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PHOTODIODE MONITORING WITH OP AMPS

With their low-input currents, FET input op amps are universally used in monitoring photodetectors, the most common of which are photodiodes. There are a variety of amplifier connections for this purpose and the choice is based on linearity, offset, noise and bandwidth considerations. These same factors influence the selection of the amplifier with newer devices offering very low-input currents, low noise and high speed.

Photodetectors are the bridge between a basic physical indicator and electronics resulting in the largest single usage of FET op amps. As a measure of physical conditions, light is secondary to temperature and pressure until the measurement is made remotely with no direct contact to the monitored object. Then, the signals of a CAT scanner, star-tracking instrument or electron microscope depend on light for the final link to signal processing. Photodiodes have made that link economical and expanded usage to detector arrays that employ more than 1000 light sensors. Focus then turns to accurate conversion of the photodiode output to a linearly related electrical signal. As always, this is a contest between speed and resolution with noise as a basic limiting element. Central to the contest is the seemingly simple current-to-voltage converter which displays surprising multidimensional constraints and suggests alternative configurations for many optimizations.

CURRENT-TO-VOLTAGE

The energy transmitted by light to a photodiode can be measured as either a voltage or current output. For a voltage response, the diode must be monitored from a high impedance that does not draw significant signal current. That condition is provided by Figure 1a. Here, the photodiode is in series with the input of an op amp where ideally zero current flows. That op amp has feedback set by R_1 and R_2 to establish amplification of the voltage diode just as if it was an offset voltage of the amplifier. While appealing to more common op amp thinking, this voltage mode is nonlinear. The response has a logarithmic relationship to the light energy received since the sensitivity of the diode varies with its voltage.

Constant voltage for a fixed sensitivity suggests current output instead and that response is linearly related to the incident light energy. A monitor of that current must have zero input impedance to respond with no voltage across the diode. Zero impedance is the role of an op amp virtual ground as high-amplifier loop gain removes voltage swing from the input. That is the key to the basic current-to-voltage converter connection of Figure 1b. It provides an input resistance of R_1/A where A is the open-loop gain of the op amp. Even though R_1 is generally very large, the resulting input resistance remains negligible in comparison to the output resistance of photodiodes.

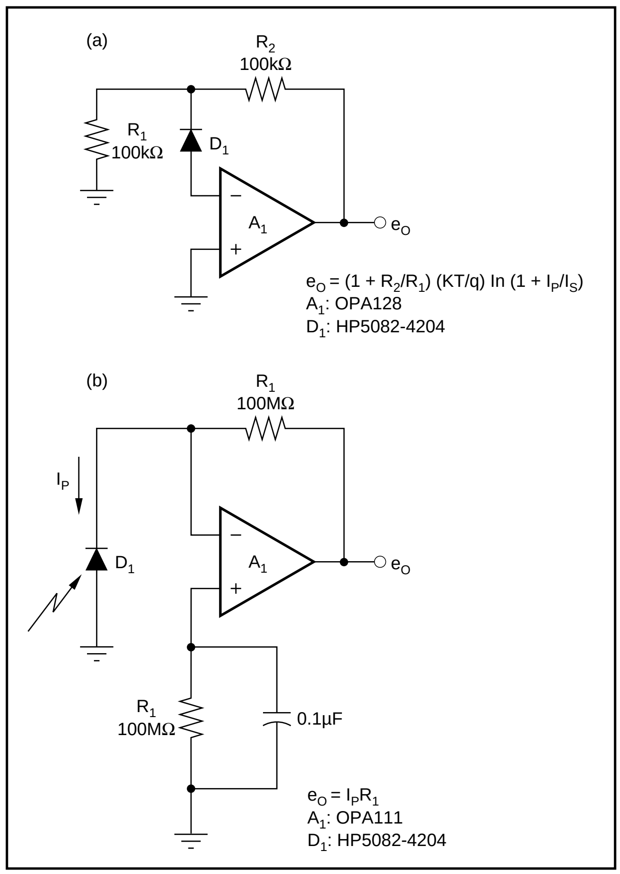

FIGURE 1a. Photodiode Output Can be Monitored as a Voltage; or, 1b, as a Current.

Diode current is not accepted by the input of the op amp as its presence stimulates the high amplifier gain to receive that current through the feedback resistor, R_1. To do so, the amplifier develops an output voltage equal to the diode current times the feedback resistance, R_1. For that current-to-voltage gain to be high, R_1 is made as large as other constraints will permit. At higher resistance levels, that resistor begins to develop significant thermal DC voltage drift due to the temperature coefficient of the amplifier input current. To compensate this error, an equal resistance R_2 is commonly connected in series with the op amp noninverting input, as shown, and capacitively bypassed to remove most of its noise. The remaining DC error is determined by the mismatches between the amplifier input currents and between the two resistors. A drawback of this error correction is the voltage drop it creates across the diode and the resulting diode leakage current. That leakage can override the correction achieved with R_2, as photodiodes typically have large junction areas for high sensitivity. Leakage current is proportional to that area which can become much larger than the op amp input currents.

Only zero diode voltage can eliminate this new error source but that is in conflict with control of a second attribute of large diode area. Large parasitic capacitance is also present creating often severe amplification of noise as will be described. To reduce that capacitance, a large reverse-bias voltage is sometimes impressed on the diode greatly complicating DC stability and making current noise from the photodiode an additional error factor. Larger diode area may actually degrade overall accuracy and higher photo sensitivity should first be sought through optical means such as a package with an integral molded lens. Monitor-circuit configurations that maintain zero diode voltage are also candidates in this optimization and are described with Figures 6, 7 and 9.

The value of the feedback resistor in a current-to-voltage converter largely determines noise and bandwidth as well as gain. Noise contributed directly by the resistor has a spectral density of $\sqrt{4KTR}$[1] and appears directly at the output of a current-to-voltage converter without amplification. Increasing the size of the resistor not only raises output noise by a square root relationship but also increases output signal by a direct proportionality. Signal-to-noise ratio, then, tends to increase by the square root of the resistance.

Noise from the op amp also influences the output with a surprising effect introduced by high feedback resistance and the diode capacitance. The amplifier noise sources are modeled in Figure 2a as an input noise current, i_n, and the input noise voltage, e_n. The current noise flows through the feedback resistor experiencing the same gain as the signal current. It is the shot noise of the input bias current, I_B, and has a noise density of $\sqrt{2qI_B}$[1]. Choice of an op amp having input currents in the picoamp range makes this noise component negligible for practical levels of feedback resistance. Input noise voltage of the amplifier would at first seem to be transferred with low gain to the output. That is true at DC where its gain $1 + R_1/R_D$ is kept small by the large diode resistance, R_D. Capacitance, C_D, of the diode alters the feedback at higher frequencies adding very significant gain to e_n. As both the capacitance and the feedback resistance are commonly large, the effect can begin at fairly low frequencies. Figure 2b illustrates the effect with an op amp gain magnitude curve plotted with the reciprocal of the feedback factor or the "noise gain." That gain curve first experiences a response zero due to C_D and begins a rise that is terminated only because of a second parasitic capacitance. Stray capacitance, C_S, shunts the feedback resistor resulting in a response pole leveling the gain at $1 + C_D/C_S$. For large area diodes C_D can be hundreds of picofarads causing the noise gain to peak in the hundreds as well. That gain continues to higher frequencies until rolled off by the op amp bandwidth limit. As feedback resistance increases, the pole and zero of this gain peaking move together to lower frequencies encompassing a greater spectrum with high gain.

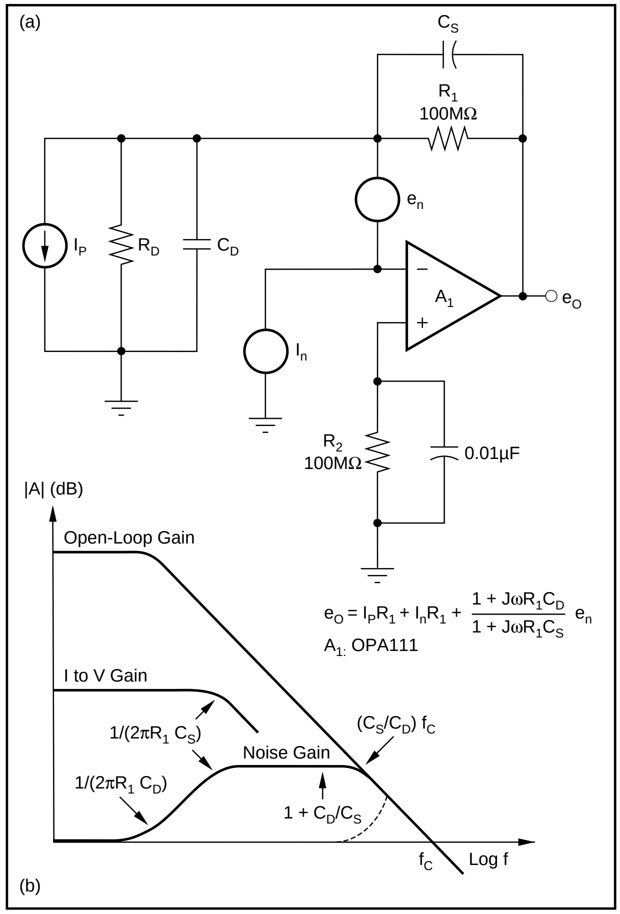

FIGURE 2a. Due to Diode Capacitance in the Feedback of the Basic Current-to-Voltage Converter, 2b, Op Amp Noise Receives Gain and Bandwidth Not Available to the Signal.

First signs of this gain peaking phenomena are familiar to anyone who has used high resistance op amp feedback in more general circuits. High output to input resistance with an op amp results in overshoot, response peaking, poor settling or even oscillation all due to the resistance interaction with amplifier input capacitance. Together the resistance and capacitance form another pole in the feedback loop resulting in the classic differentiator feedback response. Shown by the dashed line for more general op amp cases, the associated feedback factor reciprocal intercepts the amplifier open loop magnitude response with a 12dB/octave rate of closure corresponding to feedback phase shift approaching or equal to 180°. The common cure for this condition is a capacitor across the feedback resistor, which for the very high resistances of current-to-voltage converters, automatically results from stray capacitance. Such capacitance degenerates the added feedback pole to control phase shift in the feedback loop.

In understanding current-to-voltage converter noise performance it is important to note that the signal current and the noise voltage encounter entirely different frequency responses. The current-to-voltage gain is flat with frequency until the feedback impedance is rolled off by stray capacitance as shown. Gain received by the amplifier noise voltage, on the same graph, extends well beyond that roll-off and is high in that extended region. The majority of the op amp's bandwidth often serves only to amplify that noise error and not the signal. This is typically the dominant source of noise for higher feedback resistances.

Relative effects of the major noise sources of a current-to-voltage converter can be seen with the curves of Figure 3. Those curves show output noise for the basic current-to-voltage converter of Figure 1b including the effects of the noise gain represented in Figure 2b. Plotted are total output noises for three cases as a function of feedback resistance and each is the rms sum of the components produced by the feedback resistor and an op amp. Represented are three FET op amps having different performance specialties that cover the spectrum of photodiode applications with low noise, low-input bias current and high speed. While all three types have low-noise designs and low-input currents, the OPA111 offers the lowest noise in the FET op amp class at $6nV/\sqrt{Hz}$, and the OPA128 has the lowest input current at 0.075pA. Without neglecting performance in these categories, the OPA404 design pushes bandwidth to 6.4MHz. Noise due to the op amp is found by integrating the amplifier noise density spectral response over the noise gain response[2]. Also shown, by a dashed line, is the noise due to the resistor alone for the OPA111 and OPA2111 case. This resistor noise curve is different for the other op amps as each amplifier has a different bandwidth rolling off noise due to the resistor.

Different factors control the noise curves for different ranges of feedback resistance. At low resistance levels, the noise curves are largely flat with the op amp voltage noise the dominant contributor. That domination makes initial resistance increases have little effect except for the case of the very low-voltage noise of the OPA111/ OPA2111. In this region noise gain peaking has not yet been encountered so the output noise remains small. Between 10kΩ and 1MΩ, resistor noise is dominant and the curves track that error source as the dashed line shows for the OPA111/OPA2111. Here, the curves demonstrate the square root relationship with the resistance and differ only because of amplifier bandwidths. At still higher resistance, noise gain peaking takes effect returning the op amp noise to dominance and boosting the curves higher. That effect is first demonstrated by the increased slope of the OPA404 curve as that amplifier's wide bandwidth first encompasses the peaking. The noise curves level off when essentially the full amplifier bandwidth is encompassed by the gain peaking. Moving to yet higher resistance, resistor noise would return the curves to rising slopes, but resistor bandwidth is by then rolled off by stray capacitance. In this upper region, any increase in resistance is accompanied by a matching reduction in noise bandwidth so that the total resistor noise becomes a constant. Variables of diode and stray capacitances alter the point of onset of gain peaking errors, but the characteristic shape of the output noise curves remains the same for any case. Each will display ranges dominated by op amp noise, resistor noise and gain peaking effects.

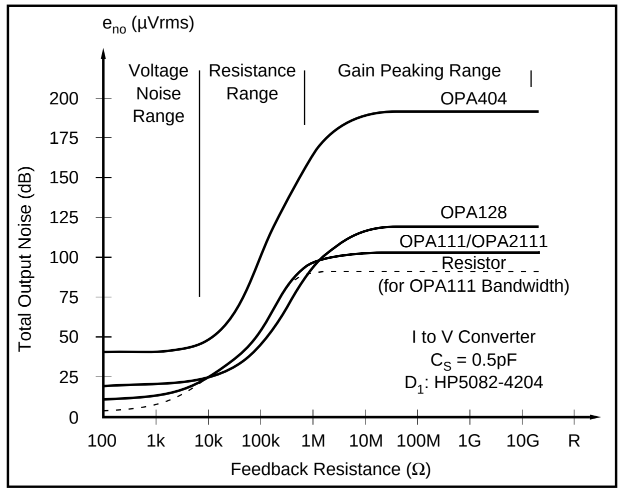

FIGURE 3. As the Feedback Resistance of a Current-to-Voltage Converter Increases, the Dominant Noise Source Changes from the Op Amp to the Resistor and Back to the Op Amp under Gain Peaking Conditions.

Comparing the curves shows that the OPA111/OPA2111 provide the lowest noise in two of the characteristic ranges. While the OPA128 shows a lower noise curve in the middle range, that is due to the amplifier's lower bandwidth and a bandwidth reduction technique to be described, removes that difference for the OPA111. Where the OPA128 excels is in very low DC error as its input currents are a mere 0.075pA which is 1/20th that of its low-noise contender. The third op amp, OPA404, produces higher total output noise overall, but that again is largely a bandwidth phenomenon. The 6.4MHz response of that amplifier accommodates noise over a much greater frequency range. While the noise curve for this amplifier is consistently higher than that of the OPA128, the OPA404 actually has lower noise density but it has six times the bandwidth. That 6.4MHz bandwidth is available to signals for feedback resistances up to 50kΩ and the amplifier still offers the best bandwidth for resistances up to 150kΩ. As the OPA404 is a quad op amp, its economy suggests consideration for use at even higher resistances along with bandwidth reduction that provides more competitive output noise.

Only a five dimensional graph could display the output noise, resistance, DC error, diode area and signal bandwidth considered in current-to-voltage converter design. Each specific application's requirements are evaluated separately with respect to these factors. To avoid suboptimizing a given design for one factor such as gain, the various effects of increasing feedback resistance are anticipated at each step. Choices such as large diode area are made considering the related capacitance and its effect on output noise and overall circuit sensitivity.

NOISE CONTROL

Gain peaking effects are the primary noise limitation with the commonly preferred high feedback resistances. To limit this effect, or to eliminate the gain rise entirely, additional capacitance is commonly added to bypass the feedback resistor. The capacitance level required can be very small for some values of R_1 and the relative significance of unpredictable stray capacitance make tuning desirable. Combined, these requirements are a challenge better resolved with a capacitor tee network as described in Figure 4a. It is capable of even subpicofarad tunable capacitance with little effect on stray capacitance in the tuning operation. The tee uses a capacitive divider formed with C_2 and C_3 to attenuate the signal applied to C_1 at the circuit input. With only a fraction of the output signal on C_1, it supplies far less shunting current to the input node as would a much smaller capacitor. Controlling the attenuation ratio is the tunable C_3, which is the largest of the capacitors, so its capacitance value is more readily available in tunable form. Since that capacitor is grounded, it has a shielding advantage to reduce stray capacitance influence while tuning.

Another option for practical feedback bypass exists with a resistor tee which is a commonly considered replacement for the high value feedback resistor. The latter is replaced in Figure 4b by elements of more reasonable value but introduces greater low frequency noise. Its operation is the dual of the capacitor tee above with R_2 and R_3 attenuating the signal to R_1 so that the latter appears as a much larger resistor to the input node. A similar opportunity for the DC error compensation resistor R_2 does not exist. DC error due to amplifier input current is no different with the tee so the large compensation resistor is still needed.

Stray capacitance across the feedback is somewhat reduced with the resistor tee by the added physical spacing of the feedback with three elements. Also, stray capacitance across each individual element has much less effect with their lower resistances. Sensitivity to other stray capacitance from the op amp output to its input has the same effect as before.

In the attenuation network of the feedback is the opportunity for intentional bypass with reasonable capacitor values. Bypassing the moderate resistance of R_2 removes the attenuation at higher frequencies leaving the net feedback resistance at the level of R_1. This operation differs from true feedback bypass in that impedance levels off, rather than continuing to fall with frequency, but the dramatic drop in

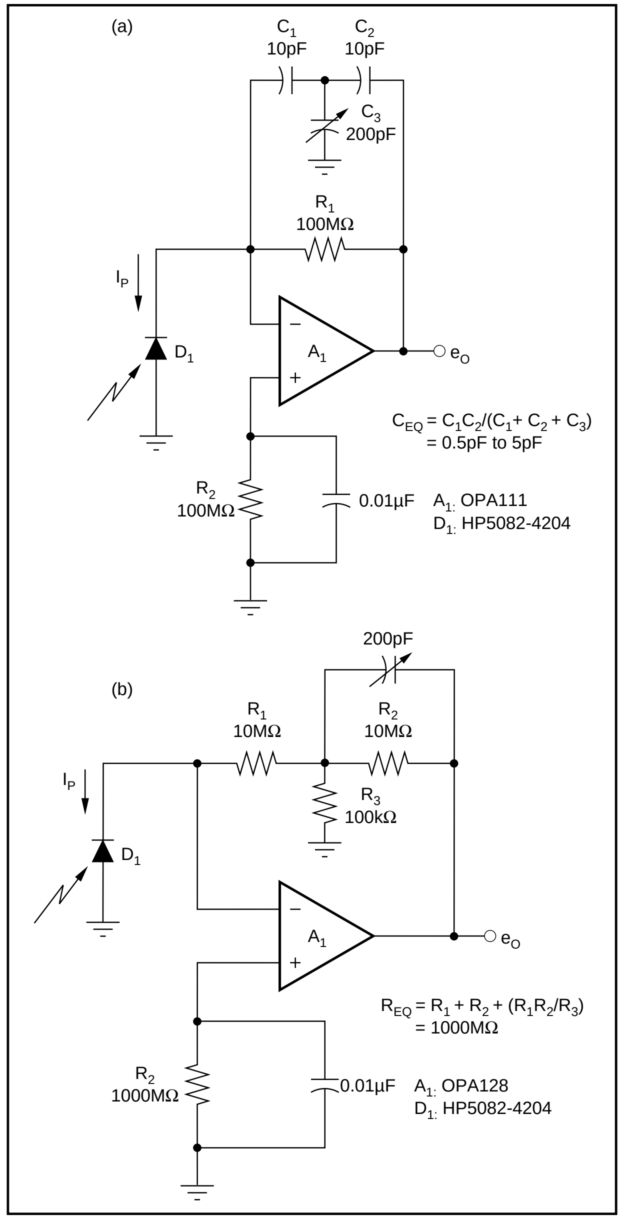

FIGURE 4a. Removal of Amplifier Gain Peaking Through Small Capacitive Bypass of Large Feedback Resistance is More Feasible with a Capacitor Tee; or, 4b, Bypass of One Element of a Feedback Resistor Tee.

equivalent resistance serves the circuit requirement. Another benefit offered by the resistor tee is more accurate DC error compensation.

Reduced high frequency noise with the tee element bypass is accompanied by an opposing increase at lower frequencies. Below the frequency of the bypass, noise gain is increased by the feedback attenuation of the tee network. That amplifies the noise and offset voltages of the op amp as well as the noise of resistor R_1 by a factor of $1 + R_2/R_3$. Countering the latter is the resistor's smaller value so that this effect is increased only by the square root of the new noise gain. Most important, however, is the bypass capacitor removal of high frequency gain as it eliminates the greatest portion of previous noise bandwidth. In the absence of other means to remove the high frequencies, the bypassed resistor tee provides lower total output noise for the higher ranges of feedback resistance.

Adding feedback capacitance is an effective means of reducing noise gain but it also decreases signal bandwidth by the same factor. That bandwidth is already low with high feedback resistance and the end result can be a response of a kilohertz or less. A more desirable solution to the noise problem is to limit amplifier bandwidth to a point just above the unavoidable signal bandwidth limit. Then, the high frequency gain which only amplifies noise is removed. Op amps with provision for external phase compensation offer this option, but those available lack the low-input currents and low-voltage noise needed for photodiode monitoring.

To achieve this bandwidth limiting with better suited op amps, a composite amplifier uses two op amps with the added one for phase compensation control as in Figure 5a. Note the reversal of the inverting and noninverting inputs of A_1 needed to retain a single phase inversion with two amplifiers in series. With the composite structure, internal feedback controls the frequency response of the gain added by A_2. At DC, that feedback is blocked by C_1 and overall open-loop gain is the product of those of the two amplifiers or 225dB for those shown. That gain is rolled off by the open-loop pole of A_1 and by the integrator response established for A_2 by C_1 and R_3. As this is a two pole roll-off, it must be reduced before intercepting the noise gain curve to establish frequency stability. A response zero does this due to the inclusion of R_4. Above the frequency of that zero, R_4 also replaces the integrator response with that of an inverting amplifier having a gain of $-R_4/R_3$. Making that gain less than unity drops the net gain magnitude curve below that of a single amplifier at high frequencies. Graphically, the noise gain response of Figure 5b is moved back in frequency much as if the op amp bandwidth had been reduced.

Eliminated is the shaded area of noise gain, which visually may not appear dramatic, but that is because of the logarithmic-frequency scale. Actually, the associated noise reduction is large because most of the amplifier's bandwidth is represented in this upper end of the logarithmic-response curve. Moving the unity gain crossover of the noise gain from 2MHz to 200kHz, as shown, drops the output noise due to A_1 by about a factor of three. To achieve the same result with feedback bypass, the signal bandwidth would have been reduced a factor of ten. That bandwidth is unaffected with the Figure 5a approach. No noise, or offset, is added by A_2 as this amplifier is preceded by the high gain of A_1. With the exceptionally low noise of the OPA111 input amplifier, this improvement reduces noise to the fundamental limitation imposed by that of the feedback resistor. This condition is retained for all practical levels of high feedback resistance. For the second amplifier, the wideband OPA404 is shown to continue its attenuating amplifier action well beyond the unity gain crossover of A_1. This avoids a second gain peak that could cause oscillation. Signal bandwidth of the current-to-voltage conversion is essentially unaffected as R_1 has not been influenced .

Where the Figure 5 technique is most useful is with lower level signals that have greater sensitivity to noise. In higher level applications that circuit can encounter a voltage swing limitation but another use of the second amplifier offers similar noise improvement. The swing limitation results from the maximum output voltage limit of A_1 and its attenuation by A_2. If the output of A_1 has a peak swing of 12V and A_2 has the gain of –1/10 illustrated, the final output is limited to a 1.2V peak swing. For lower-level signals this will be acceptable as the maximum practical level of feedback resistance already limits output swing.

Higher-level signals are not as sensitive to noise and better tolerate a more straight forward approach to filtering. An active filter following the conventional current-to-voltage converter also removes the high frequency noise. Setting filter poles at the frequency of the signal bandwidth results

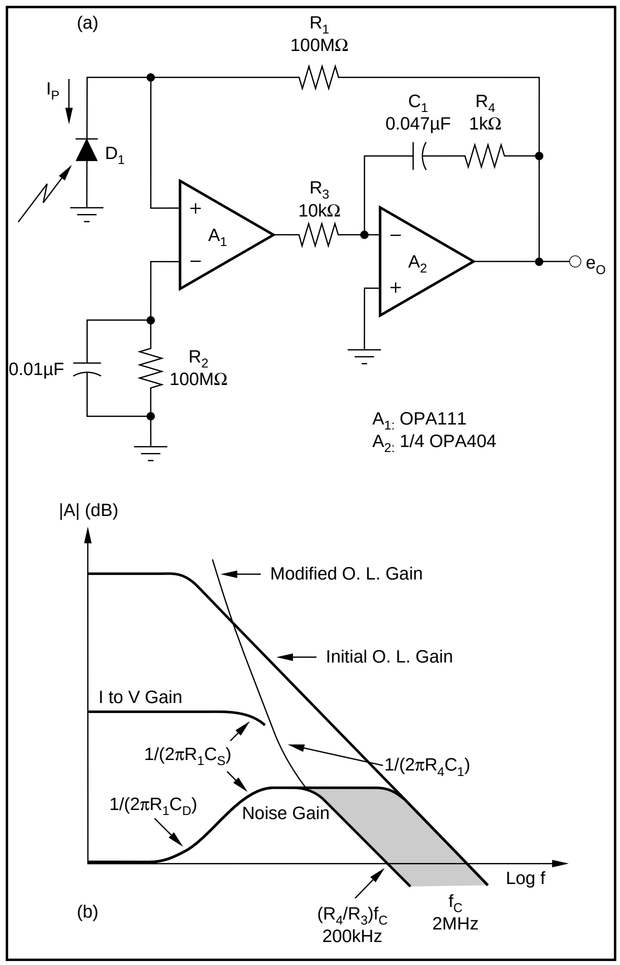

FIGURE 5a. Noise Reduction Results with a Composite Amplifier that, 5b, Restricts Noise Bandwidth Without Reducing that of the Signal.

in a system bandwidth that does not extend beyond that of useful information. Such a filter is not enclosed in a feedback loop with the converter so the input noise and offset voltage of the second amplifier are added to the signal.

BANDWIDTH

Signal bandwidth requirements are an integral part of the current-to-voltage converter noise considerations for two reasons. Total output noise increases in proportion to the square root of system bandwidth simply because a broader noise spectrum is encompassed. Added is conflict between optimum signal-to-noise ratio and signal bandwidth. That optimum occurs for very high gain but high gain current-to-voltage converters are bandwidth limited far below the roll-off of the op amp. To the signal current, the amplifier feedback factor is unity which would normally make the full amplifier unity gain bandwidth available. Yet the very high-feedback resistances that produce the desired gain are shunted by stray capacitances at much lower frequency. Just 0.5pF stray capacitance around a 100MΩ feedback resistor pulls signal bandwidth from megahertz level unity gain crossovers down to 3.2kHz. To minimize the stray shunting, low capacitance resistors and assembly precautions are used. Mounting the feedback resistor on standoffs reduces capacitive coupling with printed circuit boards and such standoffs are normally Teflon™ insulated to reduce leakage currents. That mounting must be rigid to avoid introduction of noise through the microphonic effects of mechanical stress from vibrations.

There is an ultimate limit to the effects of such measures as capacitive coupling through the air around the resistor body always remains. Bandwidth beyond that imposed by such residual limits requires lower feedback resistance and accompanying lower converter gain. To restore gain, several options are available with a first shown in Figure 6a. A second amplifier with voltage gain is simply added following the current-to-voltage converter to retain the net input to output transimpedance for $R_T = A_V R_1$. Then, the high-value resistance is reduced by a factor equal to the voltage gain for a bandwidth increase by as much as the same factor.

While an obvious alternative, its overall effect on bandwidth and noise are not so immediate. Bounding the upper end of the bandwidth increase is the response limitation of the second amplifier. The bandwidth of the two op amp circuit for a net transimpedance of 100MΩ is plotted in Figure 6b as a function of the voltage gain involved in the overall conversion. Bandwidth initially increases linearly with the voltage gain as the reduction in R_1 diminishes the roll-off effect of stray capacitance. However, the added demands of the voltage gain on A_2 eventually make that amplifier's bandwidth the controlling factor. For a given set of conditions there is an optimum gain. A_V produces the peak bandwidths shown for the three example amplifiers. That peak occurs when the amplifier closed loop bandwidth equals the stray limited bandwidth of R_1. Variables affecting this peak are the net transimpedance, R_T, and the second op amp unity gain bandwidth, f_C. Interrelating the controlling factors at the optimum bandwidth point is the expression defining the choice of R_1:

$$R_1 = \sqrt{R_\tau / 2\pi C_S f_C}$$

Bandwidth is extended to 100kHz from the original 3kHz using the wideband OPA404 for the second amplifier. That wideband op amp offers the best frequency response in Figure 6 and, although its total output noise result is greater, that is again largely due to the greater available bandwidth. If even greater bandwidth is required, either a faster op amp, with typically poorer noise performance, or lower transimpedance are the choices. Less bandwidth demands are encountered by A_1 with its unity feedback factor so an FET amplifier focused on low noise is used there like the OPA111 shown.

The price paid for improved bandwidth through voltage gain is increased output noise from that gain as well as from the presence of the added amplifier. While the lower value of R_1 does reduce its noise density, that effect is counteracted by the increase in bandwidth for a net zero change is resistor noise. That noise is now amplified by the voltage gain of the

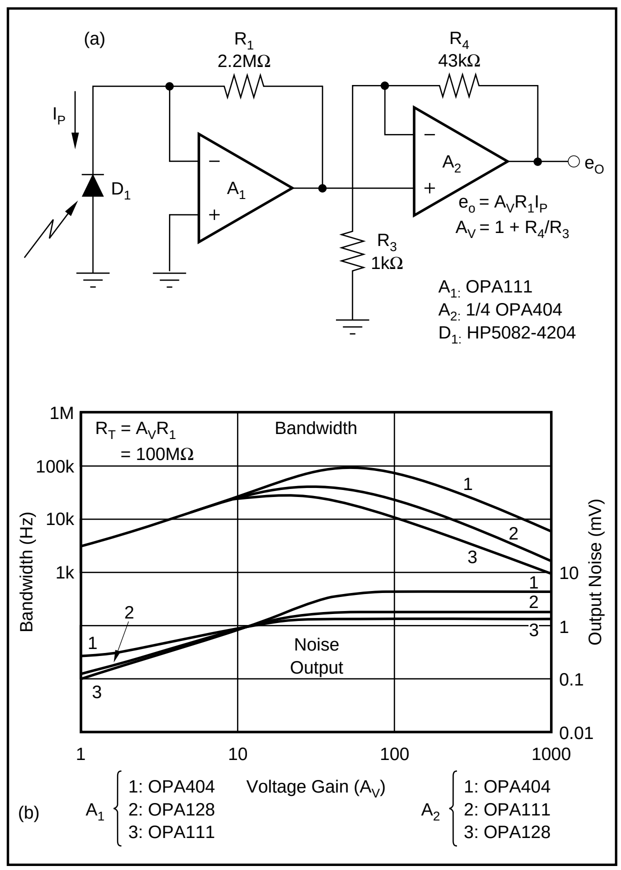

FIGURE 6. For Greater Bandwidth and the Same Net Transimpedance, (a) Voltage Gain is Added Providing (b) Bandwidth That Increases Faster Than Noise.

second amplifier causing an associated increase in output noise proportional to the voltage gain. Added to that is the noise from the op amps with the net result also shown in Figure 6b. Those noise curves are continuations of the ones presented in Figure 3 with the transition beginning at the 100MΩ level for the present example. In the lower gain ranges from one to ten, the noise is first determined largely by the op amps and their gain peaking but those effects give way to resistor noise dominance before the end of this range. Also in that range, the associated signal bandwidth plotted in Figure 6b is controlled by stray capacitance and shows a linear increase with increasing gain due to the corresponding decrease in resistance. Above the gain of ten and before 100, bandwidth begins to drop due to the encounter of A_2 limits. Simultaneous with this drop is a flattening of the output noise curve. Roll-off of the amplifier bandwidth and the simultaneous resistance drop nullify the effect of increasing voltage gain leaving output noise a constant. In the voltage gain range from 100 to 1000, these trends continue and degrade optimum performance since bandwidth is lost while noise remains constant.

While it is accepted that noise degrades with the voltage gain replacement of resistance, the overall circuit figure of merit gains. Including bandwidth in that measure shows that its improvement more than offsets the drop in signal-to-noise ratio. Mentioned before was the fact that the simple current-to-voltage converter suffers from greater bandwidth for the amplifier voltage noise than for the signal current. That discrepancy is removed with Figure 6 as the voltage gain increases and A_2 begins to filter out the higher frequencies. Evidence of this is in the noise curves that increase more gradually than the bandwidth curves—Figure 6b—up to the optimum bandwidth point. At this optimum point, no bandwidth is afforded to noise that is not also available to signal. In effect, A_2 now also serves as the output active filter discussed earlier. While each of these curves is drawn for a specific 100MΩ transimpedance and the amplifiers and photodiode specified, similar optimums are considered for any design case.

For some of the more common photodiode applications, a significant drawback of the above circuit is the need for two op amps per photodetector. Often hundreds of detectors are employed in a large arrays. As a compromise, one op amp can be made to provide the same transimpedance, still without the very large resistors, if some bandwidth and noise degradations can be accepted. A single op amp can both perform the current-to-voltage conversion and provide the subsequent voltage gain. With traditional techniques the task would be performed as in Figure 7a using R_2 for the conversion and R_3 and R_4 to set voltage gain. Current from D_1 flows in R_2 resulting in a signal voltage at the input of a noninverting amplifier. However, that signal voltage is also across the photodiode and this condition produces a nonlinear response as described before.

Instead, the diode is connected directly between the op amp inputs where zero diode voltage is maintained. Shown in Figure 7b, the resistors perform the same functions as in the last circuit but a linear response results. Current from the photodiode still flows in R_2 developing the same signal voltage. That current also flows into the feedback network but has little effect with the low resistances there. For the resistor values shown, an equivalent transimpedance of 100MΩ results—just as with the two op amp example, but bandwidth improvement is less. At 20kHz, it is increased a factor of seven rather than a factor equal to the voltage gain as in Figure 6a. A new bandwidth limitation accounts for the difference and occurs due to the new placement of the high-value resistance. That resistor is now shunted by the common-mode input capacitance of the op amp instead of just the smaller stray capacitance. To maximize bandwidth, this new shunting effect is made to coincide with the amplifier roll-off through choice of R_2 and the voltage gain. A second benefit from this choice is that resistor noise beyond the signal bandwidth encounters a two pole roll-off.

Final output noise from the resistor has the expected increase over the basic circuit by the square root of the voltage gain. Added to that would have been a small component due to the op amp as the normal source of gain peaking is

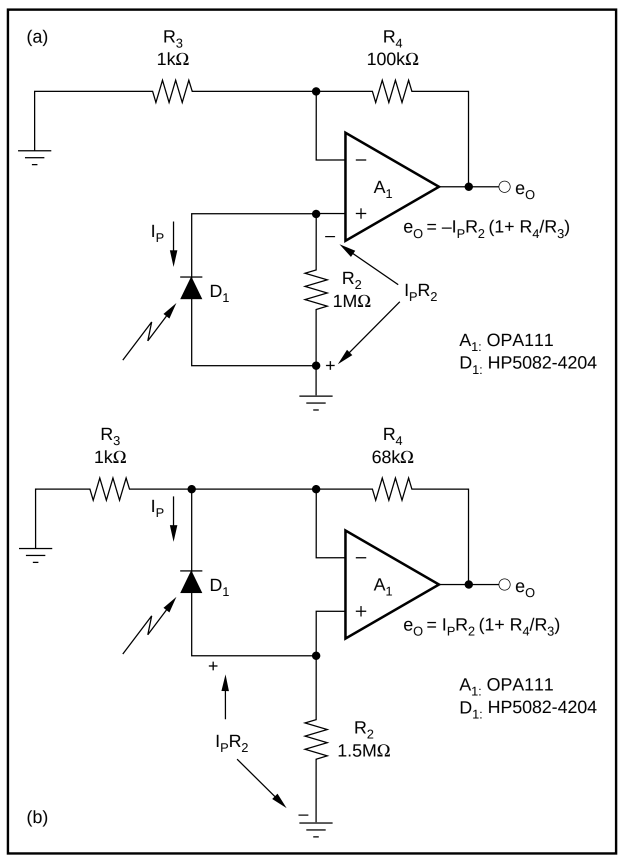

FIGURE 7. Combining Current-to-Voltage Conversion and Voltage Gain Using One Op Amp, (a) Impresses Unwanted Voltage on the Diode that (b) is Removed by Connecting the Diode Between the Op Amp Inputs.

removed. However, a new source is included in the circuit of Figure 7b, again due to the diode capacitance, as modeled in Figure 8a. Amplifier voltage noise, e_n, is impressed directly across that capacitance developing a noise current that is supplied to R_2. That creates a noise voltage at the input of the noninverting amplifier which is a multiple of e_n. The capacitive feedback network of C_D and C_{ICM} produces a noise gain that peaks at $1 + C_D/C_{ICM}$ and which exists in addition to the normal voltage gain of the noninverting amplifier.

Effects on frequency response are plotted in Figure 8b and they again produce a high-frequency peak in the noise gain. Its incidence is at a much higher frequency than with the basic current-to-voltage converter because of the lower resistance involved and it is truncated earlier by the op amp roll-off. For the low capacitance diode used in both example circuits, it now encompasses little area in the response plot corresponding to less noise effect. Larger diodes do not escape the effect, however, as represented by the dashed line for a capacitance around 200pF. Even still, the spectrum covered by the peaking is not the high end of the op amp bandwidth as it was for the basic circuit. Hence, op amp noise does not become the overriding source.

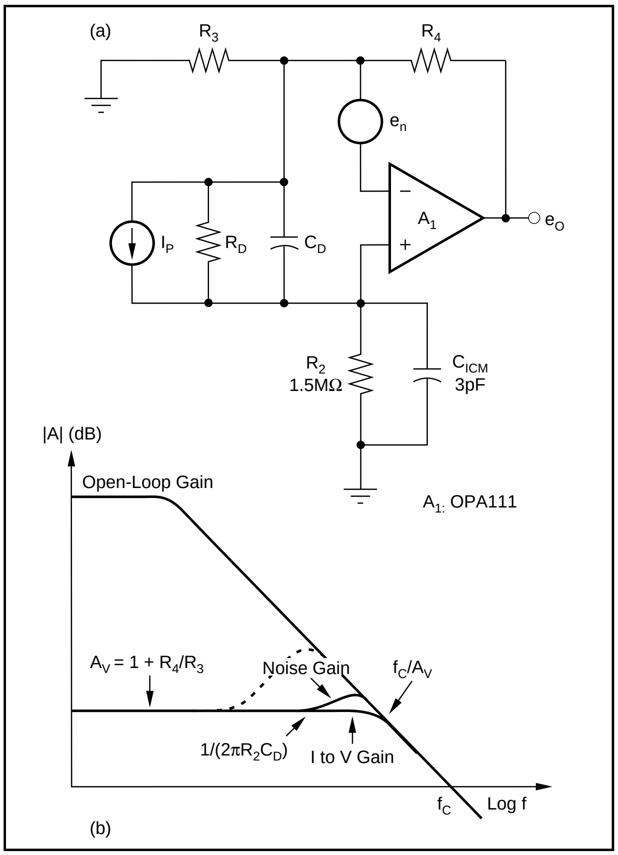

FIGURE 8. Photodiode Capacitance (a) Adds a Positive Feedback Path to Figure 7b for (b) a New but Lesser Source of Gain Peaking.

INTERFERENCE

Once diminishing returns impose a limit on reduction of the noise due to the circuit itself, consideration must be given to external noise sources. With its very high resistance, a current-to-voltage converter is extremely sensitive to noise coupling from electrostatic, magnetic and radio frequency sources. Those sources require attention to shielding, grounding, and component physical location[3], or they could otherwise become the dominant noise contributors. In each case, physical separation of the noise source from the sensitive circuitry is the most important step, but this becomes a compromise warranting other measures as well.

Electrostatic coupling, such as from the power line, supplies noise signals through the mutual capacitances that exist between any two objects. Voltage differences between the objects are impressed on those capacitances and any voltage variation couples a noise current from one to the other. To avoid that error signal, electrostatic shielding is used to intercept the coupled current and shunt it to ground. In this case, ground must be earth ground as that is the common reference for the separate objects. Such shields, however, create parasitic capacitances between the components shielded and the shields must also be returned to the signal common to avoid that coupling. Then shield carried capacitive currents from the output of a current-to-voltage converter are also shunted to ground and represent no bandwidth restriction to the feedback resistor. Even still, the shield produces a capacitance from the converter input to ground, possibly adding to gain peaking and its effect on total output noise.

As electrostatic coupling is most often of power line frequency and common to all points, it is a natural candidate for removal through the common-mode rejection of an op amp. At the line frequency, op amp CMR is very high but it is not utilized by the conventional current-to-voltage converter. This is a result of single-ended rather than differential input configurations, but that can be altered for improved noise rejection and DC error benefits as well. Op amp CMR is not a total replacement for shielding as electrostatic coupling will not perfectly common-mode to amplifier inputs. As a second defense, that rejection capability is most useful in removing the residual coupling that passes through shield imperfections.

The differential input capability of an op amp fits exactly with the signal from a photodiode. Since the diode signal is a current, it is available at both terminals of that sensor and can drive both amplifier inputs as in Figure 9a. Here, the diode current is no longer returned directly to common, but drives the amplifier noninverting input in that path. That creates a second signal voltage to double the circuit gain when $R_2 = R_1$ for, compensation. For a given gain level, the resistor value need be only one-half the normal for a similar reduction in error sensitivity to amplifier input currents. This also removes DC voltage from the diode as it is now directly across the inputs of an op amp. With the voltage between those inputs being essentially zero, photodiode leakage current is avoided.

Aside from these benefits is the added improvement in the common-mode rejection of coupled noise. Electrostatic coupling to this current-to-voltage converter is modeled in Figure 9b along with the converter's parasitic capacitances. Zero signal is assumed there to illustrate only the electrostatic coupling effects. The electrostatic noise source, e_e, couples error currents, i_e, through mutual capacitances, C_M, to the circuit's two inputs. It might seem that the coupling effects would be different to the two points because feedback makes the R_1 input node a virtual zero impedance and the other node is high impedance. Yet, the noise coupling is via currents through capacitances that only depend on voltage signals on the capacitances. Both input nodes have the same voltage due to amplifier feedback, and thus receive the same level of noise current i_e. Those equal currents develop canceling e_{ne} noise voltage effects on the two circuit resistors for a zero final output signal.

Accuracy of the error cancellation is determined by three matching conditions involving the mutual capacitances, the resistors and the parasitic capacitances shunting them. Matched mutual capacitances are best assured by locating the resistors equidistant from any significant noise source not effectively blocked by a shield. Equal resistance values assure accurate cancellation of error signals until frequencies are reached where capacitive shunting imbalances net impedances. Shunting R_1 will be only about 0.5pF of stray capacitance but across R_2 is the much larger common-mode input capacitance of the op amp. For the 3pF of the OPA111 and the 50MΩ resistance shown, a pole occurs at about 1kHz, leaving the impedances of interest unbalanced. This shunting by C_{ICM} also imposes a signal bandwidth limitation at a lower frequency than normally encountered. The bandwidth of R_2 is rolled off earlier than that of R_1 creating a response with two plateaus separated by a factor of two in gain.

For the most common electrostatic coupling at power line frequency, the above capacitive shunting has little effect. To better reject higher frequencies, capacitance can be added around R_1 to restore impedance matching, or signal swing on the common-mode input capacitance can be avoided. The latter option offers a more accurate solution and avoids the bandwidth limitation of C_{ICM} as well by using a second differential connection. Shown in Figure 10, the photodiode is connected between the inputs of two current-to-voltage converters whose outputs drive an INA105 difference amplifier. Again the diode current flows in two equal resistances that will receive equal electrostatic noise coupling. The diode current creates a differential output on the resistances, but the noise coupling generates a common-mode signal. Supplied to the INA105, those signals are separated with the diode signal passed to the output and the noise rejected.

Retained with the new differential input circuit are the 2:1 lower individual resistance and a zero diode voltage. The latter is assured by the grounded noninverting inputs of both current-to-voltage converters which establishes zero voltage on both diode terminals. These connections also avoid signal swing on common-mode input capacitances for improved bandwidth in electrostatic suppression and signal gain. Note that those noninverting inputs are not connected through high resistances for input current error correction. That is not necessary, as the input currents of A_1 and A_2 produce matching voltages at their amplifier outputs. Those voltages are a common-mode signal to the input of the INA105, so they too are rejected.

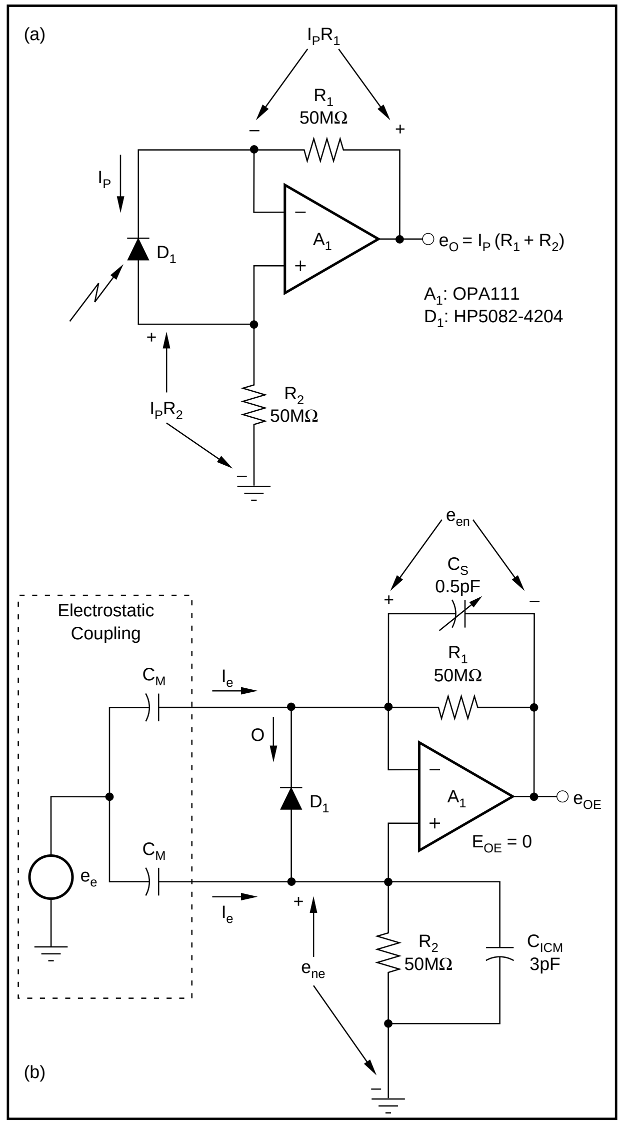

FIGURE 9. Exploiting the CMR Capabilities of the Op Amp, (a) the Differential Inputs are Driven Giving (b) Rejection of Electrostatic Coupling.

Another function available with the differential structure of Figure 10, is difference monitoring of two photodiodes. Instead of D_1, the two diodes shown in dashed lines are connected separately to the two input current-to-voltage converters. Their currents produce independent voltages at the outputs of A_1 and A_2 where they are processed by the difference amplifier to remove any common-mode portion. Left is an output proportional to the difference between the two input photocurrents as a measure of relative light intensity. A relative intensity measure is the type of signal used in position sensing or optical tracking control to direct feedback correction.

Magnetic coupling of noise can be more difficult to eliminate than the electrostatic, but its effects are also reduced by the differential input connections. Coupling is through mutual inductances in this ease, so minimum sensitive loop area is key to its control, along with shielding and maximum separation of source and receiver. Its effects are not removed by the electrostatic shield, so the first step is control of the source itself.[3] Power transformers that cannot be placed at a distance are internally shielded to largely terminate their magnetic fields at the transformer boundaries. Remaining magnetic coupling is addressed through physical and circuit configurations. High value resistors used in photodiode monitoring are sensitive to this coupling and connections must be kept short between those resistors and high impedance op amp inputs. Coupling effects that remain are made common-mode to be rejected by the op amp through loop size and distance matching. In Figure 9 and Figure 10, the high resistance is divided into two equal elements that are then physically mounted with the same orientation to and spacing from magnetic coupling sources. Noise coupled to the two resistors then causes equal signals that have canceling effects at the circuit output.

With the third class of noise coupling, radio frequency interference, less can be removed by the amplifiers so shielding and filtering are the best defenses. Sources of RFI may be close to the photodiode monitor because of digital circuitry that is most likely co-resident in the system. Due to the high frequencies involved, op amps have little gain or common-mode rejection remaining for rejection of such signals. Because of this same amplifier limitation, and the basic voltage-to-current converter bandwidth restriction, desired signals will not exist in the radio frequency range. Filtering can then be used to largely remove the unwanted signal if applied in front of the op amp. Later filtering is less effective as the op amp can act like an RF detector separating a lower frequency envelope from a carrier.[4] Further reduction of that noise is achieved with an RF shield and a ground plane layer in printed circuit boards.

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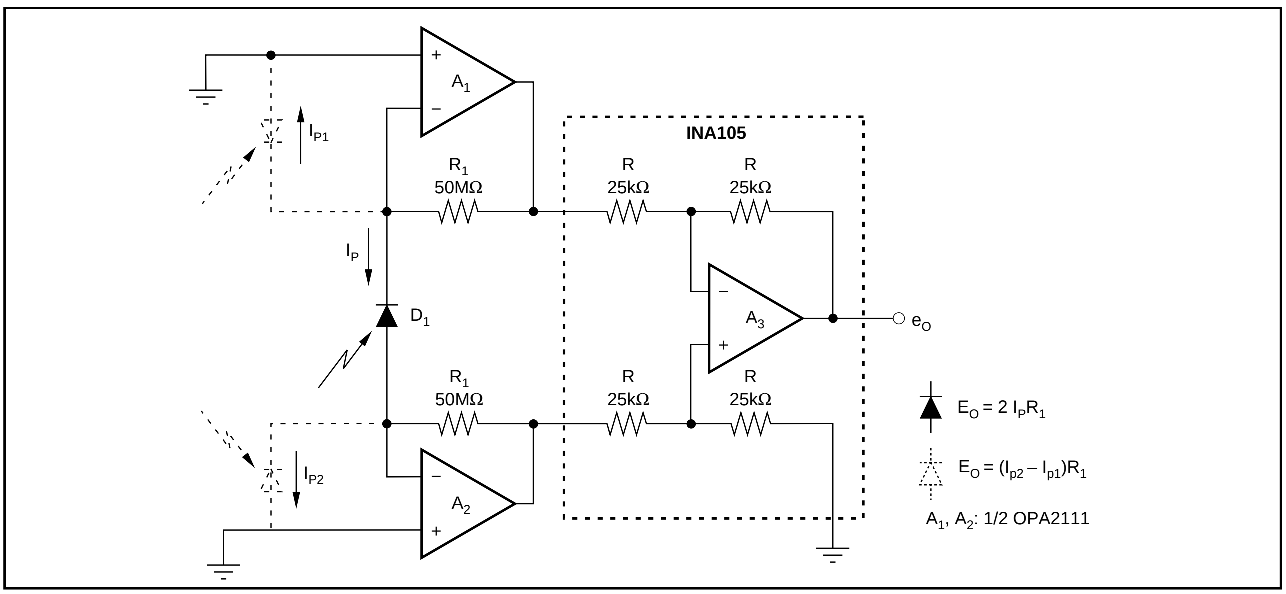

FIGURE 10. Differential Inputs with Wider Band CMR and Gain Result with Virtual Grounds Across Amplifier Common-Mode Input Capacitances.