R_f'(C_S + C_{icm})$.

7.4 Adding Current Gain

Adding current rather than voltage gain presents another bandwidth-improving alternative to high-gain photodiode amplifiers. This alternative best serves applications requiring a current output and, there, current gain replaces transimpedance as the amplifier's gain mechanism. However, for photodiode sources, direct application of the op amp current amplifier produces nonlinearity, latching, and oscillations. Like the single-amplifier case above, bootstrapping the photodiode avoids these problems by removing signal swing from the photodiode. Then, judicious component selection again optimizes the circuit's noise-versus-bandwidth compromise. A simple feedback analysis transforms previous noise results to this current-gain case.

7.4.1 Current gain alternatives

Op amp circuits readily produce current rather than voltage gain,² and direct application of the basic, op amp current amplifier to photodiodes produces Fig. 7.11. Analysis of this circuit's performance first defines

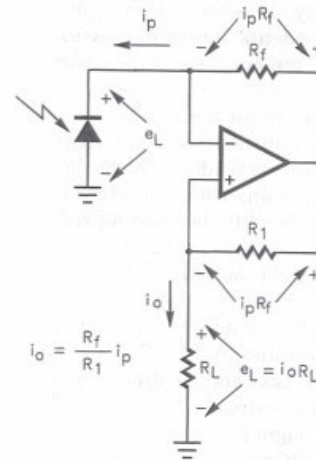


Figure 7.11 Connecting a photodiode directly to an op-amp current amplifier adds current gain but impresses signal voltage e_L across the diode, producing nonlinearity and latching.

the problems of the basic current-gain case, and then analysis of the bootstrap alternative illustrates the solution. In the figure, negative and positive feedback combine to deliver a current gain controlled by two resistors, R_f and R_1 . This circuit actually performs a current-to-voltage conversion with R_f followed by the inverse function with R_1 , adding gain in the process. The added gain again permits reducing R_f to increase bandwidth, and the circuit's single-amplifier structure avoids the noise of a second amplifier. Further, the circuit's dual function provides an extra degree of design freedom that again permits optimization of the noise-versus-bandwidth compromise.

Temporarily neglecting capacitance effects illustrates this dual function to define the circuit's low-frequency current gain A_{i0} . At lower frequencies, signal current i_p from the photodiode flows through R_f , producing the input to output voltage drop $i_p R_f$. Feedback replicates this voltage across R_1 to retain zero voltage between the op amp inputs. In doing this, the feedback action supplies the current $i_o = i_p R_f / R_1$ through R_1 and to the load R_L . Thus, the circuit's current gain i_o / i_p is $A_{i0} = R_f / R_1$ at low frequencies. This added gain permits the bandwidth-improving reduction of R_f , as described for the two preceding circuits.

However, this direct application of the basic current amplifier produces three severe problems for the photodiode amplifier.³ Each of these problems results from impressing the load voltage across the photodiode. In the first, nonlinearity develops from the inherent variation of photodiode responsivity with the diode's voltage. In Fig. 7.11, feedback replicates the load's signal voltage e_L directly across the diode, modulating the diode's responsivity. In a second problem, this signal condition produces a latch state when a larger e_L causes forward bias of the photodiode. Normally, the output signal current i_o develops an e_L voltage polarity that holds the diode in a reverse-biased state. However, the transient condition of circuit turn on can momentarily produce a forward bias. Then, the diode shorts the negative feedback of R_f to ground, leaving the positive feedback of R_1 in control of the circuit. This positive feedback dominance latches the circuit in this forward-biased state. Limiting the load impedance R_L reduces the possibility of the larger e_L required to produce this latch state. However, the uncertainty of transient conditions during circuit turn on make the circuit unreliable.

In a third problem condition, impressing e_L across the photodiode capacitance promotes oscillation as illustrated by Fig. 7.12. Voltage e_L results from a positive feedback connection and its drop across capacitance C_D produces a positive feedback current. This current flows to R_f , transferring a positive feedback signal to the negative feedback path, inviting oscillation. From another perspective, capacitance

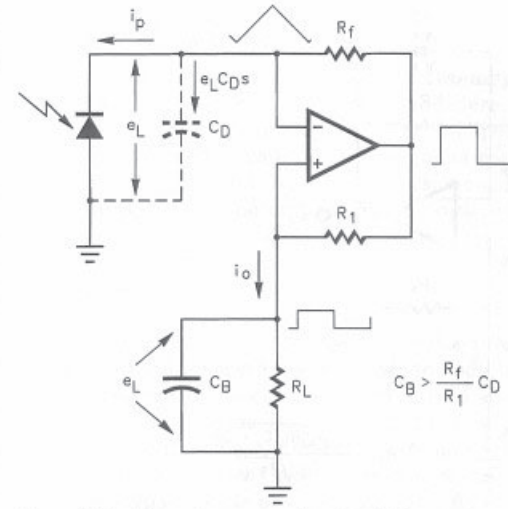


Figure 7.12 The e_L drop across the photodiode capacitance also promotes oscillation with the Fig. 7.11 connection.

C_D rolls off the negative feedback of R_f , making it possible for the circuit's positive feedback to dominate through R_1 . In fact, this feedback combination inadvertently converts the circuit to the basic op amp square-wave generator.⁴ For this generator, the op amp acts as a comparator, producing a square wave output in response to triangle- and square-wave signals at the amplifier's two inputs. These input signals result from an integrating negative feedback and a hysteresis-producing positive feedback. Here, R_f and C_D integrate the square-wave output signal, producing a triangle wave at the amplifier's inverting input. Meanwhile, R_1 and R_L transfer a square wave to the amplifier's noninverting input, shifting the threshold for the ensuing comparator action of the amplifier. Each time the triangle wave at the inverting input reaches the level of the square wave at the noninverting input, the circuit switches comparator states. To prevent this oscillation, a dominant roll off must be added to the positive feedback path through the load bypass of C_B . This requires a bypass capacitance $C_B > A_{i0} C_D = R_f C_D / R_1$ and devastates signal bandwidth.

Bootstrapping the photodiode avoids all three of the above problems by removing the load's signal voltage from the photodiode. Figure 7.13 shows this solution with the previously grounded end of the diode returned to the top of load R_L . Then, signal voltage e_L no

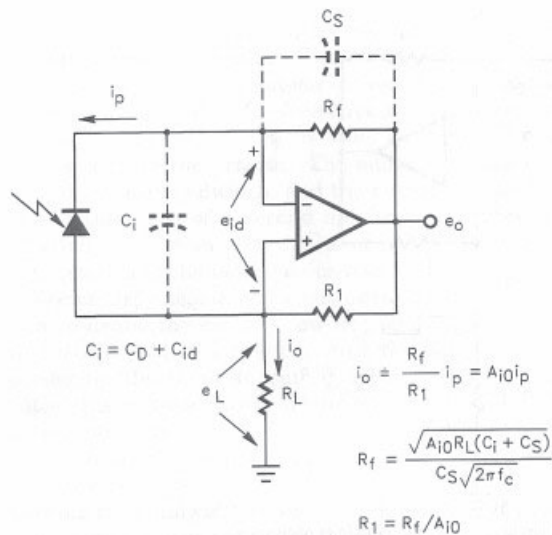


Figure 7.13 Bootstrapping the photodiode removes e_L from the diode, avoiding the problems of Fig. 7.11 and delivering the desired current gain $A_{i0} = i_o/i_p = R_f/R_1$.

longer drops across the diode, avoiding the nonlinearity, latching, and oscillation of the preceding circuit. Here, the e_L signal drives the op amp's noninverting input, and feedback virtually duplicates this signal at the amplifier's inverting input. Thus, the positive feedback signal becomes a common-mode signal at the amplifier's two inputs, and common-mode rejection virtually removes the signal from the diode. Only the small input error signal of the op amp, e_{id} , remains across the photodiode. This very small signal leaves the diode responsivity essentially unchanged and avoids the forward-biased diode state that produced latching. Also, removing the signal voltage from the diode's capacitance avoids the coupling of a positive feedback current into the negative feedback path. The negative feedback network now rides atop the positive feedback signal, preventing interaction. Given these cures, the bootstrapped current amplifier reliably produces the current gain $A_{i0} = i_o/i_p = R_f/R_1$ as originally intended for the previous circuit. The bootstrap connection actually increases this gain slightly due to the added flow of i_p from the diode's anode directly to R_L . However, high-gain applications typically produce $i_o \gg i_p$, making this added gain negligible.

7.4.2 Optimizing the current-gain bandwidth-versus-noise compromise

This compromise parallels that of the two-amplifier solution previously described in Sec. 7.2.1. There and here, both the bandwidth and noise performance of the circuit exhibit peaks in their relationships with the added gain. The two peaks occur at the same gain value, and the optimum bandwidth-versus-noise performance occurs at this peak value. The earlier discussion defines the origins of this optimum. As before, a specific A_{i0} value here optimizes bandwidth and the bandwidth-versus-noise compromise simultaneously. Thus, an analysis focus upon bandwidth versus A_{i0} automatically includes the noise compromise as well.

The bootstrapped current amplifier returns bandwidth control to the limits normally imposed upon photodiode amplifiers. Bootstrap removes the positive feedback coupling that previously required the C_B bypass of the load in Fig. 7.12. That bypass dramatically reduced signal bandwidth as described before. Removing this bypass, the bootstrap alternative returns bandwidth control to the limits imposed by capacitive bypass of R_f and the A_{OL} roll off of the op amp. In high-gain cases, the parasitic C_S bypass of R_f sets a bandwidth limit at the pole frequency $f_{pf} = 1/2\pi R_f C_S$. As before, adding circuit gain permits reduction of R_f and increases this f_{pf} limit. However, the second bandwidth limit, produced by the amplifier's A_{OL} roll off, ultimately restricts the degree of improvement. This roll off results from the amplifier's limited gain-bandwidth product beyond which the amplifier lacks the loop gain required to support the circuit's ideal response, $A_{i0} = i_o/i_p$. The intercept of the circuit's $1/\beta$ curve and the A_{OL} response defines the maximum frequency for which A_{OL} supports the gain demand of the feedback. As described with the earlier Fig. 7.6, this intercept imposes a bandwidth limit at $f_i = \beta_H f_c$, where β_H is the high-frequency value of the circuit's feedback factor β , and f_c is the unity-gain crossover frequency of A_{OL} . Making the f_{pf} and f_i bandwidth limits coincident optimizes bandwidth and defines design equations for the bootstrapped current amplifier. Aligning f_i with f_{pf} again produces a two-pole roll off for the signal response, making $BW_i = 0.64 f_{pf}$, where $f_{pf} = 1/2\pi R_f C_S$.

Equating the expressions for f_{pf} and f_i defines optimum-bandwidth design equations for R_f and R_1 in terms of known circuit quantities. However, the combined negative and positive feedback paths complicate the determination of $f_i = \beta_H f_c$. This requires knowledge of the circuit's feedback factor β , and the two feedback paths both affect this factor. Still, the combined effect of the two develops a single differential input signal e_{id} between the op amp's inputs. This signal defines the fraction of the output fed back to the amplifier's differential input

or the feedback factor. Analyzing the circuit for $e_{id}/e_o = \beta$ produces a complex result, but practical conditions reduce it to a simple equation. In practice, $Z_f \gg R_1$, where Z_f is the paralleled combination of R_f and C_S . The low value of C_S and, for high-gain cases, the high value of R_f make $Z_f \gg R_1$ over the useful range of the op amp bandwidth. Given this approximation, the circuit's feedback factor becomes

$$\beta = \frac{R_1}{R_1 + R_L} \frac{1 + R_f C_S s}{1 + R_f (C_i + C_S) s}$$

where $C_i = C_D + C_{id}$. Here, C_D is the photodiode capacitance, and C_{id} is the differential-input capacitance of the op amp. At high frequencies, the circuit capacitances dominate, reducing this feedback factor to

$$\beta_H = \left(\frac{R_1}{R_1 + R_L} \right) \left(\frac{C_S}{C_i + C_S} \right) = \beta_0 \frac{C_S}{C_i + C_S}$$

Given this expression, $f_i = \beta_H f_c$ defines f_i for the coincident bandwidth limit of the $f_{pf} = f_i$ bandwidth optimization. However, equating f_{pf} and f_i and solving for the required R_f yields another complex result. A practical consideration again offers a simplification through an approximation for β_0 , the low-frequency value of β . Here, $\beta_0 = R_1/(R_1 + R_L)$, and output compliance considerations favor the condition $R_1 \ll R_L$, making $\beta_0 \approx R_1/R_L$. Output compliance range, or the voltage limit across R_L , results from the saturation voltage limit of the op amp output. For linear response, the amplifier output voltage e_o must support the voltage drops across both R_1 and R_L without encountering this saturation limit. These two resistances conduct the same current i_o , and making $R_1 \ll R_L$ reserves most of the amplifier's voltage range for the voltage across R_L . Then, $\beta_0 \approx R_1/R_L$, making $\beta_H = (R_1/R_L)C_S/(C_i + C_S)$ and $f_i = \beta_H f_c = (R_1/R_L)f_c C_S/(C_i + C_S)$. Equating this f_i expression to $f_{pf} = 1/2\pi R_f C_S$ and solving for R_f yields the design equation

$$R_f = \frac{\sqrt{A_{i0} R_L (C_i + C_S)}}{C_S \sqrt{\omega_c}}$$

where $C_i = C_D + C_{id}$ and $\omega_c = 1/2\pi f_c$. In this equation, conditions of the specific circuit define all of the variables except A_{i0} . This last variable depends upon the current gain requirement of the specific application, as chosen by the user. Choosing a value for A_{i0} for the requirement then defines R_f . Next, the design value for R_1 follows from $R_1 = R_f/A_{i0}$. Together, the design equations of this paragraph define the circuit conditions that optimize the circuit's gain-bandwidth compromise. These equations simultaneously optimize the

circuit's bandwidth-versus-noise compromise as described earlier in this section.

7.4.3 Noise analysis for the current gain case

Noise analysis for the bootstrapped current amplifier requires two steps. First, the conventional E_{no} analysis defines the rms noise voltage at the amplifier output. Then, a conversion analysis relates the E_{no} result to the output noise current I_{no} . As before, E_{no} results from the noise components E_{noR} , E_{noi} , and E_{noe} . These components represent the rms output noise resulting from circuit resistance and from the input current and voltage noises of the amplifier. Gain and bandwidth differences distinguish the noise results here from those of the basic current-to-voltage converter. First, the circuit's positive feedback amplifies the basic noise effect for each noise component by a factor of $1/\beta_0$. Here, $\beta_0 = R_1/(R_1 + R_L)$ is the low-frequency value of the circuit's feedback factor as developed before. Next, the circuit's two-pole roll off, resulting from the $f_i = f_{pf}$ alignment, reduces the effect of each noise component through a steeper response roll off.

The amplified, two-pole results parallel those of the single-amplifier solution described in Sec. 7.3.2. Converting those earlier results to this case replaces the previous gain A_{v0} with $1/\beta_0$ here. Previously, the noise results reflected the performance of the similarly bootstrapped amplifier of Fig. 7.9. One other difference separates the two cases because of the change in the capacitance bypassing the R_f resistance. Here, stray capacitance C_S bypasses this resistance rather than the $C_S + C_{icm}$ before. Thus, replacing A_{v0} with $1/\beta_0$ and $C_S + C_{icm}$ with C_S translates the Sec. 7.3.3 noise results for application to this case. Then, for the Fig. 7.13 current amplifier the noise voltage at the amplifier output follows from the analysis steps below:

$E_{noe1} = (e_{nif}/\beta_0) \sqrt{f_f \ln(f_f/f_1)}$, where $1/\beta_0 = 1 + R_L/R_1$. Here, e_{nif} is the e_{ni} noise floor level, and f_f is the $1/f$ noise corner frequency of the op amp. As explained in Sec. 5.2.3, setting $f_1 \approx 0.01$ Hz avoids a theoretical contradiction and ensures analytical accuracy.

$E_{noe2} = (e_{nif}/\beta_0) \sqrt{f_{zf} - f_f}$, where $f_{zf} = 1/2\pi R_f (C_i + C_S)$, and $C_i = C_D + C_{id}$. Numerical evaluation here can produce the square root of a negative number. Then, substitute zero for the result as explained in Sec. 5.3.1.

$E_{noe3} = (e_{nif}/\beta_0 f_{zf}) \sqrt{(f_{pf}^3 - f_{zf}^3)/3}$, where $f_{pf} = 1/2\pi R_f C_S$.

$E_{noe4} = 0$.

$E_{noe5} = (e_{nif} f_c / \beta_0) \sqrt{1/f_i}$, where f_c is the unity-gain crossover frequency of the amplifier, and $f_i = f_{pf} = 1/2\pi R_f C_S$.

$E_{noe} = \sqrt{E_{noe1}^2 + E_{noe2}^2 + E_{noe3}^2 + E_{noe5}^2}$, the total rms output noise resulting from the spectral noise voltage e_{ni} of the amplifier input.

$$E_{noR} = (1/\beta_0) \sqrt{KTR_f \pi f_{pf}}$$

$$E_{noi} = (R_f/\beta_0) \sqrt{q \pi f_{pf} I_{B-}/2}$$

$E_{no} = \sqrt{E_{noR}^2 + E_{noi}^2 + E_{noe}^2}$, the total rms noise at the op amp output.

Having calculated E_{no} , the actual noise of interest I_{no} follows from a simple conversion. In Fig. 7.13, the positive feedback to the amplifier's noninverting input delivers a portion of e_o , including a portion of E_{no} , to the load R_L . This portion, or fraction of the output fed back to the noninverting input, equals the circuit's positive feedback factor β_+ . Thus, the portion of E_{no} noise voltage delivered to R_L equals $(\beta_+)E_{no}$, making $I_{no} = (\beta_+)E_{no}/R_L$. Two feedback paths contribute to β_+ with the primary path through R_1 . A secondary path develops through R_f , C_S , and C_i , but this path contributes much less feedback signal. For high-gain photodiode amplifiers, the parallel impedance of R_f and C_S greatly exceeds that of R_1 over the useful frequency range of the amplifier. Neglecting the secondary path, $\beta_+ = R_L/(R_1 + R_L)$, making the output noise current of Fig. 7.13

$$I_{no} = \frac{E_{no}}{R_1 + R_L}$$

References

1. J. Graeme, "The Tee-Feedback Factor in Photodiode Amplifiers," *Electronic Design*, June 26, 1995, p. 98.
2. J. Graeme, "Manipulate Current Signals with Op Amps," *EDN*, August 8, 1985, p. 147.
3. J. Graeme, "Tame Photodiodes with Op Amp Bootstrap," *Electronic Design*, May 12, 1988, p. 89.
4. G. Tobey, J. Graeme, and L. Huelsman, *Operational Amplifiers; Design and Applications*, McGraw-Hill, New York, 1971.

Reducing Power-Supply Noise Coupling

Noise on the power-supply lines couples into the signal path of the photodiode amplifier through the finite PSRR of the op amp. This noise most often results from the line impedance interaction with currents drawn by the amplifier and other circuitry powered from the same supply lines. Supply current flows through these impedances, producing voltage drops that represent noise at the amplifier supply connections. This power-supply noise couples through the amplifier's PSRR to reproduce a portion of the noise signal at the op amp inputs. There, the coupled noise combines with the amplifier's normal input noise to be amplified by the noise gain peaking of the photodiode amplifier. This gain peaking effect makes supply noise reduction especially important for the photodiode amplifier. The gain peaking also makes the photodiode amplifier more vulnerable to oscillation. That portion of the supply noise resulting from the amplifier itself constitutes a parasitic feedback signal, capable of producing this self-generated noise condition.

Numerous techniques commonly reduce this parasitic coupling through wide supply buss traces, minimized buss lengths, star connections, and supply bypass. The first two techniques, buss width and length control, reduce supply-line resistance and inductance. Star connections provide separate supply runs to sensitive and noisy circuitry, such as analog and digital, to minimize device interaction. However, most importantly, capacitive bypass of the power-supply lines greatly attenuates the noise interference and ensures frequency stability as well. To be successful, the bypass selection requires close attention to the frequency-dependent impedances of both the supply lines and the bypass capacitors. Initial intuition suggests simply placing the bypass

$E_{noe5} = (e_{nif} f_{c2} / \beta_0) \sqrt{1/f_i}$, where f_c is the unity-gain crossover frequency of the amplifier, and $f_i = 1/2\pi R_f C_S$.

$E_{noe} = \sqrt{E_{noe1}^2 + E_{noe2}^2 + E_{noe3}^2 + E_{noe5}^2}$, the total rms output noise resulting from the spectral noise voltage e_{ni} of the amplifier input.

$$E_{noR} = (1/\beta_0) \sqrt{KTR_f \pi f_{pf}}$$

$$E_{noi} = (R_f/\beta_0) \sqrt{q \pi f_{pf} I_{B-}/2}$$

$E_{no} = \sqrt{E_{noR}^2 + E_{noi}^2 + E_{noe}^2}$, the total rms noise at the op amp output.

Having calculated E_{no} , the actual noise of interest I_{no} follows from a simple conversion. In Fig. 7.13, the positive feedback to the amplifier's noninverting input delivers a portion of e_o , including a portion of E_{no} , to the load R_L . This portion, or fraction of the output fed back to the noninverting input, equals the circuit's positive feedback factor β_+ . Thus, the portion of E_{no} noise voltage delivered to R_L equals $(\beta_+)E_{no}$, making $I_{no} = (\beta_+)E_{no}/R_L$. Two feedback paths contribute to β_+ with the primary path through R_1 . A secondary path develops through R_f , C_S , and C_i , but this path contributes much less feedback signal. For high-gain photodiode amplifiers, the parallel impedance of R_f and C_S greatly exceeds that of R_1 over the useful frequency range of the amplifier. Neglecting the secondary path, $\beta_+ = R_L/(R_1 + R_L)$, making the output noise current of Fig. 7.13

$$I_{no} = \frac{E_{no}}{R_1 + R_L}$$

References

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Numerous techniques commonly reduce this parasitic coupling through wide supply buss traces, minimized buss lengths, star connections, and supply bypass. The first two techniques, buss width and length control, reduce supply-line resistance and inductance. Star connections provide separate supply runs to sensitive and noisy circuitry, such as analog and digital, to minimize device interaction. However, most importantly, capacitive bypass of the power-supply lines greatly attenuates the noise interference and ensures frequency stability as well. To be successful, the bypass selection requires close attention to the frequency-dependent impedances of both the supply lines and the bypass capacitors. Initial intuition suggests simply placing the bypass

capacitors right at the amplifier's power-supply terminals and making the capacitors large.

Here, the first intuitive assumption works but the second fails. Close placement minimizes the interconnect length and the associated inductance between the amplifier and the capacitors. For best results, the bypass capacitors should be surface-mount types mounted right between the amplifier's power-supply pins and a ground plane. However, practical layout considerations often prevent this ideal condition, and in any case, amplifier lead impedance remains unbypassed. The second intuitive assumption, making the capacitors large, fails because of the parasitic impedances of such capacitors. Large capacitor values do minimize the bypassed line impedances at lower frequencies where the capacitors remain purely capacitive. However, the internal construction of larger capacitors produces a significant parasitic inductance, interrupting the bypass effectiveness at higher frequencies. In compromise, the primary bypass capacitance value should be made large enough to sufficiently limit line impedance but no larger.

Where necessary, adding a smaller capacitor in parallel with the first commonly bypasses the inductance of the first. However, the secondary capacitor has inductance of its own, producing another bypass compromise at a somewhat higher frequency. Further, the secondary capacitor reacts with the inductance of the first, producing a resonant increase in the supply-line impedance. Careful selection of these capacitors produces a compromise solution that retains low bypass impedance throughout the amplifier's significant frequency range. For this selection, analysis of the net bypass impedance produces two simple design equations that guide the capacitor selection. For the more severe coupling cases, a resistive detuning of the bypass impedance preserves stability, or the addition of series line impedance forms decoupling filters. For both of these special cases additional design equations again guide the component selection.

8.1 The Power-Supply Bypass Requirement

Noise on the power-supply lines of an op amp circuit couples to the circuit's output with a magnitude controlled by the supply-line impedance, the amplifier's PSRR, and the circuit's noise gain. Typically, the inductive portion of the line impedance produces the most serious coupling due to the associated impedance rise with increasing frequency. Simultaneous with this rise, the amplifier's PSRR declines, increasing the parasitic coupling further. For photodiode amplifiers, yet a third increase results from the characteristic noise gain peaking. Bypass capacitors convert the line impedance rise to a decline to restrain this coupling increase. In addition, these capacitors must pre-

serve frequency stability by sufficiently reducing the line impedance at higher frequencies. Part of the supply-line noise signal results from the amplifier itself and establishes a parasitic feedback loop capable of producing oscillation.

8.1.1 The noise-coupling mechanism

Figure 8.1 illustrates the fundamental need for power-supply bypass. There, positive and negative power supplies V_+ and V_- supply the corresponding bias terminals of the amplifier with the voltages V_P and V_N . Ideally, $V_P = V_+$ and $V_N = V_-$ but intervening, supply-line inductances L_P react with the signal current i_S , producing voltage differences. Here, the signal current i_S results from the combined supply currents drawn by the op amp and other circuitry powered from the same supply lines. The L_P inductances result from an inherent parasitic effect of the connecting lines between the power supplies and the amplifier. A typical printed circuit board trace introduces about 15 nH of parasitic inductance per inch of trace and wire introduces a similar value.¹ While seemingly small, complex board and wiring connections produce inductances in the hundreds of nanohenrys. The resulting supply-line impedances can seriously degrade noise performance and frequency stability. Adding the C_B bypass capacitors shown rolls off these impedances to attenuate the supply coupling effects.

The circuit model shown intentionally oversimplifies the actual conditions of a practical circuit to provide more intuitive insight into the bypass requirement. First, the model omits the input signal to permit

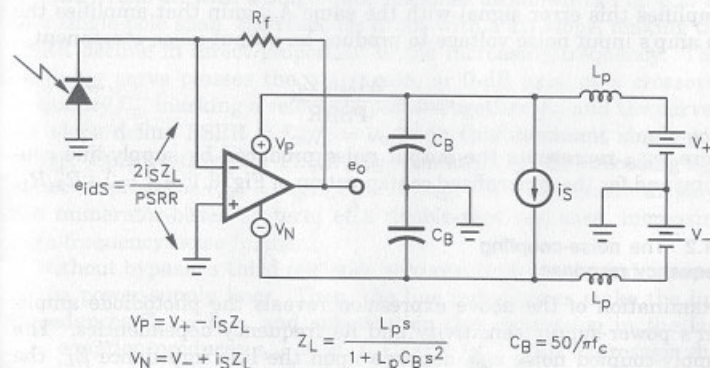


Figure 8.1 Bypass capacitors C_B suppress supply voltage changes produced by the reaction of supply current drain i_S with power-supply line inductances L_P .

focus upon only those signals relating to the line impedance and its bypass. Next, the model simplifies the line conditions. In practice, the V_+ and V_- supplies add output impedances and noise sources of their own, different circuits draw signal currents at different points upon the supply lines, and the supply-line resistance adds to the voltage drops produced with i_S . However, the bypass practices developed with this simplified model also contend with these other effects.

Intuitive evaluation of the model provides a first insight into the bypass benefit. Fundamentally, the C_B bypass capacitors shunt the line impedances to reduce the supply-line voltage drops produced by i_S . However, from another perspective, the capacitors serve as local reservoirs for the immediate supply of higher-frequency current demands. Otherwise, such currents encounter significant time delays in their travel from the power supplies and through the line inductances to the op amp. This delay produces phase shift in the amplifier response and, for a given time delay, the corresponding phase shift increases with frequency. To counteract this, the charge stored on the bypass capacitors supplies much of the high-frequency current demand locally, decreasing the time delay and the corresponding phase shift.

Circuit analysis quantifies the bypass benefit. The bypass capacitors reduce but do not eliminate the power-supply coupling effects. Voltage differences still develop with the residual impedance of the L_p , C_B combinations. The net line impedance Z_L reacts with signal current i_S , reducing the voltage magnitudes at both V_P and V_N by the amount $i_S Z_L$. Then, $V_P = V_+ - i_S Z_L$ and $V_N = V_- + i_S Z_L$, so the total supply voltage delivered to the op amp, $V_P - V_N = V_+ - V_- - 2i_S Z_L$, decreases by the amount $2i_S Z_L$. This reacts with the PSRR of the op amp, producing an amplifier input error of $e_{idS} = 2i_S Z_L / \text{PSRR}$. The circuit amplifies this error signal with the same A_{ne} gain that amplifies the op amp's input noise voltage to produce an output noise component

$$e_{noS} = \frac{2A_{ne}i_S Z_L}{\text{PSRR}}$$

Here, e_{noS} represents the output noise produced by supply-line coupling and for the generalized configuration of Fig. 8.1, $A_{ne} = 1 + R_2/R_1$.

8.1.2 The noise-coupling frequency response

Examination of the above expression reveals the photodiode amplifier's power-supply sensitivity and its frequency dependencies. The supply-coupled noise e_{noS} depends upon the line impedance Z_L , the amplifier's PSRR, and the noise gain A_{ne} . In the unbypassed state, each characteristic varies in a manner that increase e_{noS} as frequency

increases. Comparison of unbypassed and bypassed cases reveals that supply bypass reduces the resulting e_{noS} response from a triple zero to the single zero normally expected with the photodiode amplifier. Unbypassed, a rising A_{ne} gain, a declining PSRR, and a rising Z_L each introduce zeroes in the e_{noS} response. Bypassed, a declining PSRR and a declining Z_L produce canceling effects and leave the single zero of the rising A_{ne} . Examination of these effects adds insight into potential noise origins for troubleshooting a circuit, especially in light of the practical bypass deficiencies described later.

As described before, the response zero of the A_{ne} noise gain results from the amplifier input error signal across the photodiode capacitance. This parallels the amplifier input noise voltage drop described in Sec. 5.1.2. In the figure, the $e_{idS} = 2i_S Z_L / \text{PSRR}$ error voltage drops across the photodiode and its capacitance. The resulting capacitive current $e_{idS} C_D s$ flows through feedback resistor R_f , producing the amplified output noise voltage, $e_{noS} = e_{idS} C_D s R_f = 2i_S Z_L C_D s R_f / \text{PSRR}$. As frequency increases, this error rises due to the numerator-based s term here. From a circuit perspective, this rise reflects the falling capacitive impedance and the increasing capacitive error current that results. The resulting output noise would simply follow the normal gain peaking effects of the Sec. 5.1.2 noise voltage analysis, but the other two response zeros increase the noise response further.

The second response zero results from the declining PSRR of the op amp. This decline increases the $e_{idS} = 2i_S Z_L / \text{PSRR}$ input error signal to in turn increase $e_{noS} = A_{ne} e_{idS}$. An approximation to the PSRR frequency response simplifies the mathematical expression of this effect. Over most of the amplifier's useful frequency range, PSRR typically rolls off with a single-pole response as shown in Fig. 8.2. In this single-pole span, the response drops with a 1:1 slope, making the PSRR decline in direct proportion to the increasing frequency. This declining curve crosses the unity gain, or 0-dB axis, at a crossover frequency f_{cp} , marking a reference point. Together, f_{cp} and the curve's 1:1 slope define $\text{PSRR} \approx f_{cp}/f = \omega_{cp}/s$ in this dominant single-pole range. Substituting this PSRR approximation in the preceding e_{noS} equation produces $e_{noS} = 2i_S Z_L C_D s^2 R_f / \omega_{cp}$. This expression displays the numerator-based s^2 term of a double-zero response, increasing high-frequency noise further.

Without bypass, a third response zero results from the inductances of the power-supply lines. Then, the line inductances make the line impedances $Z_L = L_p s$, and substituting this expression in the last e_{noS} equation produces $e_{noS} = 2i_S L_p C_D s^3 R_f / \omega_{cp}$. This expression displays the numerator-based s^3 factor of a triple-zero response. At higher frequencies, such a response produces a very large gain for the

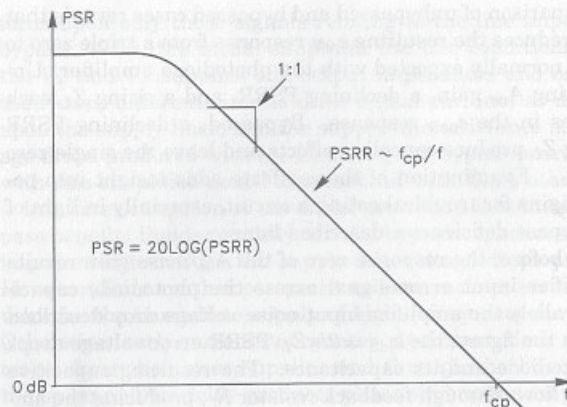


Figure 8.2 Op amp power-supply rejection displays a frequency response approximated by $\text{PSRR} \sim f_{cp}/f$, where f_{cp} is the unity-gain crossover of PSRR.

power-supply noise signal. Reducing this gain requires decreasing L_p and/or increasing ω_{cp} . Decreasing L_p reduces the supply-line noise signal, and increasing ω_{cp} improves the PSRR ability to reject this noise. These noise reduction options combine with power-supply bypass in the control of power-supply noise coupling.

Adding the C_B bypass capacitors of Fig. 8.1 dramatically reduces this noise coupling by removing two of the three zeroes in the e_{nos} response. Ideally, capacitive bypass dominates the higher-frequency line impedances, converting them from $Z_L = L_p s$ to $Z_L = 1/C_B s$. This removes a zero and adds a pole. Making this $Z_L = 1/C_B s$ substitution in the $e_{nos} = 2i_S Z_L C_D s^2 R_f / \omega_{cp}$ equation produces $e_{nos} = 2i_S C_D s R_f / C_B \omega_{cp}$. Here, a simple s factor in the numerator replaces the previous s^3 , reflecting a single rather than triple zero for the e_{nos} response. Then, the power-supply noise produced by signal current i_S , and coupled to the amplifier input through PSRR, only receives the same amplification as the amplifier's input noise voltage. However, this bypass must also ensure frequency stability, and bypass deficiencies still require attention to the noise reduction options involving reduced L_p and increased ω_{cp} as described above.

8.1.3 Power-supply coupling and frequency stability

In addition to noise reduction, the power-supply bypass must serve a second and generally more prevailing purpose, the preservation of fre-

quency stability. Lack of stability or oscillation produces the ultimate noise signal, independently sustaining itself and easily dominating the circuit's response error. As a result, the bypass selection generally focuses first upon stability and then upon improvements that further reduce noise coupling. Stability preservation requires bypass control of the parasitic feedback loop established by the supply-line impedance and the amplifier's PSRR coupling. Analysis of this feedback loop provides insight into the loop's origin and control and produces a mathematical expression of the condition required for oscillation.

As described with Fig. 8.1, this loop originates in the power-supply noise signal $2i_S Z_L$ that couples to the amplifier input in the error signal $e_{idS} = 2i_S Z_L / \text{PSRR}$. From there, the amplifier produces an output signal and a corresponding feedback error signal that actually supports e_{idS} . Figure 8.3 illustrates this feedback relationship with a modified version of the Fig. 8.1 circuit. As before, the circuit here neglects the input signal to focus the analysis upon signals related to the PSRR error. Modifications here alter the power-supply connection and the input error signal. For the supply connection, the amplifier load current i_L replaces the previous i_S , and the V_+ connection alone illustrates the supply coupling effect. Only the amplifier's i_L portion of i_S influences stability because only this portion produces a feedback relationship with the amplifier. Also, the coupling effects described here for the V_+ connection apply to those that occur with the V_- path as well. Next, a change in the input error signal shown reflects the effect of amplifier gain error. In the figure, the e_o/A gain error term appears in the more general input error

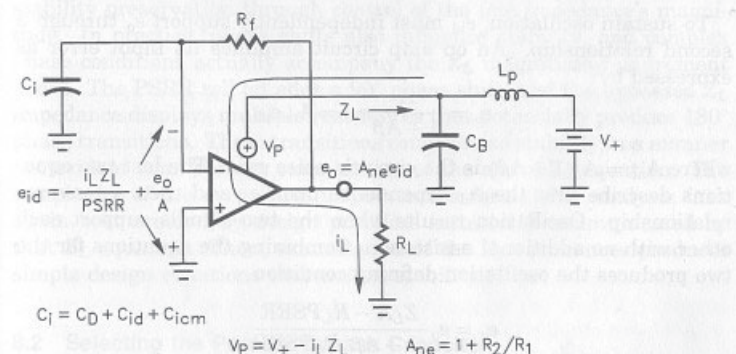


Figure 8.3 Load current i_L , drawn by the op amp, reacts with the supply-line inductance to produce an input error signal that in turn produces an output voltage and potential oscillation.

signal e_{id} , instead of the previous e_{idS} , to include the parasitic feedback mechanism.

8.1.4 The oscillation condition

Tracing a signal through the parasitic feedback loop illustrates both the feedback mechanism and its potential for oscillation. An intuitive evaluation illustrates the mechanism, and then a quantitative analysis defines the oscillation condition. In the figure, an e_o output voltage from the amplifier supplies a load current i_L to resistor R_L . The amplifier draws this current from V_+ and through the Z_L impedance of its supply line. The resulting line voltage drop produces a component of the e_{id} error signal, $-i_L Z_L / \text{PSRR}$, through the amplifier's finite PSRR. Then, the circuit amplifies this component by the amplifier noise gain A_{ne} , producing an e_o output response. That response reflects back to the amplifier inputs through the amplifier's open-loop gain A , producing the e_o/A component of the e_{id} shown. Thus, the power-supply coupling produces an input signal, which in turn produces an output signal that then produces an input signal. This simple link describes the full circle of a feedback loop capable of sustaining oscillation.

Quantifying this analysis produces two equations that describe the amplifier's feedback response and express the fundamental condition required for oscillation. First, the direct and secondary effects of power-supply coupling produce the input error signal $e_{id} = -i_L Z_L / \text{PSRR} + e_o / A$ as described above. Substituting $i_L = e_o / R_L$ removes the i_L variable for

$$e_{id} = \left(-\frac{Z_L}{R_L \text{PSRR}} + \frac{1}{A} \right) e_o$$

To sustain oscillation, e_{id} must independently support e_o through a second relationship. An op amp circuit amplifies its input error as expressed by²

$$e_o = \frac{A}{1 + A\beta} e_{id} = A_{ne} e_{id}$$

where $A_{ne} = A/(1 + A\beta)$ is the circuit's noise gain. The last two equations describe first the e_{id} dependence upon e_o and then a converse relationship. Oscillation results when the two signals support each other with no additional assistance. Combining the equations for the two produces the oscillation defining condition

$$e_o = e_o \frac{Z_L A - R_L \text{PSRR}}{(1 + A\beta) R_L \text{PSRR}}$$

While too complex for intuitive evaluation, this last equation quickly yields two simplified results. With e_o on both sides of the equation,

only two solutions satisfy the equality expressed. First, $e_o = 0$ balances the equation, indicating the stable state with no self-sustaining oscillation signal. Second, dividing both sides of the equation by e_o and solving for Z_L defines the line impedance required to produce the oscillation state

$$Z_L = \frac{R_L \text{PSRR}}{A/(2 + A\beta)} \approx \frac{R_L \text{PSRR}}{A_{ne}}$$

Preventing oscillation requires avoiding conditions that balance this new equality. Keeping Z_L low avoids oscillation as long as

$$Z_L < \frac{R_L \text{PSRR}}{A_{ne}}$$

This last Z_L equation offers several insights into the bypass requirement and the related parameters that control stability. Variations in three parameters of the equation influence the Z_L requirement. At lower frequencies, bypass capacitors fail to reduce Z_L , increasing the value of the left side of the equation. However, at these low frequencies, PSRR remains high, increasing the right side and preserving the equation's $<$ condition. At higher frequencies, PSRR declines, decreasing the value of the right side. However, bypass capacitors then reduce Z_L to retain the stability condition. At all frequencies, the value of load resistance R_L influences stability. As expressed above, lower values of R_L require correspondingly lower values of Z_L . There, lower R_L values produce greater supply-current drains that increase the supply-voltage drop developed on the Z_L impedance.

The above analysis provides intuitive but simplified guidelines to stability preservation through control of the line impedance's magnitude. In practice, phase shifts also influence stability, and complex phase conditions actually accompany the Z_L magnitude requirement above. The PSRR roll off adds a 90° phase shift, and the bypassed Z_L impedance displays multiple resonances that potentially produce 180° phase transitions. These transitions compromise stability in a manner not evident from cursory examination of the above Z_L equation. The complexity of the overall result presents a formidable analysis task again best avoided by design practices intuitively communicated by that Z_L equation. Then, examination of the Z_L resonances produces simple design equations that ensure stability.

8.2 Selecting the Primary Bypass Capacitor

As described above, supply bypass that maintains low line impedance limits supply noise coupling and preserves frequency stability. This

suggests making the bypass capacitance large to minimize the resulting $Z_L = 1/C_B s$ impedance. However, power-supply bypass does not simply replace the $Z_L = L_p s$ impedance with $Z_L = 1/C_B s$ as previously approximated. This approximation fails due to a variety of resonances formed with parasitic inductances. The basic power-supply bypass produces two such L - C resonances. Just the line inductance and the basic bypass capacitor produce the first, and the capacitor itself produces the second. The first one must be placed within the useful frequency range of the op amp in order to realize the bypass benefit. However, the second one can often be placed outside this range by avoiding an excessively large capacitance value. Larger capacitors have greater internal inductances that move this second resonance to lower frequencies. A compromise solution produces a design equation that adequately reduces line impedance but keeps this second resonance at a higher frequency. Later, the addition of a secondary bypass capacitor contends with the second resonance for higher-frequency amplifiers that cannot escape it.

8.2.1 Bypass resonances

The first results from the reaction of C_B with L_p and occurs well within the frequency range of the amplifier. In Fig. 8.3, C_B effectively appears in parallel with the L_p supply-line inductance. There, the ideal zero impedance of V_+ returns L_p to ac ground, completing this parallel connection. Then, C_B and L_p form the classic L - C tank circuit with an impedance of

$$Z_L = \frac{L_p s}{1 + L_p C_B s^2}$$

At low frequencies, $Z_L = L_p s$, and at high frequencies, $Z_L = 1/C_B s$ as previously approximated. However, at an intermediate frequency, this impedance displays a resonance maximum at the frequency $f_{rL} = 1/2\pi\sqrt{L_p C_B}$. There, the Z_L line impedance approaches infinity, totally counter to the impedance reduction intended by the power-supply bypass.

Fortunately, a parasitic line resistance R_p and the amplifier's PSRR contend with this resonance. The R_p resistance dissipates the resonant energy of the L - C tank, detuning it to dramatically restrain the resulting impedance rise. A typical printed circuit trace introduces a parasitic 12 m Ω per inch in its connection between the power supply and the amplifier. In addition, choosing C_B to place this resonance at a lower frequency where PSRR remains high lets the amplifier attenuate the associated coupling effect. Figure 8.4 illustrates the frequency response of this Z_L impedance along with the impedance's

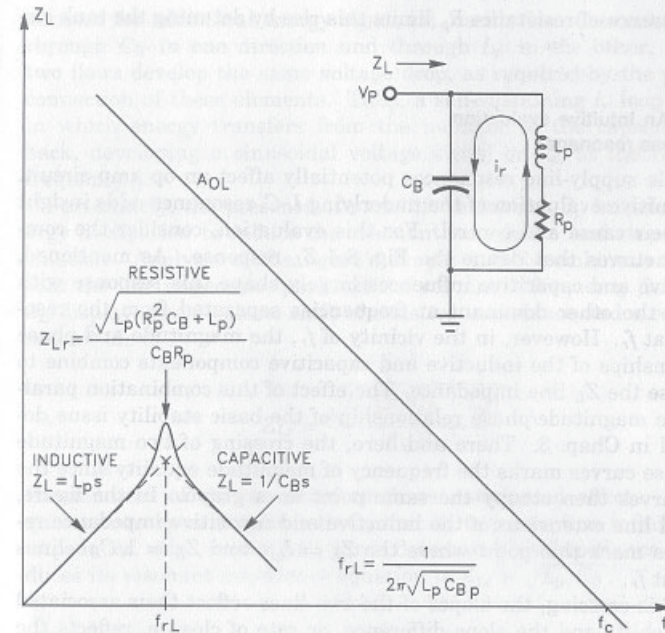


Figure 8.4 Together, the power-supply line inductance and a bypass capacitance form an L - C tank circuit with a resonant impedance detuned only by the presence of parasitic line resistance R_p .

circuit model. Superimposed on the figure, the amplifier's A_{OL} response provides a relative frequency comparison. In the model, bypass capacitor C_B shunts the series combination of the parasitic line inductance L_p and the parasitic line resistance R_p . This model assumes the zero output impedance of an ideal power supply and returns the L_p , R_p combination to ground. Then, the circuit model represents the Z_L line impedance seen from the V_p terminal of the amplifier's positive supply connection.

As shown by the Z_L response plot, the line impedance varies from inductive to resistive to capacitive as frequency increases. At lower frequencies, line inductance L_p controls Z_L , producing a rising impedance curve. At higher frequencies, bypass capacitor C_B takes control, producing the declining Z_L curve desired. Between these two frequency ranges, the inductive and capacitive response curves cross at their resonant frequency f_r . There, Z_L rises to a peak value and actually increases because of the presence of bypass capacitor C_B . Fortunately,

the presence of resistance R_p limits this rise by detuning the tank circuit.

8.2.2 An intuitive evaluation of bypass resonance

Multiple supply-line resonances potentially affect an op amp circuit. An intuitive evaluation of the underlying L - C resonance adds insight into their cause and control. For this evaluation, consider the component curves that define the Fig. 8.4 Z_L response. As mentioned, inductive and capacitive influences largely shape this response with one or the other dominant at frequencies separated from the resonance at f_r . However, in the vicinity of f_r , the magnitude and phase relationships of the inductive and capacitive components combine to increase the Z_L line impedance. The effect of this combination parallels the magnitude/phase relationship of the basic stability issue described in Chap. 3. There and here, the crossing of two magnitude response curves marks the frequency of magnitude equality since the two curves then occupy the same point on a graph. In the figure, dashed line extensions of the inductive and capacitive impedance responses mark this point where the $Z_L = L_p s$ and $Z_L = 1/C_B s$ lines cross at f_r .

At this crossing, the slopes of the two lines reflect their associated phase shifts and the slope difference, or rate of closure, reflects the corresponding phase difference. At f_r , this phase difference turns this equal magnitude point into a resonance. The single-zero rising response of the $L_p s$ component reflects a 90° phase lead, and the single-pole declining response of the $1/C_B s$ component reflects a 90° phase lag. This combination produces a 180° phase difference between the impedances of the circuit model's L_p and C_B components. For the ideal L - C tank circuit, this magnitude and phase combination produces an infinite impedance, and any supply of energy to the tank produces oscillation. Simply turning on the power supply provides this energy through the transient current drawn from the supply. After that, the ideal tank's infinite impedance successfully sustains an oscillation signal even after termination of energy input to the circuit.

Initially neglecting R_p of the Fig. 8.4 circuit model illustrates the resonant condition. Then, the parallel connected L_p and C_B support the same voltage drop. At the resonance, the equal impedance magnitudes of L_p and C_B develop currents of equal magnitudes from their common voltage drop. Yet, the 180° phase difference for the two impedances makes the two currents out of phase or of opposite polarity. This produces the circulating resonant current i_r shown with

no need for additional energy input, i.e., oscillation. Current i_r flows through C_B in one direction and through L_p in the other, yet the two flows develop the same voltage drop, as required by the parallel connection of these elements. Thus, a self-sustaining i_r loop results in which energy transfers from the inductor to the capacitor and back, developing a sinusoidal voltage signal on Z_L at the resonant frequency f_r .

Fortunately, the presence of parasitic resistance R_p introduces energy dissipation to reduce the resonant impedance and stop the oscillation. Analysis of the figure's simple R - L - C combination shows that the line impedance peaks at the resonance frequency f_r . There, Z_L displays the 0° phase shift of resistive impedance and a resonant magnitude of

$$Z_{Lr} = \frac{\sqrt{L_p (R_p^2 C_B + L_p)}}{C_B R_p}$$

Examination of this equation confirms anticipated effects. Z_{Lr} requires nonzero values for both C_B and R_p to prevent an infinite peak impedance. Also, making C_B large decreases this impedance and reduces its resonant impedance equation to $Z_{Lr} = \sqrt{L_p / C_B}$.

8.2.3 The bypass selection

Control of the above resonance and an empirical guideline produce the first two design equations for the bypass selection. The first equation places the resonance at a frequency where a higher-amplifier PSRR better contends with the associated impedance rise. The second equation sets the magnitude of the bypassed line impedance with a C_B design equation. Later, control of other resonances completes the process with secondary bypass capacitor selection.

At first, the impedance increase of the Z_{Lr} resonant condition would suggest moving f_{rL} to a frequency above the normal operating frequency range of the op amp, f_c . Here, f_c represents the unity-gain crossover frequency of the op amp in Fig. 8.4 and generally represents the upper limit of the amplifier's useful frequency range. Above f_c , the lack of amplifier gain restricts the parasitic feedback resulting from the resonance. However, making $f_{rL} > f_c$ would entirely sacrifice the benefit of the supply bypass. Preceding f_{rL} in Fig. 8.4, the Z_L impedance rises with frequency, unaffected by the bypass. The Z_L impedance only reflects the desired decline after the f_{rL} resonance where the capacitive roll off dominates Z_L . Thus, the f_{rL} resonance

must be endured within the op amp's operating frequency range in order to benefit from the higher-frequency effects of the bypass.

Moving this bypass resonance to a frequency $f_{rL} \ll f_c$ actually provides the best compromise. This places the resonance at a frequency where the amplifier retains sufficient PSRR to attenuate the coupling effect of the resonance. Further, the $f_{rL} \ll f_c$ placement moves the C_B roll off of the line impedance back into the amplifier's useful frequency range. Then, for $f_{rL} = 1/2\pi\sqrt{L_p C_B} \ll f_c$ the C_B capacitance value must satisfy the limit equation

$$C_B \gg \frac{1}{4L_p (\pi f_c)^2}$$

For a typical case, $L_p \approx 300$ nH and $f_c \approx 2$ MHz, producing $C_B \gg 0.02$ μ F. Generally, the line impedance control described next automatically satisfies this C_B condition.

Making $f_{rL} \ll f_c$ also reduces the analysis of other bypass conditions to the impedance of the capacitor alone. Without bypass, the line impedance actually includes a complex combination of inductances and resistances introduced by wire, printed circuit traces, and connectors. All of these affect the lower-frequency response of the Z_L impedance, discouraging any detailed analysis. However, making $f_{rL} \ll f_c$ reduces the Z_L impedance to just that of the capacitor in the critical higher-frequency range. There, a declining PSRR most requires a low Z_L to limit supply-line coupling. Above the f_{rL} resonance Z_L then becomes $Z_L = Z_{CB} = 1/C_B s$ independent of the preceding line impedance complexity.

Then, for most op amps, simply reducing the C_B impedance to about 1Ω well before f_c effectively counteracts the line impedance effects. This reduces the previous stability condition to $Z_L = Z_{CB} = 1 < R_L \text{PSRR}/A_{ne}$ or $A_{ne} < R_L \text{PSRR}$. Here, a noise gain as large as 1000 and a load resistance as small as $1 \text{ k}\Omega$ permit a PSRR as low as unity before oscillation results. Rarely would such a low PSRR occur where the noise gain remained this high, so the $Z_{CB} = 1 \Omega$ guideline assures stability for virtually every case. Setting $Z_{CB} = 1/2\pi f C_B = 1 \Omega$ produces $C_B = 1/2\pi f$. Then, setting $f = f_c/100$ for this $1\text{-}\Omega$ condition produces a general design equation for C_B

$$C_B = \frac{50}{\pi f_c}$$

For a general-purpose op amp with $f_c \approx 2$ MHz, this equation defines a capacitance value of $C_B \approx 7.5$ μ F. Retaining this $1\text{-}\Omega$ impedance up to and beyond f_c requires additional bypass attention as described below.

8.3 Selecting a Secondary Bypass Capacitor

Capacitor self-resonance potentially compromises the bypass described above, requiring the addition of secondary bypass capacitors. For general-purpose op amps, the above bypass selection generally places this self-resonance at a frequency beyond the amplifier's response range. Higher-frequency amplifiers encompass this frequency and may require the additional bypass. Still, even those capacitors exhibit self-resonances and require careful selection through a compromise that again produces simple design equations.

8.3.1 Bypass capacitor self-resonance

The inherent parasitic inductances and resistances of capacitors also affect the bypass effectiveness. The inductances introduce new resonances, and the resistances limit the line impedance reduction. A first new resonance results just from the capacitance and inductance of the primary bypass capacitor in a self-resonance condition. At this resonance, the bypass impedance would drop to zero except for the presence of the parasitic resistance. Above that frequency, the capacitor's parasitic inductance overrides the intended capacitance in the control of the bypass impedance.

Making $f_{rL} \ll f_c$ in Sec. 8.2 above typically requires a fairly large capacitance value in the multiple microfarad range. Capacitors this large introduce a new resonance condition that compromises the bypass effectiveness at frequencies typically within the amplifier's response range. Such cases sometimes require the addition of a secondary, smaller bypass capacitor as described later. The new resonance results from the parasitic inductance L_{BP} of the bypass capacitor reacting with the intended capacitance. This resonance reverses the Z_{CB} response slope as shown in Fig. 8.5. All capacitors possess this internal inductance, with a value corresponding largely to the lengths of the capacitor's external and internal connecting paths. Minimizing the total connecting length minimizes the parasitic inductance and maximizes the frequency of the self-resonance. First, just minimizing the capacitor lead length, including that of associated circuit board traces, addresses the external path length. Then, capacitor selection minimizes the internal path component as primarily determined by the capacitor's construction material. For a given capacitance value, capacitor materials with higher capacitance densities reduce this length. Tantalum, NPO ceramic, and mica capacitors offer the lower inductance alternatives for high, medium, and low capacitance values, respectively.

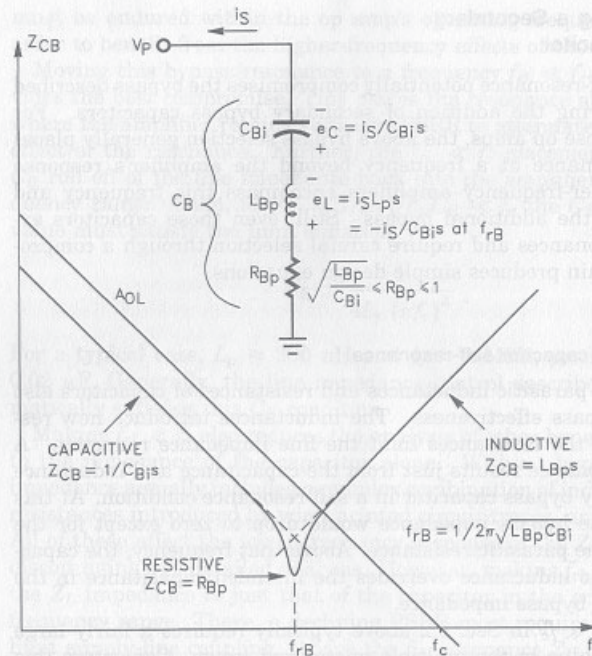


Figure 8.5 The intended impedance decline of a bypass capacitor reverses when the capacitor's internal inductance takes control at higher frequencies.

More detailed examination of the bypass capacitor's actual impedance components provides insight into its nonideal behavior and into the means to control the bypass result. As modeled in Fig. 8.5, the capacitor's parasitic inductance appears in series with the intended capacitance along with a parasitic resistance. There, L_{Bp} produces the self-resonance with C_B 's ideal capacitance, C_{Bi} . Parasitic resistance R_{Bp} sets the capacitor's impedance at the resonance. This resistance results from the same connecting path that produces the capacitor's inductance. Supply-current drain i_S develops voltages across each of the model's impedances with the capacitive and inductive impedances producing canceling effects at their resonance. Current i_S produces the voltage drops $e_C = i_S / C_{Bi} s$ across C_{Bi} , and $e_L = i_S L_{Bp} s$ across L_{Bp} . At the resonance, $s = s_{rB} = j\omega_{rB} = j / \sqrt{L_{Bp} C_{Bi}}$. Substituting this result for s in the preceding voltage drop equations yields $e_{Cr} = -j\sqrt{L_{Bp}/C_{Bi}}$ and $e_{Lr} = j\sqrt{L_{Bp}/C_{Bi}}$. Thus, at the resonance, $e_{Cr} = -e_{Lr}$ and the

two voltage drops cancel. Except for the presence of R_{Bp} , this voltage cancellation would produce zero voltage drop in response to the i_S current at the resonant frequency, indicating the ideal zero bypass impedance at the resonance.

However, parasitic resistance R_{Bp} produces a third voltage drop with i_S and limits the impedance decline to the level of this resistance. At first, this limit would seem undesirable, but the parasitic resistance also detunes the C_{Bi} , L_{Bp} resonance, taming a dramatic phase transition in the line impedance. That transition presents a broad range of phase conditions that could degrade stability. In general, this effect rarely produces a significant stability disturbance because it occurs at a frequency where amplifier PSRR remains relatively high. However, the impedance rise following the resonance may well degrade stability and require dual bypass. For the general case, the beneficial effect of the capacitor's parasitic resistance provides adequate stability assurance and eases the capacitor selection. Accepting a capacitor with higher parasitic resistance actually smoothes the phase transition of the resonance.

This resistance benefits performance as long as it detunes the resonance but does not raise the line impedance above the 1- Ω guideline level. For minimum detuning, the impedances of the R , L , and C components of the tank circuit should be equal at the resonant frequency. At this frequency, the impedance magnitudes of C_{Bi} and L_{Bp} automatically equate as shown in Fig. 8.5. There, dashed-line extensions of the two impedance curves identify equal magnitudes at their f_{rB} crossing. At this crossing, $Z_{CBi} = 1/2\pi f_{rB} C_{Bi}$, where $f_{rB} = 1/2\pi \sqrt{L_{Bp} C_{Bi}}$. Setting R_{Bp} equal to the Z_{CBi} impedance and solving produces the minimum resistance required for detuning, $R_{Bp} = \sqrt{L_{Bp}/C_{Bi}}$. Combining this minimum with the 1- Ω maximum produces the design limit equation

$$\sqrt{\frac{L_{Bp}}{C_{Bi}}} \leq R_{Bp} \leq 1$$

For the typical case, $L_{Bp} = 15$ nH and $C_{Bi} = 7.5$ μ F, making the limit condition an easily met $0.04 \leq R_{Bp} \leq 1$. Later, dual bypass conditions raise this resistance limit to counter yet another resonance.

Op amps having lesser signal bandwidths generally assure this detuned condition in the bypass capacitor selection. Such amplifiers require larger C_B values for the previous C_B design equation, directing the capacitor selection to tantalum types. Such capacitors display larger R_{Bp} values that flatten the Fig. 8.5 transition between its capacitive and inductive slopes. However, for higher-frequency amplifiers this guideline would permit use of ceramic capacitors which display much lower R_{Bp} values. Then, replacing an initial ceramic

choice with tantalum could solve occasional, unexpected stability problems. Later, a resistive detuning evaluation explores this option further.

Following the f_{rB} resonance, the Z_{CB} impedance curve rises in Fig. 8.5, again threatening stability. However, the C_B selection guidelines described above typically control the effect of this rise for general-purpose op amps. There, the avoidance of excessively large capacitance values and the limited amplifier bandwidths reduce circuit susceptibility to the impedance increase. Choosing $C_B = 50/\pi f_c$ assures adequate line impedance control at lower frequencies without introducing the excess inductance of a larger capacitor. With this compromise the Z_{CB} impedance rise remains within the $1\text{-}\Omega$ limit over the useful bandwidth of general-purpose amplifiers. Success here also depends upon minimizing the connection length between the capacitor and the amplifier.

8.3.2 Dual bypass capacitors

The simple, single-capacitor bypass that serves general-purpose op amps fails with higher-frequency op amps. At higher frequencies, inductive effects become far more serious, requiring greater attention to frequency stability and supply noise coupling. The parasitic inductance of the larger-value, single bypass capacitor defeats the bypass purpose at higher frequencies, requiring the addition of a smaller secondary capacitor. Even general-purpose amplifiers may require dual bypass if PSRR is low.

In Fig. 8.5, the parasitic L_{Bp} inductance controls the bypass impedance Z_{CB} at frequencies above the capacitor's self-resonance. At higher frequencies, the rising L_{Bp} portion of Z_{CB} increases the bypass impedance, reducing the bypass effectiveness. This rise generally begins within the amplifier response range and, there, a declining PSRR makes the amplifier more vulnerable to the line impedance. For all but the highest-frequency applications, just adding a secondary bypass capacitor in parallel with the first restores a low bypass impedance for the full amplifier response range. This produces a net $C_B = C_{B1} + C_{B2}$, and making $C_{B2} \ll C_{B1}$ generally assures that the C_{B2} self-resonance occurs outside the amplifier's response range. Both the lower capacitance value of C_{B2} and the accompanying lower parasitic inductance produce this higher resonant frequency. For even higher-frequency applications, yet a third capacitor may be required. In either case, examination of the dual bypass case here illustrates the impedance characteristics that guide the bypass selection process.

Adding secondary bypass capacitors restores the declining frequency response of the Z_{CB} bypass impedance but with new complications.

This solution introduces two additional resonances: one from the self-resonance of the secondary capacitor and the other from the reaction of the second capacitance with the inductance of the first. Figure 8.6 illustrates the first new resonance with a circuit model and the corresponding impedance responses. In the model, both capacitors display parasitic inductances and resistances and all of these parasitics affect the response curves. In the graph, two curves of the figure display the resulting capacitor impedances as Z_{CB1} and Z_{CB2} . At lower frequencies, a declining Z_{CB1} provides the lower impedance bypass shunt. However, Z_{CB1} later resonates and begins to rise at f_{rB1} . At higher frequencies, Z_{CB2} bypasses this rise and initially restores a declining bypass impedance. Later, the self-resonance of C_{B2} at f_{rB2} again produces a rise but at a lower impedance level than provided by the Z_{CB1} curve. While the resonances remain inevitable, their careful placement optimizes the overall bypass effect. To aid this placement, a third curve of the graph shows the amplifier's open-loop gain response A_{OL} . Comparison of the graph's three responses guides the secondary capacitor selection in a manner that ensures the amplifier's frequency stability.

8.3.3 Dual bypass selection

Selection of the primary C_{B1} capacitor indirectly defines the C_{B2} capacitance value through impedance interaction. The fundamental bypass requirement defines C_{B1} as described before for the single bypass case: $C_{B1} = 50/\pi f_c$. This selection reduces the line impedance to $1\text{ }\Omega$ at a frequency well within the amplifier's response range. Then, the addition of C_{B2} bypasses the higher-frequency impedance rise produced by the L_{Bp1} inductance of C_{B1} . Selecting C_{B2} to limit this rise to the same $1\text{-}\Omega$ guideline level defines this secondary capacitor's value.

In Fig. 8.6, the net bypass impedance of the C_{B1} , C_{B2} parallel combination peaks at the f_{iB} intercept of the rising Z_{CB1} curve and the falling Z_{CB2} . For higher-frequency amplifiers this peak would typically occur within the amplifier's response range. Selecting a value for C_{B2} that limits this peak to $1\text{-}\Omega$ continues the stability preservation described before for the C_{B1} selection. At their f_{iB} intercept, the two Z_{CB} curves occupy the same point on the graph, making $Z_{CB2} = Z_{CB1}$. There, inductance controls $Z_{CB1} = 2\pi f_{iB} L_{Bp1}$ and capacitance controls $Z_{CB2} = 1/2\pi f_{iB} C_{B2}$. Equating the two impedance expressions defines the intercept frequency as

$$f_{iB} = 1/2\pi \sqrt{L_{Bp1} C_{B2}}$$

Thus, the parasitic inductance of C_{B1} and the ideal capacitance of C_{B2} combine to define the frequency of the bypass impedance peak.

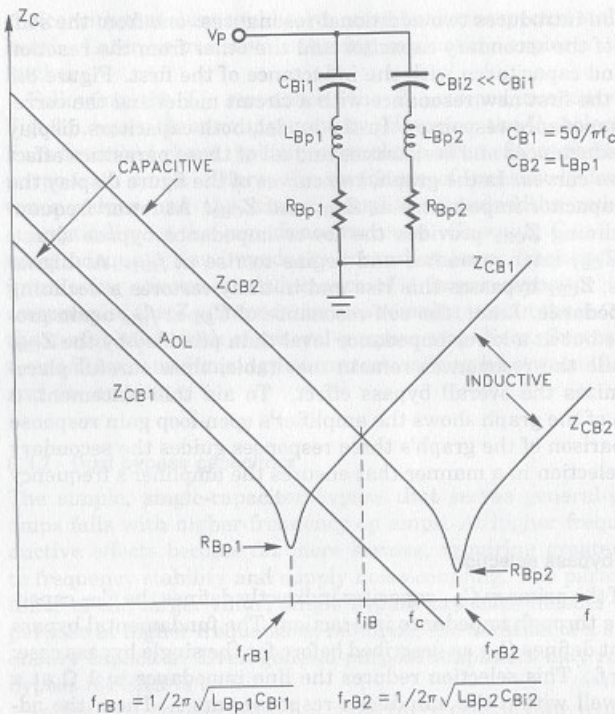


Figure 8.6 Adding a second bypass capacitor of smaller value rolls off the inductive impedance rise of the first.

Translating this f_{IB} expression into a C_{B2} design equation requires one further step. At f_{IB} , the Z_{CB2} curve follows its capacitive roll off as expressed by $Z_{CB2} = 1/2\pi f C_{B2}$. Setting $Z_{CB2} = 1 \Omega$ and $f = f_{IB}$ in this expression produces $f_{IB} = 1/2\pi C_{B2}$. Equating this result with the above f_{IB} expression yields the C_{B2} design equation

$$C_{B2} = L_{BP1}$$

Thus, simply making the magnitude of the C_{B2} capacitance equal to that of C_{B1} 's parasitic inductance transfers line impedance control from Z_{CB1} to Z_{CB2} at the 1- Ω level. In practice a resonance described later raises the peak impedance above 1 Ω at f_{IB} , but temporarily ignoring this effect yields the best starting point for the bypass selection. As described later, resistive detuning of this resonance or an

alternate capacitor type solves any related problem without altering the above design equation. Using this equation, a typical 1.5 nH of C_{B1} inductance prescribes a secondary bypass of $C_{B2} = 1.5$ nF.

The above C_{B2} equation requires knowing L_{BP1} , and impedance measurement provides the most reliable determination of this inductance. Capacitor data sheets rarely specify this parasitic. Some data sheets communicate this inductance through an impedance-versus-frequency curve. There, the higher-frequency, rising portion of this curve yields the parasitic inductance through the simple calculation $L_p = Z_c/2\pi f$. Unfortunately, even these curves can fail to communicate the actual inductance condition of practical applications. For packaged capacitors, this resonance includes the parasitic inductance of package leads that are largely removed upon installation. Chip capacitors avoid this lead effect, but their response curves still do not include the effect of the circuit board traces that connect the capacitor. Fortunately, today's impedance analyzers accurately measure the frequency response of a given capacitor and its approximated connection length. From such measurements, the rising portion of the impedance curve then accurately defines the actual inductance through the $L_p = Z_c/2\pi f$ relationship.

8.4 Bypass Alternatives

The preceding sections outline the bypass selection that adequately serves most applications. However, three special conditions may require additional attention to supply-line coupling. For these conditions, a third bypass capacitor, filtering, or resistive detuning further reduce the coupling effects. The third capacitor alternative extends the frequency range of bypass control for even higher-frequency amplifiers. There, the amplifier's A_{OL} response can encompass the self-resonances of both C_{B1} and C_{B2} . Then, adding a third, and even smaller, bypass capacitor rolls off the inductive impedance of C_{B2} , just as C_{B2} did for C_{B1} . Amplifiers with bandwidths exceeding 30 MHz may require this third capacitor.

A second alternative, filtering, removes supply-coupled signals outside the useful PSRR range of the amplifier. There, coresident circuitry may introduce high-frequency supply-line signals that couple through the amplifier with little PSRR attenuation. Beyond the PSRR's unity-gain crossover, such signals typically couple straight through the amplifier to its output. However, this PSRR crossover typically resides near the amplifier's A_{OL} crossover, presenting a filtering opportunity. The A_{OL} crossover marks the endpoint of the amplifier's useful frequency range. Thus, low-pass filtering, applied after the amplifier, can often remove the effect of high-frequency PSRR cou-

pling without restricting the useful bandwidth of the application. Still, intermodulation effects can circumvent the filtering as described later.

Resistive detuning of the bypass impedance offers a third alternative for the reduction of supply coupling effects. The dual bypass configuration potentially produces a critical resonance that degrades stability at surprising frequencies. This may occur at a frequency below or above the amplifier's response crossover at f_c . Below f_c , the C_{B2} selected above restores a line impedance below the 1- Ω guideline level, presumably preserving stable operation. However, the C_{B2} resonance with L_{Bp1} can raise the net line impedance well above this level, producing oscillation. In other cases, the resonance can even produce oscillation at frequencies above f_c . There, diminished amplifier gain limits the parasitic feedback loop and would seem to prevent instability. However, the resonant impedance rise can counteract this limit. When necessary, resistive degeneration detunes this new resonance, and an appropriate capacitor selection generally provides this through parasitic resistance.

8.4.1 Detuning the dual bypass resonance

An interaction between the two capacitors of the dual bypass configuration develops this new resonance. There, the secondary capacitance reacts with the inductance of the primary in another L - C tank configuration. Figure 8.7 illustrates this effect with the bold Z_{CB} response curve representing the net impedance of the parallel-connected bypass capacitors. There, Z_{CB} first follows the Z_{CB1} curve of C_{B1} at lower frequencies and later follows the Z_{CB2} curve of C_{B2} . The transition between the two occurs at their f_{iB} intercept where the bypass combination produces a resonance of its own. This resonance increases the Z_{CB} magnitude as shown and produces a 180° transition in the phase of Z_{CB} . The magnitudes and phases of the amplifier's PSRR and A_{OL} responses combine with these Z_{CB} characteristics to produce a complex parasitic feedback condition.

Some combinations of application conditions make the magnitude and/or phase of this resonance degrade stability. In such cases, attention to the more conventional stability determinant, the amplifier's $1/\beta$ intercept, fails to identify the problem. Similarly, adding a third bypass capacitor produces no improvement. The third capacitor only reduces the line impedance at frequencies above this interaction resonance. For such cases, bypass degeneration moderates the PSRR error signal by reducing both the magnitude and phase shift of the supply-line impedance. Careful selection of the degeneration resistance improves stability without increasing the supply-line impedance in the critical higher-frequency ranges.

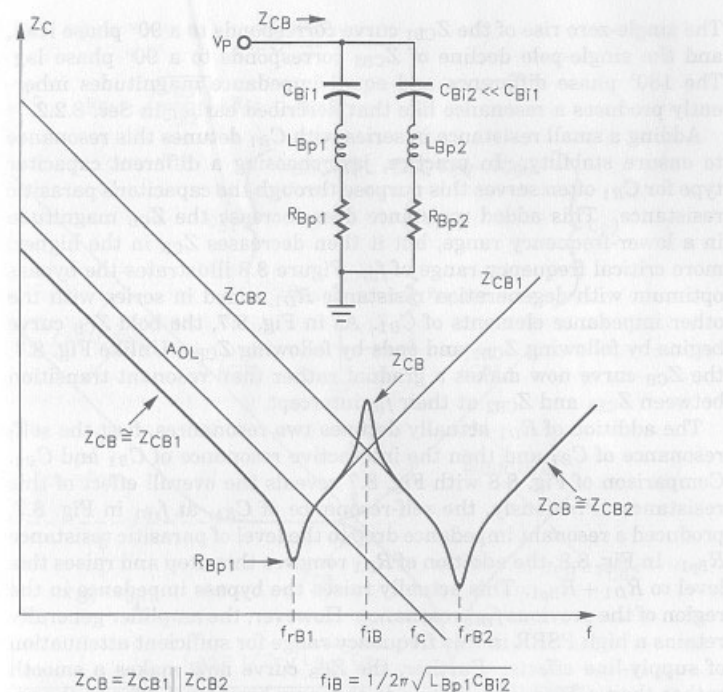


Figure 8.7 Dual bypass capacitors introduce a new resonance at their f_{iB} intercept where C_{B2} reacts with L_{Bp1} , increasing the net Z_{CB} impedance.

Graphical evaluation of this f_{iB} resonance reveals its underlying causes and displays the cure provided by the resistive degeneration. The resonance results from the reaction of C_{B1} 's parasitic inductance with C_{B2} 's capacitance. At f_{iB} in Fig. 8.7, Z_{CB1} follows a rising impedance response as controlled by L_{Bp1} , and Z_{CB2} follows a declining response as controlled by C_{B2} . At the crossing of the two responses, resonance results due to the equal magnitudes and opposite phases of the Z_{CB1} and Z_{CB2} impedances. This crossing or intercept of the two curves marks the point of equal impedance magnitudes since both curves occupy the same point in the graph. There, neither impedance dominates the parallel combination, in contrast to the impedance conditions at points separated from f_{iB} . At this intercept, the equal magnitudes of Z_{CB1} and Z_{CB2} contribute equally to the combined Z_{CB} impedance, making the phase shifts of the two also equally significant.

The single-zero rise of the Z_{CB1} curve corresponds to a 90° phase lead, and the single-pole decline of Z_{CB2} corresponds to a 90° phase lag. The 180° phase difference and equal impedance magnitudes inherently produces a resonance like that described earlier in Sec. 8.2.2.

Adding a small resistance in series with C_{B1} detunes this resonance to ensure stability. In practice, just choosing a different capacitor type for C_{B1} often serves this purpose through the capacitor's parasitic resistance. This added resistance does increase the Z_{CB} magnitude in a lower-frequency range, but it then decreases Z_{CB} in the higher, more critical frequency range of f_{iB} . Figure 8.8 illustrates the bypass optimum with degeneration resistance R_{D1} added in series with the other impedance elements of C_{B1} . As in Fig. 8.7, the bold Z_{CB} curve begins by following Z_{CB1} and ends by following Z_{CB2} . Unlike Fig. 8.7, the Z_{CB} curve now makes a gradual rather than resonant transition between Z_{CB1} and Z_{CB2} at their f_{iB} intercept.

The addition of R_{D1} actually detunes two resonances, first the self-resonance of C_{B1} and then the interactive resonance of C_{B1} and C_{B2} . Comparison of Fig. 8.8 with Fig. 8.7 reveals the overall effect of this resistance. Previously, the self-resonance of C_{B1} , at f_{rB1} in Fig. 8.7, produced a resonant impedance drop to the level of parasitic resistance R_{Bp1} . In Fig. 8.8, the addition of R_{D1} removes this drop and raises this level to $R_{D1} + R_{Bp1}$. This actually raises the bypass impedance in the region of the previous f_{rB1} resonance. However, the amplifier generally retains a high PSRR in this frequency range for sufficient attenuation of supply-line effects. Further, the Z_{CB} curve now makes a smooth rather than resonant transition to this new limit level. There, the reduced Z_{CB} response slope indicates a greatly reduced phase transition at the frequency of the previous f_{rB1} resonance. This reduces the potential phase combinations with amplifier gain and PSRR effects that could degrade stability. Accepting this temporary Z_{CB} increase allows R_{D1} to detune the more critical resonance at f_{iB} . There, the Z_{CB} curve no longer displays a resonant peak, indicating a greatly reduced Z_{CB} magnitude and phase transition at f_{iB} .

8.4.2 Selecting the detuning resistance

Detuning resistance R_{D1} should be large enough to prevent the resonant impedance rise at f_{iB} but not so large as to unnecessarily raise the Z_{CB} impedance at lower frequencies. Graphical analysis of Fig. 8.8 reveals this optimum condition and defines an R_{D1} design equation. The curves of the figure identify this optimum through the magnitudes and slopes of the impedance responses.

Detuning the Fig. 8.7 resonance at f_{iB} requires reducing the slope difference, or rate of closure, of the Z_{CB1} and Z_{CB2} curves at their f_{iB} in-

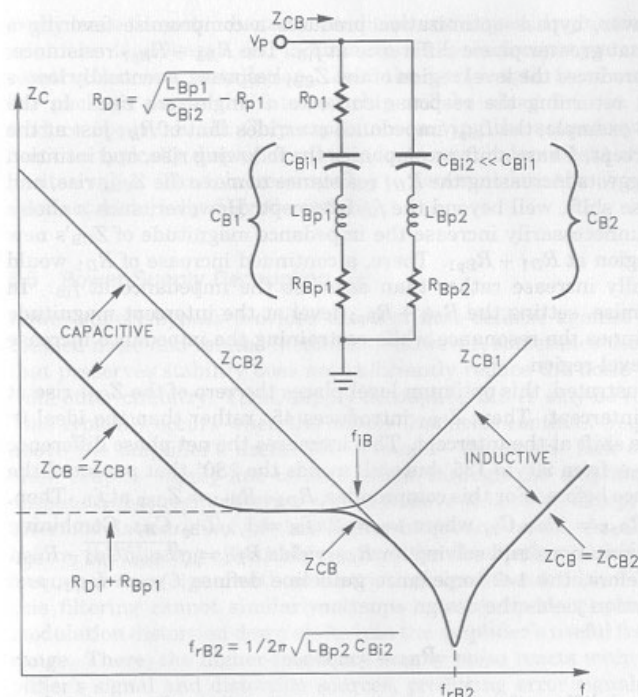


Figure 8.8 Adding resistance R_{D1} in series with C_{B1} detunes two of the Fig. 8.7 resonances at the expense of an elevated Z_{CB} level at lower frequencies.

tercept. This rate-of-closure criteria follows from the Chap. 3 stability analysis where the slopes of intersecting response curves reflect phase differences and the potential for circuit oscillation. Here, the slopes of the Z_{CB1} and Z_{CB2} curves similarly reflect phase differences and the potential for resonance. Adding R_{D1} decreases this slope difference at the f_{iB} intercept of Fig. 8.8, detuning the resonance. This added resistor produces the level zero-slope portion of the Z_{CB1} response, and this slope reflects zero phase shift. At the f_{iB} intercept, Z_{CB2} still follows its single-pole roll off, reflecting a 90° phase lag for a net 90° phase difference at the intercept. This greatly reduces the phase difference from the 180° required for resonance and would absolutely remove the resonant reaction of Z_{CB1} and Z_{CB2} .

However, bypass optimization produces a compromise favoring a somewhat greater phase difference at f_{iB} . The $R_{D1} + R_{Bp1}$ resistance, which produces the level region of the Z_{CB1} response, eventually loses control, returning the response curve to a single-zero rise. In the Fig. 8.8 example, the L_{Bp1} impedance overrides that of R_{D1} just at the f_{iB} intercept. Phase shift accompanies the following rise, and intuition first suggests increasing the R_{D1} resistance to move the Z_{CB1} rise, and its phase shift, well beyond the f_{iB} intercept. However, such a choice would unnecessarily increase the impedance magnitude of Z_{CB} 's new level region at $R_{D1} + R_{Bp1}$. There, a continued increase of R_{D1} would eventually increase rather than decrease the impedance at f_{iB} . In compromise, setting the $R_{D1} + R_{Bp1}$ level at the intercept magnitude still detunes the resonance while restraining the impedance increase of the level region.

As illustrated, this optimum level places the zero of the Z_{CB1} rise at the f_{iB} intercept. Then, Z_{CB1} introduces 45° rather than the ideal 0° of phase shift at the intercept. This increases the net phase difference with Z_{CB2} from 90° to 135° but still avoids the 180° that produced the resonance before. For this compromise, $R_{D1} + R_{Bp1} = Z_{CB2}$ at f_{iB} . Then, $R_{D1} + R_{Bp1} = 1/\omega_{iB}C_2$, where $\omega_{iB} = 2\pi f_{iB} = 1/\sqrt{L_{Bp1}C_{B2}}$. Combining these expressions and solving for R_{D1} yields $R_{D1} = \sqrt{L_{Bp1}/C_{B2}} - R_{Bp1}$. From before, the $1-\Omega$ impedance guideline defines $C_{B2} = L_{Bp1}$, and substitution yields the design equation

$$R_{D1} = 1 - R_{Bp1}$$

Expressed differently, this equation prescribes $R_{D1} + R_{Bp1} = 1 \Omega$ for a total $1-\Omega$ resistance in the C_{B1} path. This repeats the $1-\Omega$ guideline underlying bypass selection.

Typically, the above R_{D1} equation prescribes difficult resistance values of a fraction of an ohm. Such resistors exist but can be expensive and more difficult to acquire. However, when required, only this solution removes the effects of the f_{iB} bypass resonance. Fortunately, just choosing a different capacitor for C_{B1} generally provides an adequate detuning resistance. Choosing a C_{B1}' with $R_{Bp1}' \approx 1 \Omega$ detunes the f_{iB} resonance but still avoids exceeding the $1-\Omega$ guideline. Fortunately, this resistance does not have to be precise to provide sufficient detuning of the f_{iB} resonance. Often, just switching C_{B1} from a ceramic type to tantalum serves this purpose.

Note that this C_{B1} switch also requires changing the secondary bypass value C_{B2} . A tantalum C_{B1}' substitutes an increased L_{Bp1}' for L_{Bp1} , requiring recomputation of the corresponding $C_{B2}' = L_{Bp1}'$. Increasing C_{B2} here theoretically increases its inductance and would degrade higher-frequency stability. However, practical limitations gen-

erally leave the net inductance condition unchanged. In practice, some minimum lead length and inductance always accompany a capacitor construction and installation. Just the lead length from an amplifier package pin to the internal chip will be about 0.05 in and introduce about 0.8 nH. Chip capacitors serving the secondary bypass function typically display similar inductances, independent of the capacitor value. Thus, increasing C_{B2} to accommodate the C_{B1} change does not materially degrade stability.

8.5 Power-Supply Decoupling

Power-supply bypass provides the best first defense against supply-coupled noise and related instability. However, sometimes the bypass that preserves stability does not sufficiently reduce the noise coupled from other circuitry. Then, supply decoupling filters may be required. This typically occurs when the supply-line noise contains frequencies above the amplifier's useful PSRR range. There, the lack of PSRR often couples supply-line noise straight through the amplifier to its output. Subsequent filtering would remove this noise except for the intermodulation effect of this higher-frequency noise. As described before, the filtering effectively removes the direct effect of this higher-frequency error signal without restricting signal bandwidth. However, this filtering cannot similarly remove noise components that intermodulation distortion down shifts into the amplifier's useful frequency range. There, the higher-frequency supply noise reacts with the amplifier's signal and distortion sources, producing error signals in the lower-frequency range of the amplifier.

8.5.1 Decoupling alternatives

Then, prevention supersedes cure in the reduction of supply-coupled noise. Power-supply decoupling filters prevent the higher-frequency signals from even reaching the amplifier. The simplest decoupling alternatives include $R-C$, $L-C$, or $R-L-C$ low-pass filters as illustrated in Fig. 8.9. There, three low-pass filter configurations represent decoupling circuits placed between a power-supply source V_+ and the amplifier's positive supply pin V_P . While shown as a single capacitor, the C_B there represents the combined capacitance of all capacitors used for the supply bypass of V_P . In practice, the negative supply line would have an identical filter, but the positive supply case shown here serves to illustrate the decoupling principles.

Each of the three filter alternatives adds a series impedance in the supply line to combine with the bypass capacitance C_B , forming a low-pass filter. Then, the high-frequency noise contained in V_+ drops

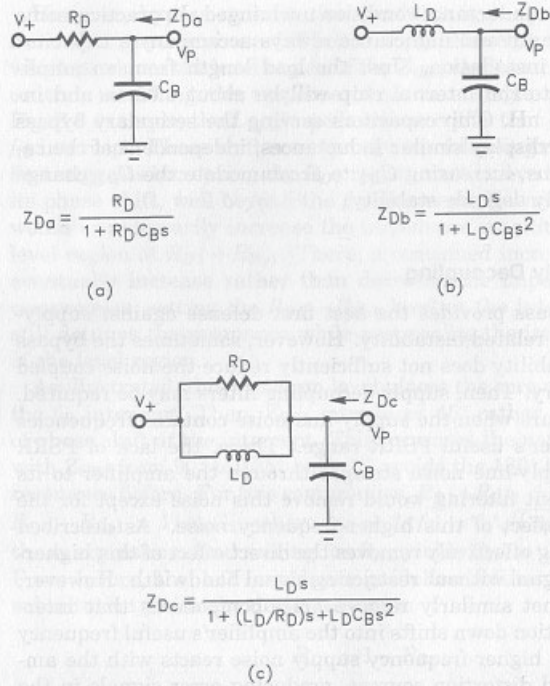


Figure 8.9 Power-supply decoupling introduces series impedance between the V_+ output of the power supply and the op amp's V_P terminal to intercept high-frequency supply transients introduced by other circuitry.

across the series impedance rather than transferring to the V_P supply pin of the op amp. The series impedance added does not have to be mounted close to the amplifier to provide the desired decoupling. However, this impedance should be placed so that it does not conduct the supply-current drain of circuitry other than the amplifier.

The type of impedance added determines the relative cost and decoupling effectiveness. This effectiveness depends both upon the filter response and the line impedance presented to the amplifier. The R - C solution of Fig. 8.9a offers the lowest cost and still produces appreciable decoupling results. With this circuit, decoupling resistance R_D absorbs and dissipates the higher-frequency, transient energy, transferring a more stable voltage to the C_B capacitance. Analysis of the decoupling circuit expresses this action in the classic R - C filter

response

$$V_P = \frac{V_+}{1 + R_D C_B s}$$

This response rolls off the supply-line noise with a single-pole at $1/2\pi R_D C_B$.

However, at lower frequencies, this simple R - C decoupling increases the supply-line impedance, potentially increasing the PSRR error. To define this impedance effect, assume that the V_+ power supply presents its ideal zero output impedance, placing R_D effectively in parallel with C_B . This makes the decoupling impedance seen from the V_P terminal

$$Z_{Da} = \frac{R_D}{1 + R_D C_B s}$$

Thus, at low frequencies the R - C filter increases line impedance by the amount R_D . To limit this effect, the R_D resistance must be kept low, and a later discussion describes its value selection.

Alternately, L - C filtering removes this impedance increase and produces a double- rather than single-pole roll off of supply-line noise. However, this alternative introduces another impedance increase through a supply-line resonance. Figure 8.9b illustrates this case with inductor L_D simply replacing the previous resistor R_D . Then, the L - C filter presents the low impedance of the inductor to lower-frequency current demands from the V_+ power supply. The filtering response of this alternative rolls off supply-line noise with a double pole at $f = 1/2\pi \sqrt{L_D C_B}$ as described by

$$V_P = \frac{V_+}{1 + L_D C_B s^2}$$

Compared with the R - C filter, the L - C filter reduces the supply-line impedance but only at lower frequencies. The impedance seen from the V_P terminal becomes

$$Z_{Db} = \frac{L_D s}{1 + L_D C_B s^2}$$

At lower frequencies, the s term of the Z_{Db} numerator assures low impedance, but this same term tends to increase Z_{Db} at higher frequencies. Counteracting this increase, the s^2 term of the denominator returns the net impedance response to a single-pole roll off like that attained with the R - C filter.

Thus, the L - C filter offers two-pole roll off of supply-line noise and reduced low-frequency line impedance. However, at an intermediate frequency, the Z_{Db} impedance increases due to a new L - C resonance.

From an energy perspective, the L_D inductor stores rather than dissipates the transient energy of supply-line noise, giving rise to this resonance. At most frequencies, the inductor simply releases its stored energy to the bypass capacitance and the amplifier in a more gradual manner than initially presented by a supply-noise transient. This smoothing of the energy release protects the amplifier from supply-line noise by reducing the associated frequency content to a range where amplifier PSRR remains high. However, at the inevitable L - C resonance the Z_{Db} impedance rises and the filtering fails. As indicated by the Z_{Db} denominator, $1 + L_D C_B s^2$, the L_D - C_B combination produces a double pole and a resonance at $f_{rDb} = 1/2\pi \sqrt{L_D C_B}$. This resonance increases rather than decreases the line impedance in the vicinity of f_{rDb} .

This resonance makes the simple L - C decoupling filter an uncertain alternative for op amp circuits having high noise gains like the photodiode amplifier. At resonance, energy stored in the inductor by supply-line transients circulates between the capacitor and inductor until dissipated by some energy drain. In the ideal L - C case, this circulation continues indefinitely, sustaining an oscillation signal. In practice, parasitic resistances and the amplifier help absorb that circulating energy, but the amplifier's PSRR error and gain offer replenishment to support this oscillation. There, supply-voltage transients or noise containing frequency components in the range of the f_{rDb} resonance can initiate this condition.

Adding a detuning resistor removes the resonance and the oscillation uncertainty but reduces the noise roll off. Shown in Fig. 8.9c, this alternative combines the R_D and L_D series impedances of the preceding two cases. The combination provides both the energy absorption of the R - C filter and the reduced low-frequency impedance of the L - C filter. For this R - L - C combination the filtering response becomes

$$V_P = \frac{[1 + (L_D/R_D)s]V_+}{1 + (L_D/R_D)s + L_D C_B s^2}$$

Here, the s^2 term of the denominator initially continues the two-pole roll off advantage of the basic L - C case. However, the s term of the numerator indicates a response zero that returns this roll off to single pole. Thus, adding R_D to the L - C filter reduces the filter attenuation.

Still, R_D improves decoupling effectiveness by detuning the line impedance resonance that otherwise adds gain to the parasitic PSRR feedback loop. Analysis of the Z_{Dc} impedance, as seen from the V_P terminal, yields

$$Z_{Dc} = \frac{L_D s}{1 + (L_D/R_D)s + L_D C_B s^2}$$

This expression repeats that of the previous Z_{Db} except for the $(L_D/R_D)s$ term of the denominator. That term separates the circuit's two poles to detune the characteristic L - C resonance and remove the associated impedance rise.

8.5.2 Selecting the decoupling components

Two compromises guide the selection of the decoupling filter components. The first repeats the bypass capacitance selection described before, optimizing the value of C_B to maintain frequency stability. Design equations developed earlier guide this selection. The second compromise guides the selection of the series impedance to limit noise coupling without significantly degrading other performance.

The series impedance, R_D or the R_D and L_D combination, must be large enough to intercept high-frequency line transients but not so large as to compromise the circuit's lower-frequency PSRR error signal. The power-supply decoupling reduces the supply-line noise produced by other circuitry but can increase the parasitic feedback produced by the amplifier itself. At lower frequencies, the bypass capacitors fail to roll off the line impedance, and adding the series impedance increases the supply-line impedance seen by the amplifier. The discussion of Sec. 8.1 describes the noise-coupling effect of this impedance. The net effect upon circuit performance remains a function of multiple variables and discourages analysis. However, following the previous bypass capacitor selection guidelines and then setting $R_D \sim 5 \Omega$ serves most amplifiers that supply output currents in the 10-mA range. This 5- Ω guideline, combined with the 1- Ω guideline for C_B , produces a filter with at least a 6:1 reduction of higher-frequency noise. Also, adding just 5 Ω generally retains lower-frequency line impedance at an acceptable level. Amplifiers supplying greater output current require correspondingly lower values for R_D . For R - C filters, this R_D selection completes the decoupling design.

For R - L - C filters, the L_D selection remains. Adding the L_D inductor of Fig. 8.9c benefits performance as long as the inductance chosen forms a detuned resonant circuit with the C_B and R_D chosen above. Making L_D larger minimizes the line impedance at lower frequencies but potentially increases it at the L - C resonance. For optimum detuning of this resonance, the L_D impedance should equal R_D at the frequency of the potential resonance. From the Fig. 8.9b discussion, the undamped resonance would occur at the frequency $f_{rDb} = 1/2\pi \sqrt{L_D C_B}$. There the impedance of L_D equals $Z_{LD} = 2\pi L_D f_r = \sqrt{L_D/C_B}$, and setting this equal to R_D yields

$$L_D = R_D^2 C_B$$

Here, capacitance C_B represents the combined value of the bypass capacitors connected to the associated supply terminal.

References

1. H. Ott, *Noise Reduction Techniques in Electronic Systems*, John Wiley, New York, 1976.
2. J. Graeme, "Feedback Models Reduce Op Amp Circuits to Voltage Dividers," *EDN*, June 20, 1991, p. 139.

Reducing External Noise Effects

Diminishing returns eventually limit the noise reduction achieved through measures focused upon the photodiode amplifier itself. External noise sources impose a background noise floor that requires attention to the amplifier's environment rather than the amplifier. This background noise typically results from the parasitic noise coupling of external electrostatic and magnetic sources. These sources can dominate noise performance, overriding any effort to further reduce the direct noise effects of the amplifier.

Limiting the effects of these external noise sources requires attention to amplifier location, shielding, the circuit structure, and the physical arrangement of circuit components.^{1,2,3} Electrostatic and magnetic noise signals enter the amplifier circuit through parasitic mechanisms. Between these sources and the amplifier, parasitic mutual capacitances couple electrostatic interference, and parasitic mutual inductances couple magnetic interference. For both sources, maximizing the physical separation of the amplifier from the source produces the greatest coupling reduction. Shielding provides the next greatest reduction but with different shield material requirements for electrostatic and magnetic coupling. Reduction of electrostatic coupling simply requires a shield material having high electrical conductivity to shunt capacitively coupled currents to ground. Reduction of magnetic coupling requires a shield material having high magnetic permeability or significant thickness to attenuate the magnetic field that can otherwise couple through a simple conductive shield. Fortunately, the higher frequencies that make magnetic coupling most effective also make simpler shield materials more effective.

Differential circuit structures and loop-minimizing component arrangements provide further reduction of coupled noise. Converting the photodiode amplifier to a differential-input form activates the CMR of the op amp for rejection of both electrostatic and magnetic coupling effects. However, limited amplifier response range limits this solution

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Differential circuit structures and loop-minimizing component arrangements provide further reduction of coupled noise. Converting the photodiode amplifier to a differential-input form activates the CMR of the op amp for rejection of both electrostatic and magnetic coupling effects. However, limited amplifier response range limits this solution

to lower frequencies. Component arrangements that minimize loop areas reduce magnetic coupling by minimizing the mutual inductances that couple the noise. Similarly, coaxial power-supply and load connections minimize higher current loops that could generate significant magnetic fields.

9.1 Reducing Electrostatic Coupling

Electric field coupling, such as from the power line, supplies noise signals through the mutual capacitances that exist between any two objects. The two objects serve as the capacitor's plates, and the intervening air or other material acts as the capacitor's dielectric layer. AC voltage differences between the objects drive these mutual capacitances, coupling noise currents from one to the other. Electrostatic shielding intercepts this noise current and shunts it to ground. Alternately, differential-input photodiode configurations turn the coupled signal into common-mode signals for removal by the op amp's CMR. In the simplest case, one op amp performs this differential function by utilizing the opposite polarity currents of the photodiode's two terminals. However, this configuration may limit bandwidth. A three-op-amp solution removes this limit and actually increases bandwidth over that achieved with the basic photodiode amplifier.

9.1.1 Electrostatic shielding

Shield material and grounding determine the shield effectiveness. Making the shield with a material of high electrical conductivity ensures that the coupled currents produce little voltage drop across the shield. Then, little of the original field continues within the shield boundaries. To be effective, the shield's ground must be earth ground because only this remains a common reference for the separate objects involved in the coupling. Further, the shield should be connected to the system ground to minimize the effects of the parasitic capacitances introduced. All shield enclosures form parasitic capacitances with each component shielded. When left ungrounded, the shield develops a signal voltage determined primarily by the circuit's larger signals through parasitic capacitance coupling. Then, the shield signal couples back into the circuit through other parasitic capacitances. Returning the shield connection to the system common removes the shield signal and the associated coupling. Then, shield-conducted capacitive currents from the output of a current-to-voltage converter shunt to ground rather than to the circuit's input. This avoids a bandwidth restriction otherwise imposed by the associated parasitic bypass of the circuit's feedback resistor. However, the shield then produces a capacitance from the circuit's input to ground, adding to the noise gain peaking described in Sec. 5.1.2.

9.1.2 The differential-input current-to-voltage converter

Most electrostatic noise coupling results from the ac power line and couples to all points of a circuit nearly equally. The equal coupling makes this interference a natural candidate for removal by CMR, and op amp CMR readily removes signals at the power-line frequency. However, the basic current-to-voltage converter fails to utilize the amplifier's CMR capability. There, the single-ended input connection of the circuit disables this feature. Switching to a differential-input connection activates CMR for reduced noise sensitivity and reduced dc error as well. However, as will be described, this connection does reduce bandwidth. Also, op amp CMR does not totally replace shielding because the electrostatic coupling to the two amplifier inputs will not be exactly the same. Often, CMR serves as a backup to shielding by removing the residual coupling that passes through shield imperfections.

Both the op amp and the photodiode possess differential qualities that combine to reject noise coupling effects. The opposite polarity inputs of the op amp separate any differentially applied signal from common-mode signals. The photodiode produces a differential signal in the form of opposite polarity currents, as seen from the two-diode terminals. Current flowing into one photodiode terminal flows out of the other, producing a differential current signal. Applying this signal to the amplifier's two inputs produces the differential-input current-to-voltage converter of Fig. 9.1. There, a second current-to-voltage

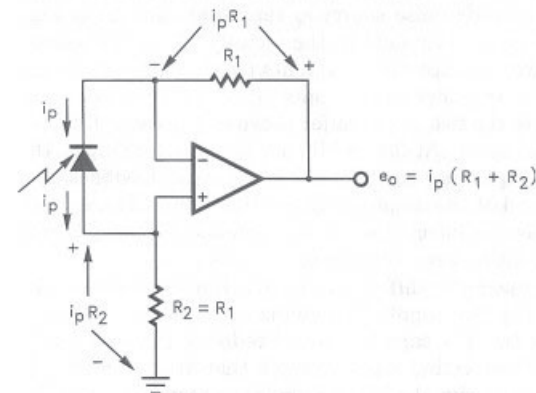


Figure 9.1 Adding a second current-to-voltage conversion resistor R_2 and bootstrapping the photodiode produces a differential-input circuit for rejection of external noise coupling.

conversion resistance R_2 and a bootstrap connection for the photodiode differentiate this circuit from the basic case. These simple differences produce easily anticipated changes in circuit gain and dc offset but less obvious changes in noise performance.

First, the circuit changes increase the current-to-voltage gain and automatically provide for dc offset reduction. In the figure, the added resistor R_2 develops a second signal voltage from the anode current of the diode. This signal $i_p R_2$ adds to the normal signal $i_p R_1$ increasing the output signal to $e_o = i_p (R_1 + R_2)$. Resistor R_2 also occupies the position of the offset compensation resistance described in Sec. 2.2.1. As described there, making the circuit's two resistances equal produces canceling offset effects with the amplifier's two input currents. As will be seen, making $R_2 = R_1$ also balances the circuit for optimum rejection of noise interference. The differential connection here reduces offset error even further through two other features. First, connecting the diode's anode to the top of R_2 bootstraps the diode on that resistor's voltage. Then, the dc compensation voltage developed across R_2 no longer drops across the diode, avoiding the associated diode leakage error. The second feature permits a reduction of resistance values. With equal resistances, this differential connection doubles the current-to-voltage gain through $e_o/i_p = 2R_1$. Then, cutting the resistance values in half maintains the same gain as before and reduces the circuit's offset sensitivity to amplifier input currents.

This differential-input connection also provides the desired CMR of coupled noise as shown in Fig. 9.2. There, setting the photodiode current to zero permits a focus upon the electrostatic coupling effects alone. Electrostatic noise source e_e represents any ac voltage source that creates an electric field in the vicinity of the photodiode amplifier. This source couples noise currents i_{ne} through mutual capacitances C_M to the amplifier's two inputs. It might first seem that the coupling effects to the two would differ because of greatly different impedances at those inputs. At the amplifier's noninverting input, the circuit presents an impedance equal to R_2 . In contrast, feedback produces a virtual ground at the amplifier's inverting input. There, feedback absorbs any current injected without producing a corresponding voltage change, indicating zero impedance.

In spite of this impedance difference, feedback also equalizes the coupling effects at the two inputs by equalizing the signal voltages impressed upon the two C_M capacitances. Feedback forces the voltage at the amplifier's inverting input to track that at the noninverting input, removing any signal voltage imbalance across the two C_M components. With equal signal voltages, the two capacitances couple equal i_{ne} noise currents to the circuit's two inputs. There, the two i_{ne}

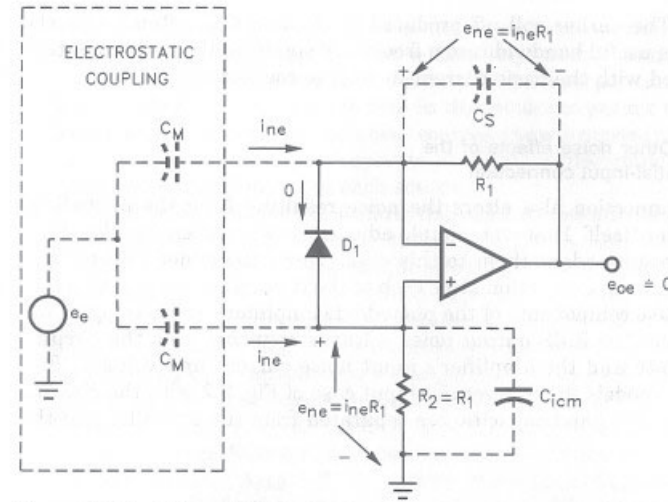


Figure 9.2 The differential input circuit of Fig. 9.1 accepts equal noise currents i_{ne} from an electrostatic source to produce two equal e_{ne} noise effects that the op amp's CMR rejects.

currents develop equal and counteracting $e_{ne} = i_{ne} R_1$ noise voltages on the circuit's two resistors. Making $R_2 = R_1$ as shown ideally produces a zeroed output response, $e_{noe} \approx 0$, from the electrostatic coupling.

The accuracy of the above noise cancellation depends upon the matching of mutual capacitances, resistors, and parasitic capacitances. For a given noise source, locating the two circuit inputs equidistant from the source largely matches the associated mutual capacitances. Then, matching the values for the two circuit resistors assures accurate cancellation of the i_{ne} signal effects, until higher frequencies are reached. There, differences in parasitic capacitive shunting imbalance the net impedances that react with the i_{ne} currents. Typically, only about 0.5 pF of stray capacitance shunts R_1 , but the significantly larger, common-mode input capacitance of the op amp shunts R_2 . This difference removes the benefit of CMR at higher frequencies, but the rejection typically remains high at the troublesome power-line frequency.

This capacitance imbalance also imposes different bandwidth limits for the circuit's two signal paths. A signal current driving R_2 encounters the shunting of the larger C_{icm} capacitance before C_S significantly shunts the current driving R_1 . This difference produces a frequency response characterized by two gain plateaus that are separated by

6 dB. The earlier roll off produced by R_2 and C_{icm} often limits the circuit's useful bandwidth to a frequency significantly lower than that achieved with the basic current-to-voltage converter.

9.1.3 Other noise effects of the differential-input connection

This connection also alters the noise resulting from the photodiode amplifier itself. However, simple adjustments to the earlier Sec. 5.3.1 noise results adapt them to this case. Here, the added resistor and the bootstrap connection alter each of the fundamental E_{noR} , E_{noi} , and E_{noe} noise components of the photodiode amplifier. These components represent the RMS output noise voltages resulting from the circuit's resistance and the amplifier's input noise current and voltage. Figure 9.3 models the differential-input case of Fig. 9.2 with the circuit's noise sources and capacitances separated from the amplifier and the

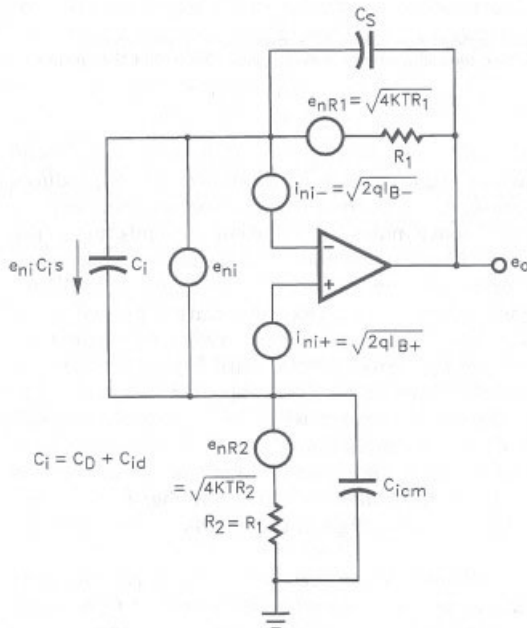


Figure 9.3 The differential-input connection of Fig. 9.2 adds the noise effects of R_2 directly through e_{nR2} and indirectly through the interaction of i_{ni+} and e_{ni} with this resistance.

resistors. There, C_i represents the net capacitance of the input circuit, $C_i = C_D + C_{id}$, where C_D is the diode capacitance, and C_{id} is the differential-input capacitance of the amplifier. Superposition analysis sets the photodiode current to zero in this model to permit a focus on the noise effects alone. The noise sources shown represent spectral noise densities and require separate analyses to determine the RMS output noise resulting from each source.

The addition of R_2 for this differential-input circuit affects noise both directly and indirectly. Directly, the presence of R_2 introduces an additional noise source to E_{noR} and, indirectly, this resistance increases E_{noi} and E_{noe} . First, the noise voltage of R_2 increases the output noise component E_{noR} produced by the circuit resistances. This added noise source, $e_{nR2} = \sqrt{4KTR_2}$, drives the amplifier's noninverting input and the shunting C_{icm} capacitance. Superposition analysis defines the resulting output noise effects by first considering all other signal and noise sources set to zero. Then, at lower frequencies, e_{nR2} effectively drives a voltage follower, transferring directly to the circuit output with a noise gain $A_{nR2} = 1$. At higher frequencies, C_{icm} absorbs the noise energy of R_2 , shorting out the e_{nR2} source. The intervening pole at $f_{p2} = 1/2\pi R_2 C_{icm}$ rolls off the corresponding noise gain as expressed by $A_{nR2} = 1/(1 + jf/f_{p2})$. As described in Sec. 5.1.3, applying the RMS noise integral to the product of the e_{nR2} noise density and the A_{nR2} noise gain defines the resulting RMS output noise voltage. Then,

$$\begin{aligned} E_{noR2}^2 &= \int_0^\infty |A_{nR2} e_{nR2}|^2 df \\ &= \int_0^\infty \frac{4KTR_2}{1 + f^2/f_{p2}^2} df \end{aligned}$$

and $E_{noR2} = \sqrt{2KTR_2\pi f_{p2}}$, where $f_{p2} = 1/2\pi R_2 C_{icm}$.

The noise density of R_1 , e_{nR1} , produces a second RMS resistor noise component. As with the basic photodiode amplifier, the $e_{nR1} = \sqrt{4KTR_1}$ noise also transfers to the circuit output with unity gain until rolled off by a pole at $f_{pf} = 1/2\pi R_1 C_s$. The resulting noise gain, $A_{nR1} = 1/(1 + jf/f_{pf})$, and the noise density e_{nR1} produce an RMS output noise analogous to that above, $E_{noR1} = \sqrt{2KTR_1\pi f_{pf}}$. An RMS summation of E_{noR1} with E_{noR2} defines the circuit's net output noise E_{noR} , produced by the circuit's resistors. For the typical $R_2 = R_1$ case this combination produces

$$E_{noR} = \sqrt{E_{noR1}^2 + E_{noR2}^2} = \sqrt{2KT\pi R_1(f_{pf} + f_{p2})}$$

where $f_{pf} = 1/2\pi R_1 C_s$, and $f_{p2} = 1/2\pi R_2 C_{icm}$.

Compared with the basic current-to-voltage converter, where $E_{\text{noR}} = \sqrt{2KT\pi R_f f_{\text{pf}}}$, the above noise component typically decreases. Comparing the $R_1(f_{\text{pf}} + f_{p2})$ and $R_f f_{\text{pf}}$ products of the two E_{noR} equations reveals this decrease. For equal gains in the two cases $R_1 = R_f/2$, making $R_1(f_{\text{pf}} + f_{p2}) = (R_f/2)(f_{\text{pf}} + f_{p2})$. Then, only the condition of $f_{p2} = f_{\text{pf}}$ would make $R_1(f_{\text{pf}} + f_{p2}) = R_f f_{\text{pf}}$ to equalize the two E_{noR} results. However, in almost all cases, $f_{p2} < f_{\text{pf}}$ by a significant amount, making $R_1(f_{\text{pf}} + f_{p2}) < R_f f_{\text{pf}}$ and producing a lower E_{noR} for the differential input case. Note, however, that this noise reduction actually results from the decreased bandwidth f_{p2} of the differential-input circuit.

The addition of R_2 indirectly introduces another noise component through this resistor's reaction with the amplifier's input noise current. With the basic photodiode amplifier, the noise current of the amplifier's inverting input, $i_{\text{ni-}}$ of Fig. 9.3 produces an output noise voltage by flowing through the feedback resistance. That also occurs here and produces a noise voltage with R_1 of $e_{\text{nRi-}} = (i_{\text{ni-}})R_1$, where $i_{\text{ni-}} = \sqrt{2qI_{B-}}$ and I_{B-} is the input bias current of the amplifier's inverting input. Being across R_1 , this noise voltage appears in series with e_{nR1} , and the amplifier supplies the same noise gain to the two noise voltages. From before, $A_{\text{nR1}} = 1/(1 + jf/f_{\text{pf}})$, and applying the RMS integral to the product of $e_{\text{nRi-}}$ and A_{nR1} yields the RMS output noise voltage produced by $i_{\text{ni-}}$, $E_{\text{noi-}} = R_1 \sqrt{q\pi f_{\text{pf}} I_{B-}}$.

With the differential case here, the noise current of the noninverting input $i_{\text{ni+}}$ also produces an output noise voltage. An analogous analysis defines its effect. Here, $i_{\text{ni+}} = \sqrt{2qI_{B+}}$, where I_{B+} is the input bias current of the amplifier's noninverting input. Noise current $i_{\text{ni+}}$ flows through the R_2 resistance, producing the noise voltage $e_{\text{nRi+}} = (i_{\text{ni+}})R_2$ across R_2 . There, $e_{\text{nRi+}}$ appears in series with e_{nR2} , and the previous $A_{\text{nR2}} = 1/(1 + jf/f_{p2})$ noise gain applies to $e_{\text{nRi+}}$ as well. This $e_{\text{nRi+}}$, A_{nR2} combination produces an RMS output noise component of $E_{\text{noi+}} = R_2 \sqrt{q\pi f_{p2} I_{B+}}$. An RMS combination of $E_{\text{noi+}}$ and $E_{\text{noi-}}$ produces the net RMS output noise due to amplifier input currents. Assuming the typical conditions of $R_2 = R_1$ and $I_{B+} = I_{B-}$,

$$E_{\text{noi}} = R_1 \sqrt{q\pi I_{B-} (f_{\text{pf}} + f_{p2})}$$

where $f_{\text{pf}} = 1/2\pi R_1 C_S$, and $f_{p2} = 1/2\pi R_2 C_{\text{icm}}$. Compared with the basic current-to-voltage converter, where $E_{\text{noi}} = R_f \sqrt{q\pi f_{\text{pf}} I_{B-}}$, the above noise component increases due to the added f_{p2} term but decreases for a gain-equivalent choice of $R_1 = R_f/2$. Typically, the combined increase and decrease effects produces a net decrease in noise but, again, only due to the reduced bandwidth imposed by f_{p2} .

The third and final noise effect of the R_2 addition increases the E_{noe} effect of the amplifier input noise voltage e_{ni} . Evaluation of the

net e_{ni} effect requires RMS integration of the product $A_{\text{ne}} e_{\text{ni}}$, where A_{ne} is the noise-gain response presented by the circuit to e_{ni} . However, the multiple capacitances of the differential-input photodiode amplifier produce a complex A_{ne} response, making direct mathematical analysis cumbersome. For noise analysis purposes, an intuitive examination of the circuit produces a simpler and adequate approximation of the A_{ne} response. For this examination consider all signal sources except for e_{ni} set to zero in Fig. 9.3. Then, only e_{ni} produces an output signal, and at lower frequencies, the circuit capacitances have no effect. For this combination, a loop equation shows that $e_o = e_{\text{ni}}$ in response to a low-frequency noise gain of $A_{\text{neL}} = 1$. At higher frequencies, the presence of e_{ni} across C_i develops a noise current $e_{\text{ni}} C_i s$ that flows through both R_1 and R_2 . The resulting noise voltages on these resistors produce an A_{ne} gain rise, marked by a feedback zero at approximately $f_{zf} = 1/2\pi (R_1 + R_2) C_i$. For the typical $R_2 = R_1$ case, this zero becomes $f_{zf} = 1/4\pi R_1 C_i$. At even higher frequencies, C_{icm} and C_S bypass R_2 and R_1 , limiting the effect of the noise current developed in C_i . There, the circuit's capacitances control the response to e_{ni} , and a loop equation defines a high-frequency noise gain for e_{ni} approximated by

$$A_{\text{neH}} = 1 + C_i \frac{C_{\text{icm}} + C_S}{C_{\text{icm}} C_S}$$

Figure 9.4 displays the resulting $e_{\text{noe}} = A_{\text{ne}} e_{\text{ni}}$ output noise-density response. This result closely resembles the original e_{noe} response of Sec. 5.2.3, making the earlier results applicable here with just a few modifications. As before, five regions separate the $A_{\text{ne}} e_{\text{ni}}$ response of the figure to simplify an otherwise complex integration. Here, changes in f_{zf} , f_i and the shape of region 3 response differentiate this case from the earlier result. In regions 1 and 2 of Fig. 9.4, the low-frequency noise gain $A_{\text{neL}} = 1$ makes the e_{noe} response again follow the e_{ni} noise response of the amplifier. Then in region 3, the e_{noe} response rises due to the A_{ne} response zero at $f_{zf} = 1/4\pi R_1 C_i$, as identified above. The resulting response rise continues with a constant slope in region 3 until temporarily interrupted by the C_{icm} bypass of R_2 . This bypass cuts the noise gain in half and introduces a dip in the curve. However, this dip occurs late in the region 3 span, and ignoring it produces little error. Thus, assuming a straight line response between f_{zf} and f_{pf} adequately approximates the region 3 response and extends previous results to this case.

The C_{icm} capacitance also affects regions 4 and 5 of the curve where the intercept frequency f_i defines a boundary for the two. This frequency marks the crossing of the circuit's $1/\beta$ curve and the A_{OL} response. There, $f_i = \beta_H f_c$, where β_H is the high-frequency value of

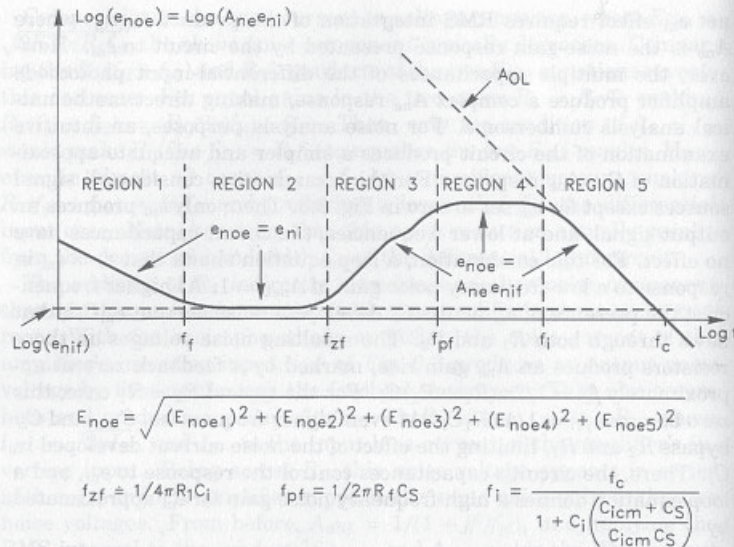


Figure 9.4 The differential-input model of Fig. 9.3 modifies the frequency response of the photodiode amplifier's output noise density with a negligible dip in region 3 and a change in the f_i boundary between regions 4 and 5.

feedback factor and f_c is the amplifier's unity-gain crossover frequency. The factor β_H follows from the Fig. 9.3 derivation of the high-frequency noise gain A_{neH} through the relationship $\beta_H = 1/A_{neH}$. Then, substituting the previous A_{neH} result in $f_i = \beta_H f_c = f_c/A_{neH}$ produces

$$f_i = \frac{f_c}{1 + C_i(C_{icm} + C_S)/C_{icm}C_S}$$

Making the preceding changes to E_{noR} , E_{noi} , f_{zf} , and f_i adapts the original noise analysis of Sec. 5.3.1 to this differential-input case. This analysis covers only the noise produced by the photodiode amplifier components and not the noise introduced by external sources. Then, the analysis of self-induced noise for the Fig. 9.2 follows the steps listed below:

$E_{noe1} = e_{nif}\sqrt{f_f \ln(f_f/f_1)}$, where e_{nif} is the e_{ni} noise floor level and f_f is the $1/f$ noise corner frequency. As explained in Sec. 5.2.3, setting $f_1 \approx 0.01$ Hz avoids a theoretical contradiction and ensures analytical accuracy.

$E_{noe2} = e_{nif}\sqrt{f_{zf} - f_f}$, where $f_{zf} = 1/4\pi R_1 C_i$, and $C_i = C_D + C_{id}$. Here and in the E_{noe4} analysis, numerical evaluation can produce the square root of a negative number. Then, substitute zero for the result as explained in Sec. 5.3.1.

$E_{noe3} = (e_{nif}/f_{zf})\sqrt{(f_{pf}^3 - f_{zf}^3)/3}$, where $f_{pf} = 1/2\pi R_f C_S$, and C_S is the stray feedback capacitance.

$E_{noe4} = (1 + C_i/C_S)e_{nif}\sqrt{f_i - f_{pf}}$, where $f_i = f_c/[1 + C_i(C_{icm} + C_S)/C_{icm}C_S]$.

$E_{noe5} = (e_{nif}f_c)\sqrt{1/f_i}$, where f_c is the op amp's unity-gain crossover frequency.

$E_{noe} = \sqrt{E_{noe1}^2 + E_{noe2}^2 + E_{noe3}^2 + E_{noe4}^2 + E_{noe5}^2}$, the total RMS output noise resulting from the spectral noise voltage e_{ni} of the amplifier input.

$E_{noR} = \sqrt{2KT\pi R_1(f_{pf} + f_{p2})}$, where $f_{pf} = 1/2\pi R_1 C_S$, and $f_{p2} = 1/2\pi R_2 C_{icm}$.

$E_{noi} = R_1\sqrt{q\pi I_B - (f_{pf} + f_{p2})}$.

$E_{no} = \sqrt{E_{noR}^2 + E_{noi}^2 + E_{noe}^2}$.

9.1.4 An alternate differential-input photodiode amplifier

The preceding photodiode amplifier reduces sensitivity to electrostatic coupling but sacrifices signal bandwidth. There, the C_{icm} bypass of R_2 imposes a new bandwidth limit at a frequency lower than the normal limit imposed by the C_S bypass of R_1 . An alternative circuit restores bandwidth by removing signal swing from C_{icm} . Shown in Fig. 9.5, this differential-input circuit connects the photodiode between the inputs of two current-to-voltage converters and subtracts the two converter output signals with a difference amplifier. The resulting circuit becomes the common three-op-amp instrumentation amplifier with the photodiode replacing the normal gain-setting resistor. With this connection, grounded noninverting inputs for A_1 and A_2 assure zero signal swing across the op amps' C_{icm} capacitances to avoid the preceding bandwidth limit. Then, the bandwidth limit returns to the stray capacitance bypass of the converter feedback resistors R_1 and R_2 . Making $R_2 = R_1$ sets this limit at $1/2\pi R_1 C_S$ for both converters and balances the circuit for optimum rejection of electrostatic coupling. Compared with the basic photodiode amplifier, bandwidth actually improves here due to the dual gain setting resistors. For a given gain

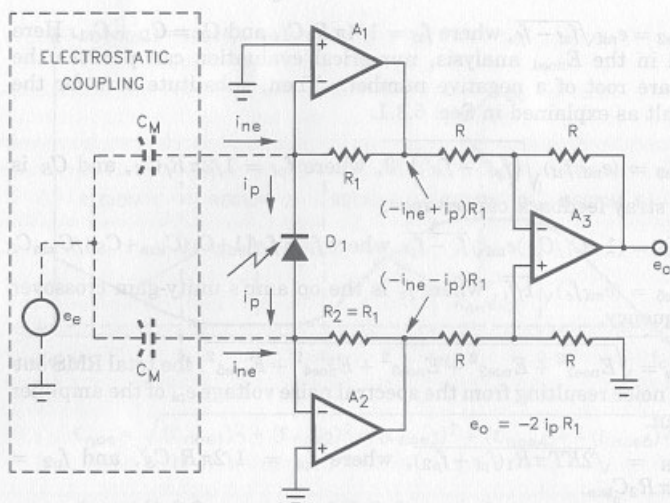


Figure 9.5 Connecting the photodiode to a grounded-input instrumentation amplifier removes the common-mode swing of Fig. 9.1, avoiding the bandwidth limit imposed there by common-mode input capacitance.

level, dual resistors permit a 2:1 resistance reduction, reducing bandwidth sensitivity to stray capacitance by the same factor.

As with the preceding circuit, the differential inputs of this alternative circuit separate the electrostatic interference for rejection by CMR while amplifying the desired signal. Mutual capacitances C_M again couple equal electrostatic interference currents i_{ne} to R_1 and R_2 . For $R_2 = R_1$, the equal currents produce equal and in-phase voltages $-i_{ne}R_1$ at the outputs of A_1 and A_2 . This makes the $-i_{ne}R_1$ signals a common-mode signal at the input of the A_3 difference amplifier. From there, the subtracting action of this amplifier rejects these identical signals. In contrast, the differential nature of the photodiode current i_p produces opposite-phase voltages at the outputs of A_1 and A_2 . This current flows into one diode terminal and out of the other, producing i_pR_1 at the output of A_1 and $-i_pR_1$ at the output of A_2 . Then, the subtracting action of the A_3 difference amplifier delivers an output voltage of $e_o = -2i_pR_1$. This separation of differential and common-mode signals depends upon the CMR of A_3 , and this rejection decreases with increasing frequency. However, the rejection generally remains high at the power-line frequency most often involved in electrostatic noise coupling.

9.1.5 Other noise effects of the differential-input alternative

While it does reduce electrostatic noise, the Fig. 9.5 alternative adds the noise effects of several circuit components. Instead of one amplifier and one resistor, this alternative includes three amplifiers and six resistors. Fortunately, A_3 and the four resistors of its difference amplifier enter the signal path after the signal gain provided by A_1 and A_2 . This placement makes the noise effects negligible for the five components of the A_3 difference amplifier. Then, the noise introduced by A_1 , A_2 , R_1 , and R_2 represents the noise contribution of the photodiode amplifier. These circuit components form two matching current-to-voltage converters that each introduce the noise components previously summarized in Sec. 5.3.1. Simple RMS addition of the two converter's noise effects produces the net result for this case. For matched converter components, this addition of two equal noise components, $\sqrt{E_n^2 + E_n^2}$, simply produces $E_n\sqrt{2}$, introducing a $\sqrt{2}$ multiplier to the circuit's noise components. Adding this multiplier to the results of the earlier Sec. 5.3.1 results adapts those results to this case. This addition modifies E_{noR} , E_{noi} , and E_{noe1} through E_{noe5} , which represent the noise effects of the circuit's resistance noise, current noise, and the five subcomponents of voltage noise. Making these changes to the Sec. 5.3.1 summary produces the equivalent noise analysis summary for the differential-input alternative of Fig. 9.5.

$E_{noe1} = e_{nif}\sqrt{2f_f \ln(f_f/f_1)}$, where e_{nif} is the e_{ni} noise floor level, and f_f is the $1/f$ noise corner frequency. As explained in Sec. 5.2.3, setting $f_1 \approx 0.01$ Hz avoids a theoretical contradiction and ensures analytical accuracy.

$E_{noe2} = e_{nif}\sqrt{2(f_{zf} - f_f)}$, where $f_{zf} = 1/2\pi R_1(C_i + C_S)$, and $C_i = C_D + C_{id} + C_{icm}$. For applications requiring feedback compensation C_f , replace C_S with $C_f + C_S$ here and in the steps that follow. Here and in the E_{noe4} analysis, numerical evaluation can produce the square root of a negative number. Then, substitute zero for the result as explained in Sec. 5.3.1.

$E_{noe3} = (e_{nif}/f_{zf})\sqrt{2(f_{pf}^3 - f_{zf}^3)/3}$, where $f_{pf} = 1/2\pi R_1 C_S$, and C_S is the stray feedback capacitance.

$E_{noe4} = (1 + C_i/C_S)e_{nif}\sqrt{2(f_i - f_{pf})}$, where $f_i = f_c C_S/(C_i + C_S)$.

$E_{noe5} = (e_{nif}f_c)\sqrt{2/f_i}$, where f_c is the op amp's unity-gain crossover frequency.

$E_{\text{noe}} = \sqrt{E_{\text{noe}1}^2 + E_{\text{noe}2}^2 + E_{\text{noe}3}^2 + E_{\text{noe}4}^2 + E_{\text{noe}5}^2}$, the total RMS output noise resulting from the spectral noise voltage e_{ni} of the amplifier inputs.

$E_{\text{noR}} = 2\sqrt{KTR_1\pi BW_t}$, where $BW_t = 1/2\pi R_1 C_S$. For applications that add a feedback capacitance C_f , replace C_S with $C_f + C_S$.

$$E_{\text{noi}} = R_1 \sqrt{2q\pi BW_t I_{B-}}$$

$$E_{\text{no}} = \sqrt{E_{\text{noR}}^2 + E_{\text{noi}}^2 + E_{\text{noe}}^2}$$

9.2 Reducing Magnetic and RFI Coupling

Magnetic noise coupling and radio-frequency interference (RFI) introduce circuit noise through a common coupling mechanism, mutual inductance. There, the interference source acts like the primary of a transformer, and circuit loops resemble secondary windings. While often considered separately, RFI simply represents the higher-frequency form of parasitic magnetic coupling. However, in shielding, this frequency distinction suggests the separation. The frequency of the magnetic source greatly influences the effectiveness of magnetic shielding materials. At lower frequencies, only the magnetic property of ferrous metals permits practical shield thickness. However, at higher frequencies, decreases in both the magnetic response and in the shield thickness requirement make copper a good alternative material. There, even the copper layer of a ground plane becomes effective. In addition to shielding, careful choosing of the amplifier's physical and electrical configurations reduces magnetic coupling effects. Minimizing the physical areas of circuit loops reduces the coupling efficiency, and using a differential-input structure makes CMR reduce the residual effect.

9.2.1 Magnetic shielding

A comparison of magnetic coupling with electrostatic coupling and a transformer analogy illustrate the added shielding requirement of the magnetic case. The electrostatic shielding described before simply requires a shield material of high electrical conductivity. There, the high conductivity shorts the currents transmitted through mutual capacitances to ground. As long as these currents develop little voltage across the shield, no significant field continues within the shield boundaries. However, magnetic fields couple through mutual inductances, rather than capacitances, and produce voltage rather than current signals. The coupled signals develop in all circuit loops within the field and have relative magnitudes determined by loop areas. For the magnetic field, a high-conductivity electrostatic shield only forces

an equipotential condition at the shield boundaries. Grounding this shield does not terminate the magnetic field and may just simply establish a zero voltage reference. By analogy, grounding the center tap of a transformer's secondary winding establishes a reference voltage but does not terminate the transformer's coupling. Thus, an electrostatic shield distorts the magnetic field but does not necessarily remove the field energy. Some portion of the field's energy continues the field within the shield boundaries.

Fortunately for noise reduction, some of the magnetic field energy dissipates in eddy currents and ohmic absorption within the shield. Shield material selection optimizes this dissipation with different results for low and high frequencies. To effectively reduce magnetic coupling, the shield must absorb the field energy as the field travels through the shield's walls. A combination of high conductivity and high magnetic permeability produces the greatest field absorption. High electrical conductivity ensures that the induced shield currents produce little voltage drop across the shield, preventing continuation of the field through electrostatic coupling within. High magnetic permeability ensures efficient absorption of the magnetic field moving through the shield. At the power-line frequency, common to the most serious magnetic coupling effects, the magnetic permeability of ferrous materials reduces magnetic coupling an order of magnitude better than other shield materials.² For such materials, the realignment of magnetic dipoles in the ferrous material consumes magnetic field energy in the form of eddy currents. This improvement occurs even though the electrical conductivity of steel is about an order of magnitude lower than that of copper. Also, at power-line frequencies, the shield thickness required for a copper shield becomes prohibitive due to the great skin depth of this material.

At radio frequencies (RF), copper becomes a more reasonable, but never superior, alternative shield material. There, both the magnetic permeability of ferrous metals and the skin depths of all metals drop dramatically. Above 10 kHz, the magnetic permeability of ferrous metals drops due to the finite time required to realign the material's magnetic dipoles. The shorter periods of high-frequency signals preclude the realignment and the conversion of field energy to eddy currents. Then, a shield must remove magnetic field energy through ohmic absorption as reflected by skin depth. Skin depth indicates the thickness of a given shield material required to attenuate a magnetic field by a factor of $e = 2.73$. Fortunately, skin depth also decreases at higher frequencies, due to the shorter wavelength of the signal. Thus, at high frequencies, the decreased permeability of ferrous materials reduces their relative advantage over copper, and the reduced skin depths

make the required copper thickness more practical. However, for a given shield thickness, steel still retains about a factor of 3 absorption advantage.

9.2.2 Circuit reductions of magnetic coupling

As described above, physically separating and shielding the photodiode amplifier from a magnetic noise source offers the best protection against magnetic noise coupling. However, the amplifier's physical and electrical configurations also offer powerful alternatives. Configuring the amplifier layout for minimum physical loop areas and designing for CMR also reduce magnetic noise coupling. Minimizing the areas of the circuit's loops minimizes the mutual inductances that couple the magnetic noise signal. Careful component layout achieves this by placing the photodiode amplifier's components close to the op amp. Simply minimizing the lengths of component interconnections generally produces the desired area reductions. Finally, for lower-frequency noise, an op amp's CMR reduces coupling effects when using a differential-input configuration for the photodiode amplifier. Then, by matching loop areas and distances from a noise source the amplifier's CMR rejects the resulting, equal noise coupling effects.

For photodiode amplifiers, the most confusing task in magnetic coupling reduction can be the identification of the receptor loops. The physical arrangement of the circuit's components form these loops in several ways, and Fig. 9.6 shows the loops produced by the differential-input photodiode amplifier of Fig. 9.1. Here, three loops illustrate those commonly formed with photodiode amplifiers through the circuit connections of the photodiode, the feedback, and the load. The op amp would seem to break these loops, but as will be seen, the amplifier's feedback action continues them.

Differing amplifier responses and a grounding connection differentiate the three loops shown. First, the photodiode's connection to the op amp inputs forms loop L_3 that develops magnetic coupling signal e_{m3} . In this loop, it would seem that the amplifier's very high input impedance would break the loop by interrupting an otherwise continuous conductive path. However, feedback forces the voltage between the amplifier inputs to zero, just as if a short-circuit connected them. This completes the L_3 loop in spite of the high input impedance. The equivalent short circuit of this case parallels the condition of a shorted transformer secondary. There, the transformer continues its magnetic coupling and simply delivers its short-circuit output current. Here, the L_3 loop acts as the receptor or secondary side of a parasitic transformer and similarly delivers its short-circuit current i_{ss} . This current can-

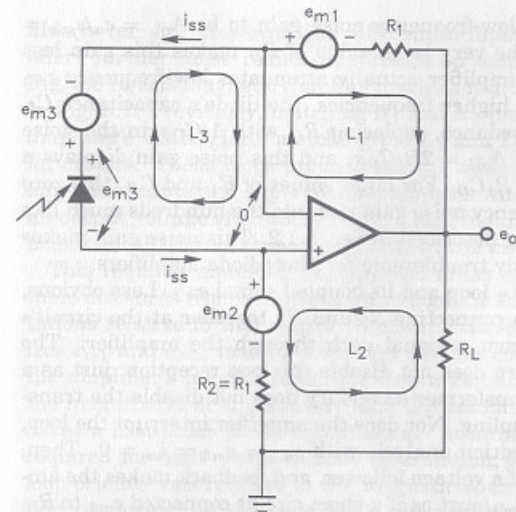


Figure 9.6 Photodiode amplifiers form magnetic receptor loops, in spite of the high-impedance interruption of the op amp inputs, due to the amplifier's output response to signals at its inputs.

not flow into the amplifier inputs because of their high impedances. Instead, feedback supplies this current through R_1 and forces the return current into R_2 . To do so, the amplifier develops an output voltage supporting the resulting voltage drops on the two resistors. This condition maintains the necessary zero voltage between the amplifier inputs and retains a short-circuit appearance to the L_3 loop.

The above feedback action and the photodiode capacitance make the L_3 loop a serious noise source for the photodiode amplifier. The output voltage developed by feedback above is actually an amplified replica of the e_{m3} noise. To determine the corresponding noise gain, let superposition set e_{m1} and e_{m2} to zero and consider i_{ss} to be only that current produced by e_{m3} . This current develops a voltage across the photodiode that counteracts e_{m3} in the L_3 loop and retains zero voltage between the op amp inputs. Generally small, e_{m3} does not produce a forward bias condition for the photodiode but keeps the diode junction near its off state. At lower frequencies, the off impedance of this junction equals the photodiode's very large dark resistance, R_D . Together, e_{m3} and R_D produce a very small current of $i_{ss} = e_{m3}/R_D$. The circuit conducts this current from R_1 and to R_2 , producing an output voltage of $e_o = i_{ss}(R_1 + R_2) = 2i_{ss}R_1$. Combining the last two

equations shows the low-frequency noise gain to be $A_{nL} = e_o/e_{m3} = 2R_1/R_D$. Typically, the very large value of R_D makes this gain less than unity, and the amplifier actually attenuates low-frequency e_{m3} signals. However, at higher frequencies, the diode's capacitance C_D shunts the diode impedance, replacing R_D with $1/C_D s$ in the noise gain equation. Then, $A_{nH} = 2R_1 C_D s$, and this noise gain displays a response zero at $1/4\pi R_1 C_D$. For large values of R_1 and C_D , this zero makes the high-frequency noise gain rise into the hundreds much like the noise gain peaking described in Sec. 5.1.2. This noise gain makes the L_3 loop particularly troublesome for photodiode amplifiers.

Next, consider the L_2 loop and its coupled signal e_{m2} . Less obvious, this loop results from connecting R_2 and R_L together at the circuit's ground return and from a signal path through the amplifier. The ground connection here does not disable the loop reception, just as a grounded tap on a transformer secondary does not disable the transformer's magnetic coupling. Nor does the amplifier interrupt the loop, as seen by a superposition analysis with $e_{m1} = e_{m3} = i_{ss} = 0$. Then, e_{m2} drives the input of a voltage follower, and feedback makes the amplifier output follow e_{m2} , just as if a short circuit connected e_{m2} to R_L . This equivalent short circuit completes the L_2 loop through the amplifier. Due to the follower response of this case, the circuit presents a noise gain of $A_n = 1$ to e_{m2} .

The L_2 loop demonstrates another benefit of the differential-input photodiode amplifier of the figure. There, the circuit's bootstrap connection isolates the diode capacitance from the e_{m2} signal, reducing noise gain. Returning the anode of the diode to the amplifier's noninverting input bootstraps the diode on the e_{m2} signal at that input. In contrast, common photodiode amplifiers connect the anode to ground, impressing e_{m2} across the diode. Then, e_{m2} reacts with the diode capacitance, increasing high-frequency noise gain, just as e_{m3} did before.

The final circuit loop L_1 depends upon the equivalent short-circuits described for L_2 and L_3 . As described, feedback action produces these equivalents from the amplifier's noninverting input to both the inverting input and the output. Visual examination of Fig. 9.3 shows that making the two equivalent connections completes the L_1 loop. Superposition analysis again determines the corresponding noise gain but using $e_{m2} = e_{m3} = i_{ss} = 0$ in this case. Under these conditions, one side of the e_{m1} source connects to the virtual ground that feedback produces at the amplifier's inverting input. The other side connects to the output through R_1 and, with $i_{ss} = 0$, no current flows in this resistor. Thus, $e_o = -e_{m1}$ for a noise gain of $A_n = -1$.

Minimizing the preceding loops reduces but does not eliminate the noise from magnetic coupling. Some finite loop areas and coupling

always remain. However, the differential-input photodiode amplifier offers further noise reduction through op amp CMR. The circuit of Fig. 9.6 rejects magnetic coupling much as it did electrostatic coupling in Fig. 9.1. Previously, matching R_1 and R_2 resistances activated the amplifier's CMR. Here, matching the L_1 and L_2 loops produces similar results. These loops produce the e_{m1} and e_{m2} signals at opposite amplifier inputs and superposition analysis with $e_{m3} = i_{ss} = 0$ defines an output voltage of $e_o = e_{m2} - e_{m1}$. Thus, making $e_{m2} = e_{m1}$ produces noise cancellation of these two effects at the circuit output.

This requires matching the physical areas of the L_1 and L_2 loops, their distances from an interfering magnetic source, and their orientations relative to that source. Matching these three features equalizes e_{m1} and e_{m2} , making their net effect a common-mode signal at the amplifier's inputs. Matching loop areas and distances equalizes the magnitudes of e_{m2} and e_{m1} , and the matching distances also produces a first-order phase equalization. Accurate phase matching, as required for high common-mode cancellation, also requires matched loop orientations relative to the magnetic source. Then, the subtraction performed by the amplifier's differential inputs produces optimum rejection of magnetic noise coupling. Most often, this noise reduction opportunity aids in the rejection of low-frequency local noise sources such as power transformer interference. At higher frequencies, the amplifier's CMR declines, reducing its benefit to noise reduction.

There, RFI filtering somewhat replaces CMR to counteract the accompanying increase in magnetic coupling efficiency. This increase potentially dominates noise performance, often due to the RFI of coresident digital circuitry. However, the bandwidth limitation of the amplifier permits a filtering solution. The limited bandwidth of the op amp typically restricts the photodiode amplifier's response to frequencies well below the RF range. This permits filter removal of RFI signals at the amplifier output without restricting the circuit's useful bandwidth.

Still, output filtering does not necessarily remove the effects of higher-level RFI. The input circuit of an op amp can act like an RF detector, separating a lower frequency envelope from a carrier.⁴ There, larger RF signals drive the emitter-base junctions of bipolar input transistors to produce a rectifying action. Then, the transistor's junction capacitances store the envelope level of the RF signal at the amplifier input. If not disabled by this interference, the amplifier transmits an amplified replica of the envelope to the circuit output. FET input op amps significantly reduce the likelihood of this effect due to the larger voltage swing required to produce a rectifying action. Fortunately,

the lower input bias currents of FET amplifiers also make them the predominate choice for photodiode applications.

9.3 Reducing Multiple Coupled-Noise Effects

Coupled noise seldom provides clear clues to its origin, and successful reduction of this noise requires a modified, troubleshooting approach. In a given application, any one of the preceding noise reduction techniques may offer the best solution, or the application may require a combination of two or more. Without definitive clues, trial-and-error application of the noise reduction practices prevails. However, the combinational possibility requires a modification to this approach. Traditionally, when a given trial implementation fails to solve the problem, you remove this implementation and proceed to the next. However, with the potential for combinational effects, effective noise coupling reduction requires a delay in this removal. When the trial of a first noise reduction technique fails to solve the problem, do not remove it before applying the next technique. The effect of the first may be masked by the requirement for a second, and removing the first at this point may suboptimize the final solution. Instead, proceed through the series of noise reduction practices described above, leaving all implementations in place until the end. Then, with the coupling reduction optimized, begin to remove those implementations which originally appeared to have no effect. In the reduced noise background of the optimized condition, the actual noise effects of the individual implementations become apparent. Remove the ineffective ones at this point but not before.

9.4 Minimizing Magnetic Field Generation

The preceding discussions focus upon the photodiode amplifier's sensitivity to coupled noise, overlooking this amplifier's potential for producing such noise. Simple current flow in a conductor produces a magnetic field capable of coupling noise back into the amplifier or to other circuitry. Current flows associated with a photodiode amplifier typically lack the magnitude needed to produce a significant field, except at the amplifier output. There, the load current supplied by the amplifier potentially produces a magnetic noise source, but attention to coaxial returns minimizes this potential.

As illustrated in Fig. 9.7, the photodiode amplifier supplies the load current i_L to load resistance R_L , often producing a significant current flow in connecting conductors. Closer examination of this current's path reveals current flow in a field-generating loop. For the

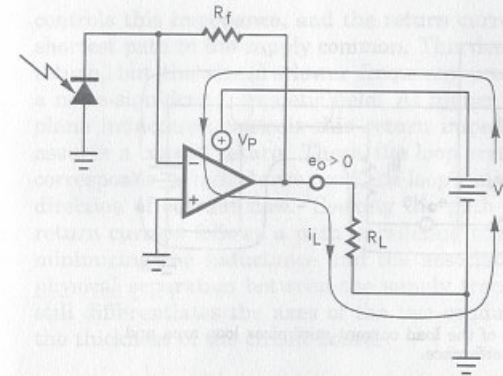


Figure 9.7 Load current supplied by the amplifier actually flows in a loop area capable of producing magnetic interference.

case shown, the amplifier produces a positive output voltage e_o by drawing the load current i_L from the V_+ positive supply and delivering it to R_L . Then, R_L 's ground connection returns the current to V_+ , completing the loop. A similar current conduction loop results for negative e_o values through the amplifier's negative power supply. In each case, the magnitude and frequency content of i_L and the physical area of the conduction loop determine the strength of the magnetic field generated. Application requirements dictate the characteristics of i_L , but physical layout options permit minimizing the loop area to greatly reduce the resulting field strength.

Coaxial conduction of the supply and return of the i_L current minimizes this loop area. For the simplest illustration of this principle, Fig. 9.8 shows this current conduction with a coaxial cable. Other, more practical options generally suffice, and later discussion presents these options using the principles illustrated here. In the figure, the cable's center conductor supplies the current to the amplifier, and the cable shield returns it to the power-supply common. Just the appearance of the resulting figure communicates the reduced area of the conduction loop, supplying an intuitive sign of magnetic field reduction. In practice, the relative area reduction far exceeds that presented by the short cable length of the illustration.

The significance of this area reduction becomes apparent when considering the magnetic fields produced by the supply and return conductors. To minimize loop area, the supply and return conductors must follow essentially the same physical path. These conductors carry the i_L current in opposite directions along this common path,

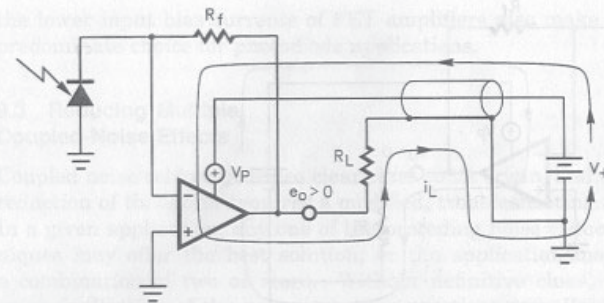


Figure 9.8 Coaxial return of the load current minimizes loop area and the associated magnetic interference.

producing equal magnetic fields of opposite orientation. These fields cancel, neutralizing the net magnetic field. The coaxial cable optimizes this canceling effect by producing a common centroid or axis for the two current flows. While difficult to illustrate, the return current conducted in the cable shield of Fig. 9.8 distributes throughout the shield, enveloping the center conductor. Averaged around the circular shield, the net magnetic field produced has a centroid coincident with the center conductor. Thus, the supply and return conductors of the coaxial cable permit precise magnetic field neutralization. However, coaxial cable connections represent an impractical solution to the multitude of current conduction loops found on a typical circuit board.

Fortunately, adjacent board traces or a ground plane provide nearly coaxial performance for the practical case. Adjacent circuit board traces illustrate this solution, but the ground plane best supplies it. Conceptually adjacent traces, supplying and returning a given current, approximate the coaxial condition. This configuration again minimizes loop area and only slightly compromises the ideal coaxial conduction of the cable above. With adjacent traces, the space separating the two differentiates the central axes of the two magnetic fields produced, slightly reducing the magnetic field cancellation. Still, the adjacent trace solution also remains impractical for all but the simplest of circuit boards. Layout implementation of a return trace adjacent to each signal trace soon becomes a product of conflicting demands.

In practice, a ground plane avoids the adjacent trace complication and provides similar coaxial performance. There, serendipity assures the coaxial return at the higher frequencies where magnetic coupling becomes increasingly significant.⁵ A return current delivered to the ground plane follows the path of least impedance back to the power-supply common. At lower frequencies, the ground plane resistance

controls this impedance, and the return current typically follows the shortest path to the supply common. This does not establish a coaxial return, but the signal's lower frequency prevents the production of a noise-significant magnetic field. At higher frequencies, the ground plane inductance controls this return impedance and automatically assures a coaxial return. There, the loop area of the conduction path corresponds to inductance, and the loop's magnetic field controls the direction of current flow. Seeking the path of least impedance, the return current follows a path paralleling that of the supply current, minimizing the inductance and the associated magnetic field. The physical separation between the supply trace and the ground plane still differentiates the axes of the two conduction paths but only by the thickness of the circuit board.

References

1. R. Morrison, *Grounding and Shielding Techniques in Instrumentation*, 2nd ed., John Wiley, New York, 1977.
2. H. Ott, *Noise Reduction Techniques in Electronic Systems*, John Wiley, New York, 1976.
3. J. Graeme, "Systematic Approach Makes Op-Amp Circuits Resist Radiated Noise," *EDN*, July 20, 1995, p. 93.
4. Y. Sutu and J. Whalen, "Statistics for Demodulation RFI in Operational Amplifiers," *IEEE International Symposium on Electromagnetic Compatibility*, August 23, 1983.
5. P. Browkaw and J. Barrow, "Grounding for Low- and High-frequency Circuits," *Analog Dialog*, 23-3, 1989.

Position-Sensing Photodiode Amplifiers

Photodiodes sense the position of a light beam through the diodes' photo responses, but several variables potentially degrade sensing accuracy. These variables introduce offset and gain errors in the light-to-voltage conversion process. A single photodiode provides some measure of a light beam's position through the magnitude of the diode's output signal. However, background light conditions also influence this magnitude and could require calibrated and often impractical measurement conditions. Fortunately, background light only adds an offset to the diode output, and differential measurement removes its effect. Adding a second, matching photodiode and monitoring the difference between the two diode outputs removes the equal offsets produced by the two. The differential-input nature of op amps readily extracts this difference signal through several circuit configurations.¹ There, two photodiodes provide one axis of position information, and replicating the two-diode solution provides multiaxis information.

Still, these position-sensing solutions remain vulnerable to variations in the light source intensity and the photodiode's responsivity. Both variables affect the gain of the position measurement, and simple differential monitoring does not remove gain error effects. Once again, calibrated measurement conditions can temporarily remove these effects, but light sources age and the photodiode responsivity displays a strong temperature sensitivity. Signal normalization provides a more permanent solution by dividing the diodes' difference signal by their mean signal. There, gain variations affect the mean and difference signals equally, making their normalized quotient immune to gain changes. An analog divider extracts this normalized signal with the greatest accuracy, but analog multiplier replacements for the divider

simplify the circuitry. Alternately, linear photodiode arrays remove the need for normalization by providing a digital indication of position. There, the diodes of the array need only indicate which diode resides closest to the incident beam as resolved by evaluating their relative rather than absolute signal magnitudes.

10.1 Direct-Displacement Monitor Amplifiers

In the simplest case, two photodiodes directly indicate a beam position through the difference in their output signals. Four circuits below produce the desired difference signal with varying degrees of circuit complexity and performance. Using three op amps, a differential-photodiode amplifier provides the best overall performance and offers simplified noise filtering. Alternately, connecting the photodiodes differentially reduces the amplifier requirement to the basic current-to-voltage converter. However, this simpler circuit compromises accuracy, offset, and noise. Modifying this simpler circuit bootstraps the photodiodes to restore offset and noise performance, but this modification potentially reduces bandwidth. These first three circuits provide single-axis monitoring with two photodiodes. A fourth circuit extends the monitoring to two axes using four photodiodes.

10.1.1 Single-axis monitoring with a differential-photodiode amplifier

While somewhat complex, the differential-photodiode amplifier of Fig. 10.1 develops the photodiode difference signal with the greatest accuracy. This circuit first converts the two photodiode currents into voltages and then subtracts those voltages. In the figure, A_1 and A_2 perform as traditional current-to-voltage converters, and A_3 performs as a difference amplifier. Fundamentally, the circuit becomes the common three-op-amp instrumentation amplifier² with the gain set resistor removed and photodiodes connected to the gain set points. The output voltages of A_1 and A_2 contain both difference and common-mode signals. A_3 separates the two, delivering the difference signal to the circuit output as $e_o = (i_{p1} - i_{p2})R_f$.

This difference signal reflects the displacement information desired through the difference in the two photodiode currents. Note that equal illumination of the two photodiodes makes $i_{p1} = i_{p2}$ and $e_o = 0$, indicating a beam centered between the two diodes. Similarly, greater illumination of D_1 makes $i_{p1} > i_{p2}$ and $e_o > 0$, indicating a displacement in the D_1 direction, and the converse conditions result for greater illumination of D_2 . Currents i_{p1} and i_{p2} respond linearly to illumi-

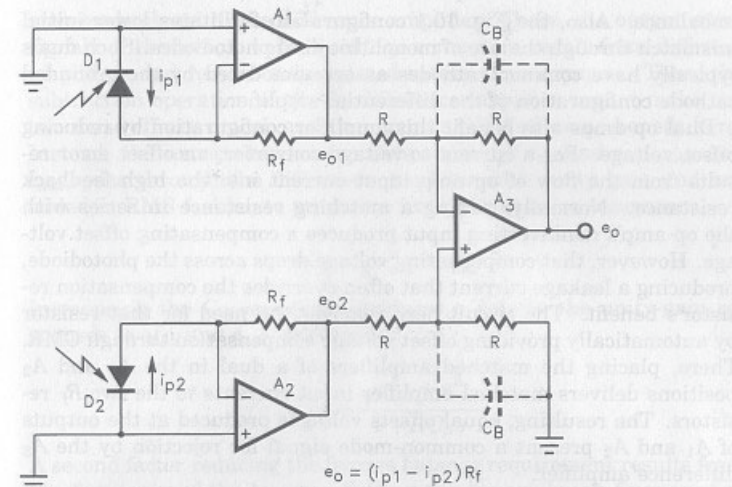


Figure 10.1 The differential-photodiode amplifier resolves the position of a light source relative to two photodiodes as represented by the difference in diode output currents.

nation intensity and make the magnitude of e_o a direct measure of displacement.

Evaluation of the e_o equation confirms the circuit's removal of background light source effects. A photodiode produces an output current of $i_p = r_\phi \phi_e$, where r_ϕ is the diode's flux responsivity and ϕ_e is the radiant flux in watts.³ The combination of signal and background light sources makes $\phi_e = \phi_{es} + \phi_{eB}$ and $i_p = r_\phi(\phi_{es} + \phi_{eB})$. Applying this i_p expression to $e_o = (i_{p1} - i_{p2})R_f$ produces $e_o = [(r_{\phi1}\phi_{es1} - r_{\phi2}\phi_{es2}) + (r_{\phi1} - r_{\phi2})\phi_{eB}]R_f$. Here, the term $(r_{\phi1} - r_{\phi2})\phi_{eB}R_f$ describes the residual offset produced by the background flux ϕ_{eB} that illuminates the two diodes equally. Simply matching the photodiode responsivities, $r_{\phi1} = r_{\phi2}$, reduces this offset to zero, leaving $e_o = (r_{\phi1}\phi_{es1} - r_{\phi2}\phi_{es2})R_f$.

Circuit mismatches produce error in the actual e_o signal, but resistor adjustments produce compensating effects. Specifically, the two photodiodes will not have identical responsivities, and the high value of the R_f resistors makes their matching difficult. Both effects produce gain imbalance in the circuit's two signal channels, and this disturbs the circuit's difference function. Adjusting the R_f resistors would compensate for both effects, but again, the high resistance value makes such adjustment difficult. Instead, adjusting the more reasonable R resistors of the difference amplifier produces a compensating gain

imbalance. Also, the Fig. 10.1 configuration facilitates lower initial mismatch through the use of monolithic dual photodiodes. Such duals typically have common cathodes as accommodated by the grounded cathode configuration of the differential amplifier.

Dual op amps also benefit this amplifier configuration by reducing offset voltage. For a current-to-voltage converter, an offset error results from the flow of op amp input current into the high feedback resistance. Normally, adding a matching resistance in series with the op amp's noninverting input produces a compensating offset voltage. However, that compensating voltage drops across the photodiode, producing a leakage current that often overrides the compensation resistor's benefit. The circuit here removes the need for that resistor by automatically providing offset voltage compensation through CMR. There, placing the matched amplifiers of a dual in the A_1 and A_2 positions delivers matched amplifier input currents to the two R_f resistors. The resulting, equal offsets voltages produced at the outputs of A_1 and A_2 present a common-mode signal for rejection by the A_3 difference amplifier.

10.1.2 Noise reduction with the differential-photodiode amplifier

This amplifier permits reducing the noise effects of both internal and external noise sources. Simplified filtering reduces the effects of the amplifier's noise sources, and CMR reduces the effects of noise coupled from external sources. For a different application, Secs. 9.1.4 and 9.1.5 develop detailed noise analyses for both noise effects of the differential amplifier, and those analyses apply to this application as well.

The simplified filtering opportunity results from a difference in signal and noise bandwidths. As described in Chap. 6, the current-to-voltage converter often delivers a greater bandwidth to internal noise sources than to the signal, but filtering can restrict both bandwidths to the same frequency and remove noise without affecting signal. The previous discussion also presents several such filtering methods for the basic current-to-voltage converter. Any of these methods could be applied to the two current-to-voltage converters of this circuit; however, this circuit's third op amp offers a simpler solution. As shown with dashed lines, simple capacitance bypass of two resistors of the difference amplifier produces the desired filtering. This adapts the difference amplifier to also serve as a low-pass filter and avoids the added complexity of previous solutions.

Such bypass can disturb the circuit balance that produces the difference amplifier's CMR, and ideally, matching bypass responses are required. However, two factors make this matching less critical here. First, the dominant noise sources of the circuit do not present a com-

mon-mode signal to the difference amplifier. There, the output noise signals of the two current-to-voltage converters result from different amplifiers and remain uncorrelated rather than common. Thus, the subtraction operation of the A_3 difference amplifier cannot make the two noise effects cancel, and the circuit's CMR offers no benefit here. Instead, the uncorrelated noise components of e_{o1} and e_{o2} , e_{no1} and e_{no2} , pass through the difference amplifier with unity gain and are added in RMS fashion to produce a final output noise component

$$e_{no3} = \sqrt{e_{no1}^2 + e_{no2}^2}$$

Introducing the C_B capacitors produces a pole in the unity-gain responses to e_{no1} and e_{no2} , making

$$e_{no3} = \frac{\sqrt{e_{no1}^2 + e_{no2}^2}}{1 + RC_B s}$$

A second factor reducing the bypass balance requirement results from the frequency of the bypass. Setting the bypass break frequency at the signal bandwidth limit moves the primary effect of the bypass to frequencies beyond the range of interest. Still, the circuit depends upon CMR for accurate development of the difference signal, and the effect of bypass imbalance should be evaluated when using this option.

Note that another obvious filtering option does not yield an equivalent result. Capacitive bypass of the R_f conversion resistance would also reduce noise bandwidth. However, as described in Sec. 6.1.1, this bypass simultaneously reduces signal bandwidth by the same factor, leaving the ratio of the two bandwidths unchanged. Then, signal bandwidth declines, and noise still retains the bandwidth advantage.

The differential-photodiode amplifier also aids in the reduction of noise coupling from external sources. Electrostatic and magnetic coupling can be significant noise contributors especially because of the small signals typically developed by photodiodes. The differential nature of this circuit converts those coupling effects into common-mode signals for rejection by the difference amplifier. Chapter 9 describes this noise reduction in detail. Note that this reduction depends upon the common-mode nature of the coupled signals and the circuit's CMR. Thus, the filtering described above potentially compromises the reduction of coupled noise effects as well.

10.1.3 Connecting the photodiodes differentially

With their voltages held to zero by op amp inputs, photodiodes become two-terminal current sources, and current signals sum or subtract

easily. This permits a simpler monitor circuit but introduces some performance compromises. Simply connecting two photodiodes in parallel and with opposite orientations in Fig. 10.2 produces the differential output current $i_{p1} - i_{p2}$. Common-mode current merely circulates in the diode loop resulting in no signal output. Then, the simple current-to-voltage converter again serves as the photodiode amplifier. This circuit produces the same $e_o = (i_{p1} - i_{p2})R_f$ as the preceding differential-photodiode amplifier.

However, this simpler circuit also degrades response accuracy, offset, and the rejection of coupled noise. Response accuracy degrades due to the different cathode connections of the diodes. Previously, the grounded cathodes of the two photodiodes permitted the use of monolithic devices that ensure matched responsivities for the two. Offset degrades because the offset compensation resistor R_C now develops a voltage across the photodiodes. That resistor produces a dc offset voltage with the amplifier's noninverting input current to compensate for the voltage developed by the inverting input current with the feedback resistor. However, the voltage on the compensation resistor also appears across the photodiodes, inducing diode leakage currents. These leakages flow through the feedback resistor and produce an output offset. Previously, the differential amplifier held both diode voltages at zero and achieved offset compensation through CMR without the need for compensation resistors. Rejection of coupled noise also degrades, again due to the lack of CMR that the previous circuit used to remove this noise effect.

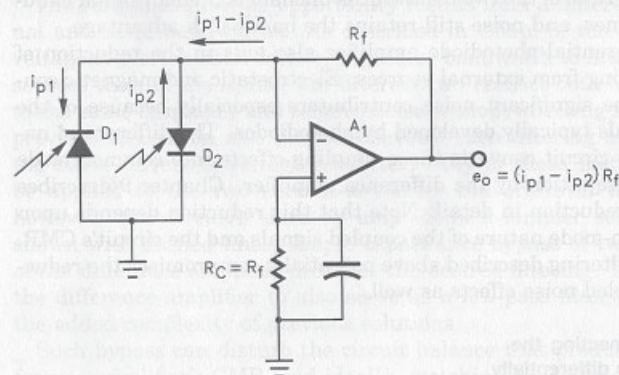


Figure 10.2 Connecting two photodiodes in parallel and with opposite orientations automatically produces a differential signal, reducing the photodiode amplifier to the basic current-to-voltage converter.

10.1.4 An alternate differential-photodiode connection

A simple modification to the Fig. 10.2 circuit restores the previous offset and noise rejection performance at the expense of bandwidth. Shown in Fig. 10.3, the photodiodes now return to the op amp's non-inverting input rather than ground. Previously, Sec. 9.1.2 introduced this circuit for a single-diode application. The accompanying noise analysis of Sec. 9.1.3 applies to this case as well. Here, the modified diode return again bootstraps the diodes upon the voltage developed across the compensation resistor R_C . Now, the diodes connect directly across the zero voltage of the op amp's differential inputs, removing diode leakage and the associated offset result. This modified circuit also removes the Fig. 10.2 bypass capacitor that rolled off the noise of R_C there. Then, R_C participates in the current-to-voltage conversion and restores CMR. The $i_{p1} - i_{p2}$ return current of the diodes now flows through R_C , producing a second component of output signal e_o . For its compensation purpose, $R_C = R_f$, and the added e_o component doubles the circuit's current-to-voltage gain. Also, the balanced impedances now connected to the op amp's inputs restore CMR of coupled noise. Electrostatic noise currents coupled to the amplifier's two inputs produce equal and canceling voltages on the circuit's two resistors. Section 9.1.2 describes this noise cancellation in detail.

However, this alternate differential-photodiode circuit potentially reduces bandwidth. Two factors affect bandwidth here, with the first

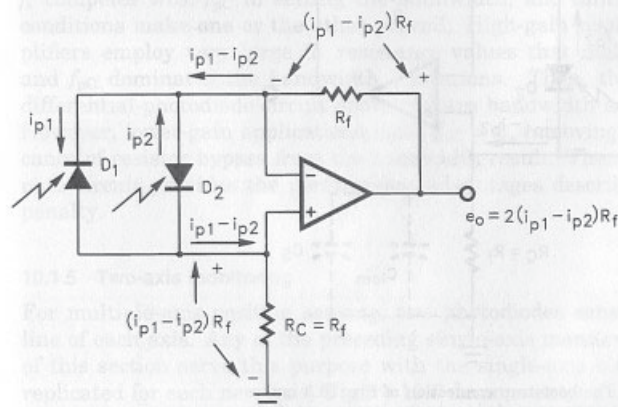


Figure 10.3 Bootstrapping the photodiodes removes Fig. 10.2's major performance compromises but potentially reduces bandwidth.

producing an increase but with the second producing a greater decrease. Bandwidth tends to increase due to the doubled current-to-voltage gain $2R_f$ described above. This permits reducing R_f a factor of 2 to produce the same gain as the preceding circuits. Then, parasitic capacitance bypass of the R_f resistances presents a bandwidth limit at twice the frequency experienced before. However, this alternative connection also adds another bypass capacitance across the compensation resistance as shown in Fig. 10.4. There, the inevitable C_S stray capacitances bypass both resistors, but the op amp adds its common-mode input capacitance C_{icm} to the bypass of R_C . Typically, $C_{icm} \approx 6C_S$, and the combined bypass of R_C rolls off its gain contribution at about one-seventh the frequency of the C_S bypass of R_f . The latter bypass often determines the bandwidth of the photodiode amplifier, so the C_{icm} effect here potentially reduces bandwidth a factor of 7. Combined, the two bandwidth effects potentially produce about a 3.5:1 bandwidth reduction for the alternative differential-photodiode circuit.

The actual frequency response of this circuit exhibits two gain plateaus instead of the customary single plateau. For the typical $R_C = R_f$ case, Fig. 10.5 illustrates this response starting with a low-frequency I -to- V gain plateau at $2R_f$. At an intermediate frequency,

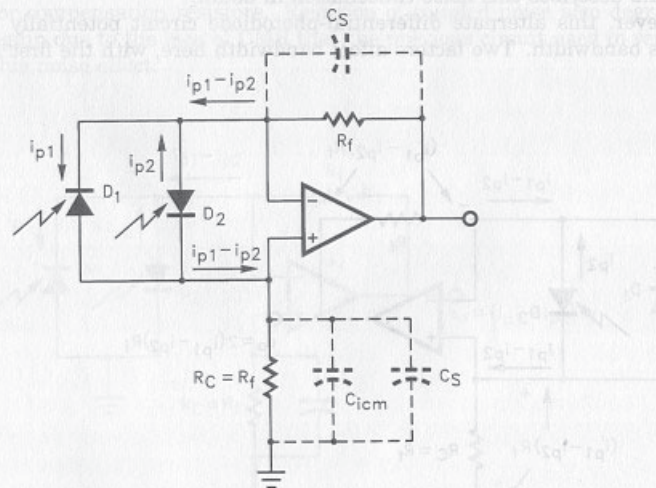


Figure 10.4 The bootstrap connection of Fig. 10.3 makes the amplifier's common-mode input capacitance part of the bandwidth-limiting bypass of current-to-voltage conversion resistance.

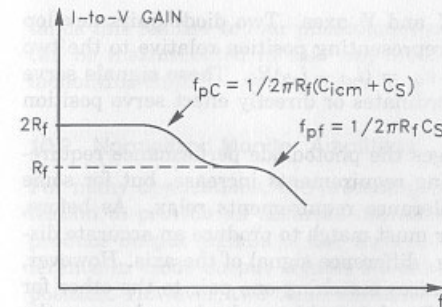


Figure 10.5 The added C_{icm} bypass of Fig. 10.4 introduces a second response pole at f_{pC} that potentially dominates the circuit's bandwidth.

C_{icm} shunts $R_C = R_f$ and produces a response pole at $f_{pC} = 1/2\pi R_f (C_{icm} + C_S)$. This shunt totally removes the gain contribution of R_C but that of R_f remains, interrupting the first gain roll off with a second plateau at a level equal to R_f . At a higher frequency, the stray C_S shunts R_f , producing the second roll off at $f_{pf} = 1/2\pi R_f C_S$. Technically, the second plateau extends the frequency response, but this extension typically serves little practical purpose. For most applications, the roll off at f_{pC} marks the useful bandwidth limit as set by resistor bypass.

However, this limit alone does not necessarily define the final bandwidth. As described in Chap. 3, other limits potentially set the actual bandwidth of the photodiode amplifier. A second limit occurs at the $1/\beta$ intercept with the A_{OL} response at the frequency f_i . Then, f_i competes with f_{pC} in setting the bandwidth, and different circuit conditions make one or the other prevail. High-gain photodiode amplifiers employ very large R_f resistance values that make $f_{pC} < f_i$, and f_{pC} dominates the bandwidth limitations. Then, the alternate differential-photodiode circuit above reduces bandwidth as described. However, lower-gain applications make $f_{pC} > f_i$, removing the significance of resistor bypass from the bandwidth result. Then, this alternate circuit provides the performance advantages described without penalty.

10.1.5 Two-axis monitoring

For multiple-axis position sensing, two photodiodes sense along the line of each axis. Any of the preceding single-axis monitoring circuits of this section serve this purpose with the single-axis circuit simply replicated for each new axis. Figure 10.6 demonstrates this extension for two-axis monitoring using the Fig. 10.1 differential-photodiode amplifier. There, the graphical representation of the diodes depicts their

physical placements along X and Y axes. Two diode pairs develop independent output signals representing position relative to the two axes, $e_{ox} = (i_{x1} - i_{x2})R_f$ and $e_{oy} = (i_{y1} - i_{y2})R_f$. These signals serve as indicators of X and Y coordinates or directly effect servo position control.

Two-axis monitoring changes the photodiode performance requirements in two ways. Matching requirements increase, but for some applications, responsivity tolerance requirements relax. As before, the two diodes of an axis pair must match to produce an accurate displacement indication in the e_o difference signal of the axis. However, two-axis monitoring also requires matching one pair to the other for vector angle measurement. Resolving this angle depends upon the ratio of the X and Y signals, and interpair mismatch produces angular error. However, this ratio remains immune to the highly variable diode responsivity, potentially relaxing a requirement there. Responsivity variations due to manufacturing tolerances and temperature changes do not disturb this ratio as long as all diodes experience the same variation. This eases the diode requirements for applications requiring only angular information.

Monolithic photodiode arrays best serve the more stringent matching requirement of this case. As described before, the Fig. 10.1 differential-photodiode amplifier facilitates its use through the grounded connection of the diode cathodes. With both cathodes connected to the same point, that circuit accommodates the common-cathode structure of monolithic photodiodes. Using the Fig. 10.1 amplifier, Fig. 10.6 ex-

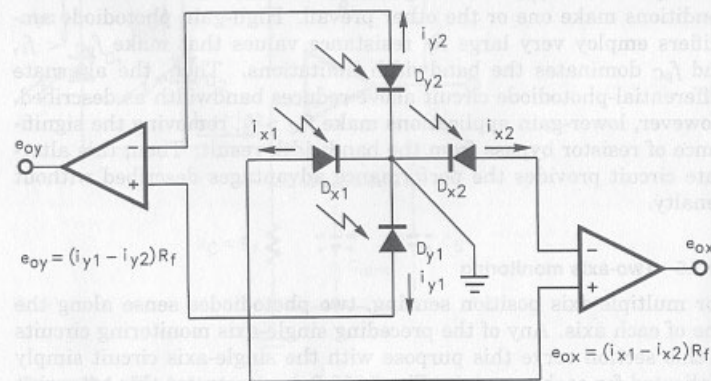


Figure 10.6 Four photodiodes provide two-axis displacement monitoring, and the differential-photodiode amplifier permits the common-cathode connection of a monolithic array.

tends this feature to four photodiodes as shown. Any number of axes can be instrumented in this way with all photodiodes on the same monolithic chip.

10.2 Normalized Monitor Amplifiers

For many applications, the preceding circuits require signal normalization to provide an accurate indication of position. These circuits produce output signals of the form $e_o = (i_{p1} - i_{p2})R_f$, where a difference in diode output signals should directly reflect a light beam's position. However, two other variables typically affect this difference signal by introducing gain error in the light-to-voltage conversion process. Variations in both the light intensity received and the photodiode responsivity to light also affect the magnitudes of the i_{p1} and i_{p2} currents. Received light intensity varies due to changes in the light source and in the medium separating the source from the diodes. Photodiode responsivity varies with manufacturing tolerances and temperature changes. The resulting magnitude variations of i_{p1} and i_{p2} also alter the $i_{p1} - i_{p2}$ difference of the e_o equation, confounding the position sensing. This potentially reduces the e_o difference signal to a relative rather than absolute indication, where e_o only reflects which diode is closest to the source.

To quantify these effects, consider the $e_o = (i_{p1} - i_{p2})R_f$ equation expressed in terms of the light intensity and responsivity variables. The Sec. 10.1.1 discussion reduced this equation to $e_o = (r_{\phi1}\phi_{es1} - r_{\phi2}\phi_{es2})R_f$, where r_{ϕ} represents photodiode diode flux responsivity and ϕ_e represents radiant flux. Consider the case of matched photodiodes where $r_{\phi1} = r_{\phi2} = r_{\phi}$, making $e_o = (\phi_{es1} - \phi_{es2})r_{\phi}R_f$. Clearly, a variation in responsivity r_{ϕ} alters gain and changes e_o . Similarly, a variation in light source intensity alters ϕ_{es1} and ϕ_{es2} by the same proportion to also change gain. Thus, both variables potentially make e_o an inaccurate indicator of absolute position. Calibration can remove the effects of these extra variables, but light source aging and diode responsivity changes with temperature make calibration a temporary condition. Some tracking applications only require relative position information because servo adjustments only need to reduce the difference signal to a null. However, the added variables in this signal then require slowing the servo response to prevent the overshoot and ringing otherwise produced by gain increases.

Alternately, signal normalization automatically removes the effects of the added variables to reestablish absolute rather than relative position indication. Normalizing the difference signal against the mean signal,⁴ $(i_{p1} - i_{p2})/(i_{p1} + i_{p2})$, divides the difference signal by a signal having equal sensitivity to the added variables. Then,

variations in the light intensity or photodiode responsivity change the i_{p1} and i_{p2} variables in both the numerator and denominator of the normalized expression for a net zero result. For example, doubling the responsivity of the diodes replaces an original i_{p1} and i_{p2} with $2i_{p1}$ and $2i_{p2}$, and the change produces canceling effects in the normalized expression above. Three circuits below produce the normalized output with differing degrees of precision and circuit complexity. The first provides the best performance using four op amps and a dedicated analog divider. Next, substituting the less accurate but more common multiplier for the divider eases component selection without increasing circuit complexity. Finally, capitalizing upon circuit summing capabilities reduces the circuit to two op amps and a multiplier. However, this simpler circuit further compromises performance.

10.2.1 Normalizing the differential-photodiode amplifier

The straightforward and most accurate approach to normalization combines the differential-photodiode amplifier with an analog divider as in Fig. 10.7. There, A_1 through A_3 and their associated resistors form the differential amplifier discussed with Fig. 10.1, and A_4 serves as an inverting summer. Together, the amplifier and summer drive the divider's N and D numerator and denominator inputs, and the divider responds with an output response of $10N/D$. The differential amplifier supplies the required difference signal $(i_{p1} - i_{p2})R_f$ directly from the output of A_3 . To derive the mean signal, the summer adds and inverts the signals at the outputs of A_1 and A_2 , producing $(i_{p1} + i_{p2})R_f$. Then, the divider produces the normalized output

$$e_o = 10 \frac{(i_{p1} - i_{p2})}{(i_{p1} + i_{p2})}$$

As an additional benefit, that the divider introduces a gain of 10 to this result.

Evaluation of the e_o equation confirms the circuit's removal of the effects of variations in light source and photodiode responsivity. As described in Sec. 10.1.1, a photodiode produces an output current of $i_p = r_\phi \phi_e$, where r_ϕ is the diode's flux responsivity and ϕ_e is the radiant flux. Applying this i_p expression to the above e_o produces

$$e_o = 10 \frac{(r_{\phi 1} \phi_{e1} - r_{\phi 2} \phi_{e2})}{(r_{\phi 1} \phi_{e1} + r_{\phi 2} \phi_{e2})}$$

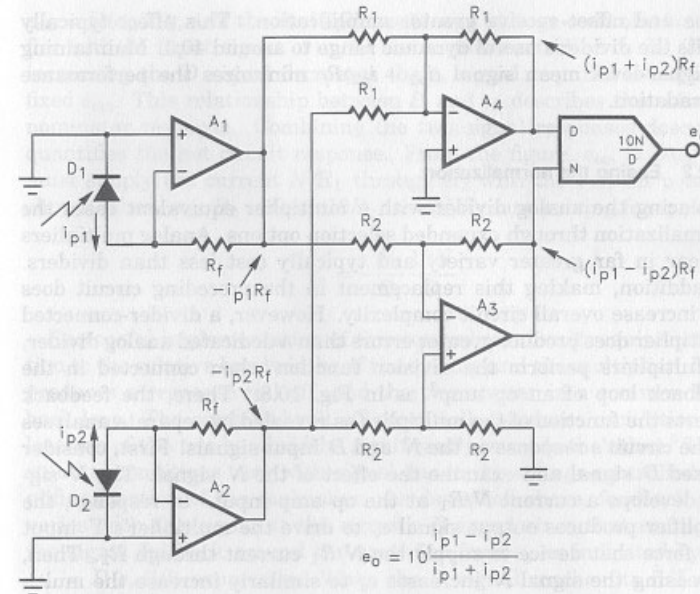


Figure 10.7 Adding an analog divider and a summing amplifier to the differential-photodiode amplifier produces a differential output signal normalized against the mean signal.

Then, assuming responsivity matching for the two diodes, $r_{\phi 1} = r_{\phi 2}$, removes the effects of responsivity manufacturing tolerance and temperature variation in

$$e_o = 10 \frac{(\phi_{e1} - \phi_{e2})}{(\phi_{e1} + \phi_{e2})}$$

There, any change in light source intensity alters ϕ_{e1} and ϕ_{e2} by the same proportion and adds the same multiplier factor to the numerator and denominator of the above expression for a net zero change. For example, doubling the light intensity replaces ϕ_{e1} and ϕ_{e2} with $2\phi_{e1}$ and $2\phi_{e2}$ but leaves e_o above unchanged.

However, the divider added for normalization degrades the circuit's linearity, noise, and offset performance, restricting the dynamic range of the normalized response. Typically, the divider's high noise and offset become dominate error sources, and their effects increase as the denominator signal magnitude decreases. There, a denominator decrease raises the divider's gain of $10N/D$, and the divider's input

noise and offset receive greater amplification. This effect typically limits the divider's useful dynamic range to around 40:1. Maintaining a higher-level mean signal $(i_{p1} + i_{p2})R_f$ minimizes the performance degradation.

10.2.2 Easing the normalization

Replacing the analog divider with a multiplier equivalent eases the normalization through expanded selection options. Analog multipliers appear in far greater variety and typically cost less than dividers. In addition, making this replacement in the preceding circuit does not increase overall circuit complexity. However, a divider-connected multiplier does produce greater errors than a dedicated analog divider.

Multipliers perform the division function when connected in the feedback loop of an op amp³ as in Fig. 10.8. There, the feedback inverts the function of the multiplier as revealed by separate analyses of the circuit's response to the N and D input signals. First, consider a fixed D signal and examine the effect of the N signal. The N signal develops a current N/R_1 at the op amp input. In response, the amplifier produces output signal e_o to drive the multiplier's Y input and force that device to supply the N/R_1 current through R_2 . Then, increasing the signal N increases e_o to similarly increase the multiplier output signal e_{om} . This relationship between N and e_o describes that of a numerator response. Next, consider a fixed N signal and examine the effect of the D signal. A fixed N signal requires a fixed e_{om} output from the multiplier as seen from the preceding discussion. In response to the D signal, the multiplier produces an output voltage

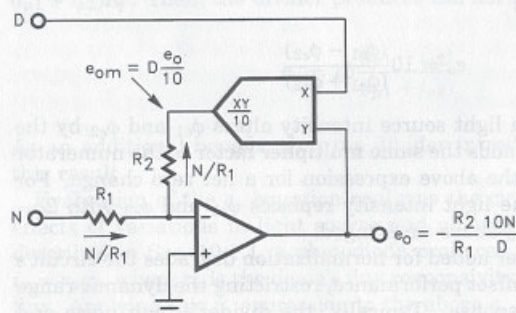


Figure 10.8 Connecting an analog multiplier in the feedback loop of an op amp produces a divider response for replacement of the Fig. 10.7 divider with a more commonly available component.

$e_{om} = D(e_o/10)$, and the multiplier acts as a voltage-controlled amplifier with an input signal of $e_o/10$ and a gain controlled by D . Then, increasing the D signal decreases the e_o signal required to supply a fixed e_{om} . This relationship between D and e_o describes that of a denominator response. Combining the two signal responses described quantifies the net circuit response. From the figure, $e_{om} = D(e_o/10)$ must supply the current N/R_1 through R_2 with the current polarity shown. Then, $e_{om}/R_2 = D(e_o/10R_2) = -N/R_1$, and solving for e_o yields

$$e_o = -\frac{R_2}{R_1} \frac{10N}{D}$$

Successful implementation of this equivalent circuit requires attention to the relative bandwidths of the op amp and the multiplier. Each produces a response roll off, and the two appear in a common feedback loop. The combined roll off potentially produces oscillation unless one of the two bandwidths significantly exceeds the other.⁵ Here, stability conditions vary with signal due to the voltage-controlled amplifier function of the multiplier. As described above, feedback gain varies with the signal at the circuit's D input since that multiplies the signal fed back around the amplifier from e_o as indicated by the $e_{om} = D(e_o/10)$ expression. When the D signal reaches its full-scale level, typically 10 V, the gain to the e_o signal reaches its maximum of unity and frequency stability reaches its greatest vulnerability. Stability should be evaluated under this condition. Typically, multipliers achieve greater bandwidths than op amps, and choosing a multiplier with a bandwidth well beyond the f_c crossover of the op amp assures a single, dominant pole for all levels of feedback gain provided by the multiplier.

Above, the divider-connected multiplier produces the response $e_o = -(R_2/R_1)10N/D$. Note the gain inversion provided by the minus sign and the added degree of gain control provided by the R_2/R_1 ratio. Both features simplify the divider replacement in the normalized differential-photodiode amplifier. As described above, the multiplier equivalent of the analog divider requires placing the multiplier in the feedback loop of an op amp. This would seem to add an op amp to a circuit, but the added gain inversion and gain control avoid this for the differential amplifier.

Shown in Fig. 10.9, the modified circuit still uses just four op amps because the circuit no longer requires the previous inverting summing amplifier. Previously, that amplifier developed the mean signal $(i_{p1} + i_{p2})R_f$ from the outputs of A_1 and A_2 and supplied the divider's denominator input. Here, the combination of the multiplier A_4 and the R_3 , $2R_3$ resistances form the multiplier equivalent of the previous

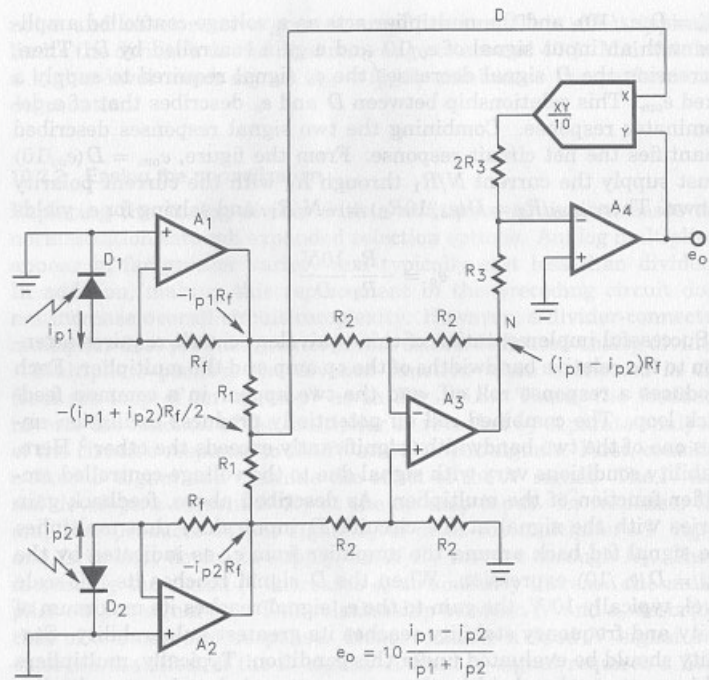


Figure 10.9 Replacing the Fig. 10.7 analog divider with a multiplier adds the A_4 feedback amplifier but eliminates the previous summing amplifier, leaving circuit complexity unchanged.

divider and summer. The equivalent denominator input D senses the center tap of the R_1 voltage divider and that tap produces the signal $-(i_{p1} + i_{p2})R_f/2$. This signal differs from the desired mean through a minus sign and a factor of 2 divisor. However, the gain inversion of the multiplier equivalent neutralizes the minus sign, and the R_3 , $2R_3$ resistor network cancels the factor of 2. As a result, the Fig. 10.9 circuit repeats the previous result and delivers the normalized output

$$e_o = 10 \frac{(i_{p1} - i_{p2})}{(i_{p1} + i_{p2})}$$

However, the factor of 2 divisor of $D = -(i_{p1} + i_{p2})R_f/2$ does decrease the denominator signal, increasing noise and offset effects as described at the beginning of this section.

10.2.3 Simplifying the normalized differential-photodiode amplifier

Two circuit summation opportunities greatly simplify this amplifier at the expense of some performance characteristics. Shown in Fig. 10.10, this option requires only two op amps instead of the four of the preceding circuits. Both photodiodes again connect directly across the inputs of op amps for the preferred zero diode bias. This circuit also retains canceling effects for dc offsets and coupled noise through the matching R_f conversion resistors connected to the two amplifier inputs. There, amplifier input currents and electrostatically coupled noise currents produce canceling error effects. However, the photodiodes no longer share the common-cathode connection that previously permitted use of matched, monolithic dual devices. Also, as will be seen, this circuit further reduces the D denominator signal, degrading noise and offset performance even more.

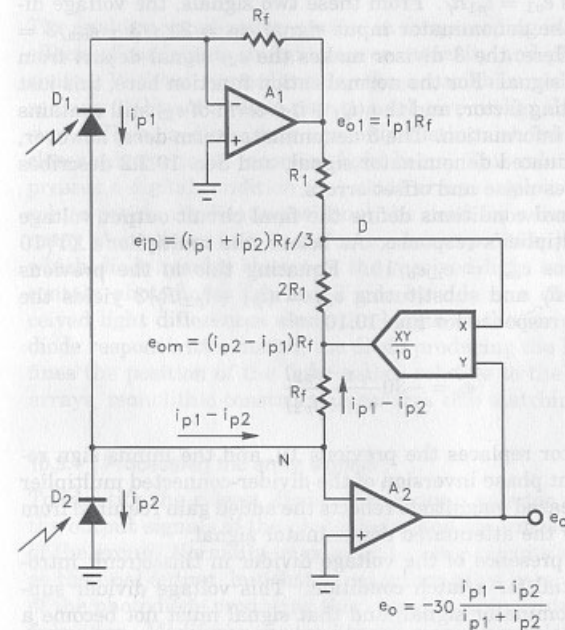


Figure 10.10 Connecting the photodiodes together and adding the R_1 , $2R_1$ voltage divider eliminates two of the circuit's four op amps through unique summation opportunities.

Circuit combinations simplify the summations required to generate the difference and mean signals required for normalization. Simply a diode interconnection and a voltage divider provide these summations. In the figure, the multiplier A_2 and the lower R_f resistor again produce the divider function but without the customary numerator input resistor. That resistor would normally be connected between the inverting input of A_2 and the circuit's difference signal voltage. However, current drive removes the need for that resistor here. Connecting the anode of D_1 to the cathode of D_2 results in current summation at their junction. The resulting current, $i_{p1} - i_{p2}$, supplies the numerator input N with the difference signal.

Next, the circuit's mean signal must be developed to drive the denominator input D . The $R_1, 2R_1$ voltage divider derives this signal by summing components of the multiplier and A_1 output signals e_{om} and e_{o1} . The multiplier output signal results from the flow of the $i_{p1} - i_{p2}$ current in the lower R_f resistor and $e_{om} = (i_{p2} - i_{p1})R_f$. Amplifier A_1 performs as a current-to-voltage converter monitoring the current of D_1 and produces $e_{o1} = i_{p1}R_f$. From these two signals, the voltage divider produces the denominator input signal $e_{ID} = 2e_{o1}/3 + e_{om}/3 = (i_{p1} + i_{p2})R_f/3$. Here, the 3 divisor makes the e_{ID} signal depart from the actual mean signal. For the normalization function here, this just introduces a scaling factor, and the $(i_{p1} + i_{p2})$ term of e_{ID} still contains the mean signal information. The 3 denominator term does, however, indicate an attenuated denominator signal, and Sec. 10.2.2 describes how this increases noise and offset errors.

The above signal conditions define the final circuit output voltage through the multiplier's response. As before, the multiplier's $XY/10$ response produces $e_{om} = e_{ID}e_o/10$. Equating this to the previous $e_{om} = (i_{p2} - i_{p1})R_f$ and substituting $e_{ID} = (i_{p1} + i_{p2})R_f/3$ yields the final normalized response for Fig. 10.10

$$e_o = -30 \frac{(i_{p1} - i_{p2})}{(i_{p1} + i_{p2})}$$

Here, a -30 factor replaces the previous 10, and the minus sign reflects the inherent phase inversion of the divider-connected multiplier circuit. The increased magnitude reflects the added gain required from the multiplier by the attenuated denominator signal.

Note that the presence of the voltage divider in this circuit introduces the potential for a latch condition. This voltage divider supplies the e_{ID} denominator signal, and that signal must not become a negative voltage for feedback polarity reasons. Positive e_{ID} values assure a positive gain through the multiplier's $e_{om} = e_{ID}e_o/10$ response, and this retains a net negative feedback around the op amp. How-

ever, negative values of e_{ID} reverse the polarity of the multiplier's gain, producing positive feedback and the latch condition. In normal operation, the $e_{ID} = (i_{p1} + i_{p2})R_f/3$ assures positive voltages because i_{p1} and i_{p2} can only be positive values. The positive voltage delivered to the voltage divider by A_1 and the voltage divider's ratio assure this normal condition. However, transient conditions during turn on potentially produce a momentary negative voltage at the D input. This does not significantly affect the preceding normalized circuits because the low impedances of op amp outputs drive that input and restore the D signal after the transient. However, in this case, the resistance of the voltage divider may permit a condition in which a negative multiplier output voltage sustains a negative e_{ID} voltage. Should this occur, capacitively coupling an initialization voltage from the positive power supply to the D input prevents the latch condition.

10.3 A Digital Alternative to Normalization

The analog normalization applied in the preceding section removes the effects of variations in light source intensity and photodiode responsivity. Without normalization, those variables reduce the position-sensing ability of a diode pair to a relative rather than absolute measure. Then, a diode pair can only distinguish which diode resides closer relative to an incoming light beam. This relative indication expresses a digital condition that diode arrays exploit to produce an absolute position indication without the need for normalization. There, many photodiodes in a linear array detect position just by identifying which diode resides closest to the received light beam. Such arrays contain photodiodes fabricated side by side in a strip and detect received light differences along the length of the strip. As long as the diode responsivities match, the diode producing the largest output defines the position of the light source relative to the array. For these arrays, monolithic construction assures this matching.

10.3.1 Processing the array signals

To identify the closest diode, a maximum selector circuit⁶ compares the output signals of the photodiodes and identifies the largest signal of the group. Normally, maximum selector circuits deliver this signal as the final output, but that is not informative here. Only the identity of the photodiode producing this signal communicates the position information. Modifying the traditional maximum selector produces this indication as shown in Fig. 10.11. There, separate monitor amplifiers receive the output voltages of the individual photodiodes, and the

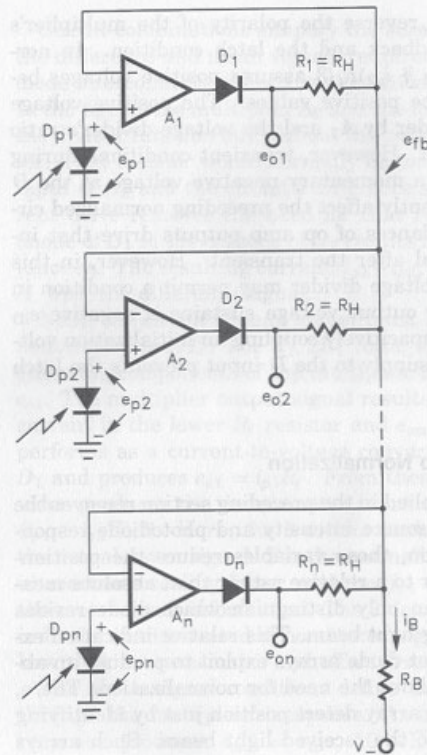


Figure 10.11 A modified maximum selector monitors a photodiode array, producing a series of digital output signals that indicate a light beam's position on the array by the diode producing the high output state.

amplifiers control separate output signals, labeled e_{o1} through e_{on} . All of the photodiodes have grounded cathodes, and this common return permits the use of the common-cathode configuration characteristic of monolithic arrays.

Also, all of the op amps have their inverting inputs connected to a common feedback-sensing line. This common connection produces the signal magnitude discrimination. To this line, each amplifier acts as a voltage follower that would make the line's signal e_{fb} track that amplifier's photodiode voltage. Each amplifier would require this to maintain zero voltage between its own inputs. However, the photodiode voltages will differ, and only one amplifier can achieve this feedback control, as determined by the signal diodes connected in series with the amplifier outputs. Those diodes form a gating net-

work that lets the largest amplifier output take control. Bias resistor R_B connects to the negative supply and would pull e_{fb} to that level except for the counteracting i_B current supplied by an amplifier output. Any of the amplifiers can supply such current to raise the e_{fb} voltage but none can reduce it. To raise this voltage an amplifier supplies a positive i_B current to R_B through the corresponding output diode. Reducing e_{fb} from this level would require a negative amplifier output current and the output diodes D_1 through D_n block such currents. Thus, the amplifier demanding the highest e_{fb} level prevails and sets the voltage of the feedback sensing line. That makes e_{fb} greater than the levels that would satisfy feedback for the other amplifiers. Then, the high open-loop gains of those amplifiers drive their outputs negative, reverse biasing the corresponding output diodes. Then, only the dominant amplifier retains a positive output voltage.

10.3.2 Defining the logic output levels

For this digital application, the circuit's other components translate the amplifier output signals into digital signal levels at the e_{o1} through e_{on} output terminals. These terminals assume low and high digital output levels as determined by a diode voltage drop and a voltage divider. Figure 10.12 illustrates these levels with a two-amplifier simplification of the circuit. For this example, assume $e_{p2} > e_{p1}$ as shown. Then, amplifier A_2 prevails in the control of e_{fb} , producing $e_{fb} = e_{p2}$. This makes $e_{fb} > e_{p1}$ and produces the voltage $e_{p2} - e_{p1}$ between the inputs of A_1 . A_1 responds, driving its output voltage negative and reverse biasing D_1 . With D_1 in the off state, A_1 operates open-loop, and the high open-loop gain drives this amplifier's output to its negative saturation level near V_- . Also with D_1 off, the current in R_1 is zero, and $e_{o1} = e_{p2} \approx 0.6$ V, representing a low logic level. In contrast, the prevailing amplifier A_2 supplies the current $I_B = (e_{p2} - V_-)/R_B$ through R_2 and produces a high logic level at its corresponding output terminal

$$e_{o2} = e_{p2} + \frac{e_{p2} - V_-}{R_B} R_H$$

Thus, a voltage difference of $(R_H/R_B)(e_{p2} - V_-)$ separates the two logic levels. Note that V_- here will be a negative value, and its presence in these equations increases rather than decreases the voltages expressed. The operation of the complete Fig. 10.11 circuit parallels that described for the two-amplifier example here. The amplifier having the largest input voltage will control e_{fb} and produce a high logic

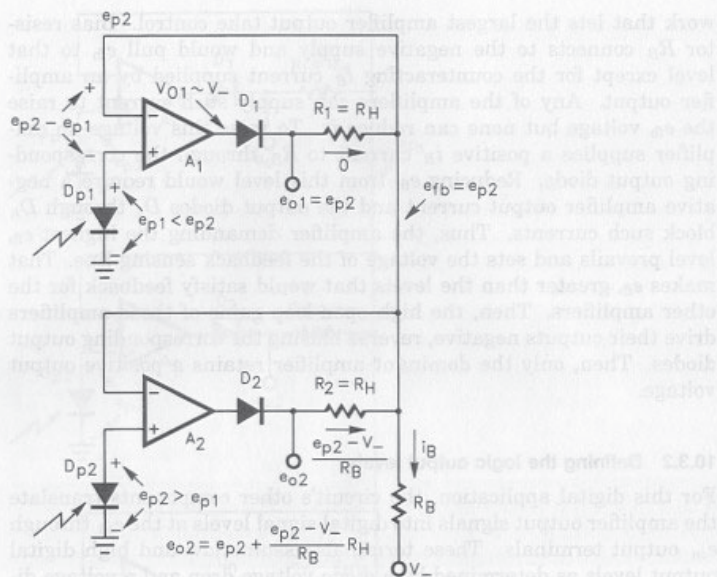


Figure 10.12 In a two-amplifier simplification of the Fig. 10.11 circuit, a larger diode signal e_{p2} overrides a smaller one e_{p1} to produce a high output state at e_{o2} and a low state at e_{o1} .

level at its corresponding e_o output terminal. All other amplifiers will produce low logic levels.

Diode matching and the amplifiers' offset and gain errors determine the accuracy of this digital position sensor. Together, they produce a net input error voltage $V_{i\epsilon}$. Diode responsivity mismatches make the light-to-voltage conversions differ for the diodes of an array, and for a given illumination condition, this produces a voltage mismatch error ΔV_p . Amplifier input offset voltage V_{OS} appears between the op amp inputs and adds directly to the diode error voltage. Similarly, amplifier gain error e_o/A_{OL} appears between an op amp's inputs, and the net input error voltage becomes

$$V_{i\epsilon} = \Delta V_p + V_{OS} + e_o/A_{OL}$$

where e_o is the output voltage of the op amp and not that of the maximum selector.

References

1. J. Graeme, "Use Op Amps to Design Optical Position-Sensing Circuitry," *EDN*, November 26, 1987, p. 229.
2. G. Tobey, J. Graeme, and L. Huelsman, *Operational Amplifiers; Design and Applications*, McGraw-Hill, New York, 1971.
3. S. Gage, *Optoelectronics/Fiber-Optics Applications Manual*, 2nd ed., McGraw-Hill, New York, 1977, p. 4.3.
4. F. Daghighian, "Optical Position Sensing with Duolateral Photoeffect Diodes," *Sensors*, November 1994, p. 31.
5. J. Graeme, "Generalized Op Amp Model Simplifies Analysis of Complex Feedback Schemes," *EDN*, April 15, 1993, p. 175.
6. J. Graeme, *Designing with Operational Amplifiers; Applications Alternatives*, McGraw-Hill, New York, 1977.

Glossary

avalanche photodiode A photodiode intended to be operated on the verge of avalanche breakdown so that impact ionization produces carrier multiplication and current gain.

bootstrap A circuit technique that removes signal voltage swing from a source by driving the source's common return with the signal appearing at the source's output terminal.

built-in voltage φ_B The potential barrier of a semiconductor junction as established by diffusion of carriers across the junction under zero bias.

common-mode input capacitance C_{icm} The effective capacitance between either input of a differential amplifier and common ground.

common-mode input resistance R_{icm} The effective resistance between either input of a differential amplifier and common ground.

common-mode rejection (CMR) The logarithmic form of common-mode rejection ratio (CMRR) as expressed by $CMR = 20 \log (CMRR)$.

common-mode rejection ratio (CMRR) The ratio of the differential gain of an amplifier to its common-mode gain.

common-mode voltage The average of the two voltages applied to differential amplifier inputs.

compensation resistor A resistor added in series with the noninverting input of an op amp circuit to develop an offset voltage component with the input current there that compensates for the offset component generated by the current of the noninverting input.

composite amplifier An op amp circuit enclosing two or more op amps within a common feedback loop.

current-to-voltage converter The op amp circuit configuration consisting of the op amp and a single, negative feedback resistor for converting an input current signal into an output voltage signal.

damping ratio ζ A figure-of-merit representation of the response damping characteristic of a second-order system.

dark current I_D The reverse saturation current of a photodiode or that current which flows through the diode under reverse bias and in the absence of illumination.

decoupling phase compensation A phase compensation technique that decouples an op amp from a capacitive load through an isolating resistor and a feedback bypass capacitor.

difference amplifier An op amp with a feedback configuration that results in an output voltage proportional to the difference of two input voltages.

differential input amplifier An amplifier having two inputs of opposite gain polarity with respect to the output.

differential input capacitance C_{id} The effective capacitance between the two inputs of a differential amplifier.

differential input resistance R_{id} The effective resistance between the two inputs of a differential amplifier.

duo-lateral photodiode A lateral photodiode having two current-dividing resistive surfaces for two-dimensional position detection.

electromagnetic noise coupling The parasitic signal coupling from a magnetic source through mutual inductances.

electrostatic noise coupling The parasitic signal coupling by an electric field through mutual capacitances.

feedback factor β The fraction of an output signal fed back to the input in a feedback system.

flux ϕ_e The radiant flux energy of a light source in watts.

gain error signal The differential signal voltage developed between the two inputs of an op amp and equal to the output signal voltage divided by the amplifier's open-loop gain.

I-to-V gain Current-to-voltage gain.

impact ionization The generation of free carriers through carrier collisions with semiconductor atoms.

input bias current I_B The dc biasing current drawn by both inputs of an op amp.

input capacitance See Common-mode input capacitance and Differential input capacitance.

input noise current i_{ni} The noise current drawn by both inputs of an op amp.

input noise voltage e_{ni} The input-referred equivalent noise voltage of an op amp that would reproduce the noise at the output produced by internal noise sources.

input offset current I_{OS} The difference between the two input bias currents of an op amp.

input offset voltage V_{OS} The dc voltage impressed between the inputs of an op amp that produces zero output voltage.

input resistance See Common-mode input resistance and Differential input resistance.

instrumentation amplifier A differential input amplifier with internal feedback committed for voltage gain.

integrator An op amp with a feedback configuration that results in an output signal proportional to the time integral of the input signal.

intercept frequency f_i That frequency at which an op amp circuit's $1/\beta$ response intercepts the op amp's A_{OL} response, marking the highest frequency for which the circuit's available gain supplies the feedback gain demand.

intrinsic layer A semiconductor layer containing no P or N doped atoms.

junction capacitance An equivalent capacitance that models the charge storage of a semiconductor junction.

lateral photodiode A position-sensing photodiode with a resistive surface path that divides the diode's output current into multiple portions according to the position of a light beam on the diode's surface.

loop gain $A\beta$ The excess gain available to supply increasing feedback gain demand.

maximum selector A device which selects the largest of a set of input signals and connects the selected signal to the circuit output.

noise See Input current noise and Input voltage noise.

noise density The spectral noise contained in a 1-Hz bandwidth at a given frequency.

noise gain peaking A noise gain phenomena common to photodiode amplifiers in which the gain supplied to the op amp's input noise voltage rises and peaks at high frequencies.

open-loop gain The ratio of an op amp's output signal magnitude to the signal magnitude appearing between the amplifier's inputs.

phase compensation Frequency response tailoring for the stability of a feedback system through the addition of response poles and zeroes that reduce high-frequency phase shift.

phase margin ϕ_m The margin separating the phase shift around a feedback loop from 360° at the frequency where loop gain equals unity.

photoconductive The current-output operating mode of a photodiode.

photodiode A semiconductor junction that converts the photon energy of light into an electrical signal.

photovoltaic The voltage-output operating mode of a photodiode.

PIN photodiode A photodiode formed with an intrinsic layer between the normal *P*- and *N* layers to expand the diode's depletion region and increase spectral bandwidth.

radio-frequency interference (RFI) The higher-frequency form of electromagnetic noise coupling.

rate of closure The difference in slopes of the $1/\beta$ and A_{OL} responses at their crossing as expressed in decibels.

responsivity r_ϕ The light-to-current gain of a photodiode as expressed by $r_\phi = i_p/\phi_e$, where ϕ_e is the radiant flux energy of the light in watts.

slewing rate S_r The maximum rate of change of output voltage.

tetra-lateral photodiode A lateral photodiode having two orthogonal sets of output electrodes on one current-dividing resistive surface for two-dimensional position detection.

unity-gain crossover f_c The frequency at which the open-loop gain of an op amp crosses unity.

virtual ground A groundlike characteristic of the feedback input of an op amp at which feedback absorbs injected current without developing a voltage.

voltage follower The short-circuit feedback connection of an op amp which results in an output signal that follows the signal at the amplifier's noninverting input.

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