Control frequency response and noise in broadband, photodetector, transimpedance amplifier

Michael Steffes - July 04, 1996

Converting the current output of a wideband photodetector to a voltage, minimizing noise, and achieving the desired frequency response can tax the patience of even an experienced designer. Careful selection and correct application of an op amp can often provide the best and most cost-effective solution.

Choosing and properly applying a transimpedance amplifier for broadband photodetector applications involves understanding trade-offs. Manufacturers of components for wideband transimpedance applications often spec products only for selected detector diodes. These components can be prohibitively expensive for all but the narrowest applications. After selecting op amps fast enough to achieve the desired bandwidth, the key becomes balancing each op amp’s voltage and current noise to get the lowest broadband integrated noise.

Even with the right op amp, incorrect frequency response compensation can cause peaking in the response, leading to poor signal performance and increased output noise. Common approaches to setting the compensation capacitor for op-amp transimpedance stages ignore the second-order nature of the closed-loop response. Given the detector capacitance, desired transimpedance gain, desired bandwidth, and op-amp gain-bandwidth product (GBP), you can develop a simple solution for the compensation capacitor across feedback resistor $R_F$ to achieve a desired second-order, lowpass response (Figure 1). Finally, determine the total integrated noise equation and use it to select the op amp that yields the lowest noise.

Transimpedance op-amp configuration

Low-noise, voltage-feedback op amps can provide even better bandwidth and lower noise than can dedicated trans-impedance amplifiers. Current-feedback amplifiers can operate as transimpedance stages, but, because the feedback resistor determines these amplifiers' bandwidth, it is difficult to separately set the gain and the bandwidth. For low-signal-level applications, voltage feedback offers superior noise performance and bandwidth control.

Your first hurdle in using an op amp for transimpedance is setting the compensation to limit peaking
in the frequency response. If the closed-loop transimpedance stage peaks significantly in the frequency domain, the result is ringing in the pulse response and higher noise than necessary. The most commonly recommended setting for this compensation typically gives about 1.25-dB frequency-response peaking.

**Figure 1** shows the basic transimpedance configuration for an op amp. You must select an op amp and then set $C_F$ to control the frequency response, given a detector diode and a desired transimpedance gain, $R_F$. The key selection criteria for the op amp include the GBP, the differential and common-mode input capacitance, the inverting input-current noise, and the noninverting input-voltage noise.

The design proceeds in two steps. First, from the detector's capacitance at the intended reverse bias, $V_B$, and from the desired transimpedance gain and signal bandwidth, you compute a minimum GBP for the amplifier. To see how to do this step, first develop a solution for $C_F$, the main factor controlling peaking in the frequency response; the amplifier's GBP, along with the peaking, sets the transimpedance bandwidth.

The second step is to compare the total integrated noise among various amplifiers that offer sufficient GBP. They differ in their input parasitic capacitance and noise terms. For low detector capacitances, you need to iterate for a best solution, because the amplifier's input parasitic capacitance can dominate bandwidth and noise performance. As the desired signal bandwidth increases, bipolar input stages can offer lower total noise than do FET input stages, due to the bipolar stages' lower input-voltage noise.

**Controlling the frequency response**

You can derive the transfer function for the transimpedance op-amp configuration using the analysis circuit of **Figure 2**, which shows the op-amp parasitic input capacitances. Because $R_D >> R_F$ for any high-speed application, ignore the real part of the diode's impedance. The key op-amp characteristics are common-mode input capacitance, $C_{CM}$; differential input capacitance, $C_{DIFF}$; and open-loop gain, $A(s)$, where (s) indicates the Laplace domain description of the frequency-dependent open-loop gain of the op amp. All three input capacitors appear in parallel, and you can treat them as a single capacitance to ground on the inverting op-amp input. For very low-capacitance diodes, these op-amp terms can dominate, thus limiting achievable bandwidth and noise. With all other things equal, a lower input capacitance for the amplifier provides better performance than a higher capacitance value.

Due to the diode's capacitance, the op amp in **Figure 2** behaves as if it is in a differentiator configuration. Without $C_F$, this circuit often oscillates, even when using unity-gain-stable op amps. From a Bode-analysis perspective (**Figure 3**), you select $C_F$ to limit the noninverting (or noise) gain to a finite level before its intersection with the op amp's open-loop gain curve. At high frequencies, the op amp sees a noninverting signal gain set by $1+(C_S/C_F)$. This gain determines the op-amp stability. A related issue is that the gain for the op amp's noninverting input-noise voltage increases with frequency, as the noise-gain part of **Figure 3** shows. This differentiated input-noise voltage often dominates the total output noise for wideband photodiode transimpedance amplifiers.

The transfer function from the input current to output voltage, $[V_O/I_D]$, is:
**Equation 1** is written in standard Bode analysis form, where the op amp's frequency-dependent gain shows up only in the denominator. If \( A(s) \) were very large and had infinite bandwidth, \( Z_F \) would simply establish the transfer function: a dc gain set by \( R_F \) and a single dominant pole set by \( 1/R_F C_F \). In any real op amp, however, \( A(s) \) has a high dc gain and a low-frequency dominant pole that rolls off the gain. For Bode analysis, you compare this response to the frequency response of the noninverting signal gain, \( 1+sC_S Z_F \). When the noise gain and open-loop response become equal in magnitude, the loop gain decreases to 1, and the closed-loop bandwidth typically starts to roll off. Substituting for \( Z_F \) and rewriting the noise-gain portion of the loop gain in pole/zero format results in **Equation 2**, which shows the zero and pole that the noise-gain plot in **Figure 3** illustrates.

\[ (2) \]

Because the numerator for the transfer function from \( V_O \) to \( I_D \) of **Equation 1** includes a pole due to \( Z_F \), you can't directly predict the closed-loop response from just the Bode analysis of **Figure 3**. However, you can make some important observations:

First, placing the pole for the noise gain at the intersection with the open-loop roll-off appears to yield a low phase margin (**Figure 3**). The phase shift due to the open-loop pole at \( \psi_A \), minus the phase shift due to the zero in the noise gain, gives 180° phase shift up to the pole for the noise gain. The pole in the feedback network decreases the phase around the loop but adds only 45° at the pole's corner. This leaves just a 45° phase margin, which can significantly peak the closed-loop response. You must place the high-frequency pole for the noise gain before its intersection with the open-loop response to get a flat closed-loop frequency response.

Second, because the noise gain intersects the open-loop response at a high gain set by \( 1+C_S/C_F \), it's unnecessary for you to use a unity-gain-stable op amp for the transimpedance application. In fact, the achievable transimpedance bandwidth depends directly on the GBP of the op amp. In this case, you can use a very fast but non-unity-gain-stable op amp.

**Determine the feedback capacitance**

You can expand the full transfer function and derive a solution for \( C_F \) that yields the desired frequency response, using **Equation 1** and substituting for \( Z_F \) and \( A(s) \), along with a single-pole model of the op amp's open-loop gain. Using a single-pole, open-loop approximation is always sufficient for unity-gain-stable op amps. For non-unity-gain-stable op amps, this approximation is acceptable if \( 1+[C_S/C_F] \) is greater than the specified minimum stable gain for that op amp. **Equation 3** is this complete transfer function, organized as a standard second-order, lowpass equation.

\[ (3) \]

where \( K_O = \text{dc open-loop gain for the op-amp gain} \), and \( \psi_A = \text{dominant-pole frequency for the open-loop response} \).

You can do a quick check at dc by setting \( s=0 \). The gain is \( R_F x[K_O/(K_O+1)] \), which is correct for a negative-feedback op amp. A second-order transfer function in \( \psi_A \) and Q format can describe this entire transfer function. Rewriting **Equation 3** to show this function yields:

\[ (4) \]
Some reasonable approximations dramatically simplify your subsequent analysis. Assume that \( K_o + 1 \sim K_o \) (a good approximation for high-open-loop gain amplifiers) and that \( K_o x[C_f/(C_f+C_s)] >> 1 \), which states that the maximum noise gain is less than the open-loop gain for the op amp, normally a safe assumption. This calculation results in considerable simplification in the equations for \([\text{psi}]_a\) and \(Q\). Simplify the expressions shown in Equation 4 by using \( K_o [\text{psi}]_a = \text{GBP} \), the GBP for the op amp in radians and setting \([1/(R_F x(C_F+C_s))] = Z_1\) (radian low-frequency zero for the noise gain). This procedure yields Equations 5a and 5b.

\[
C_s \text{ and the desired transimpedance gain, } R_{fy}, \text{ predominately determine } Z_1, \text{ because you set } C_f \text{ low relative to } C_s \text{ to get a high noise gain at } P_1. \text{ Removing } C_f \text{ from the equation for } Z_1 \text{ significantly simplifies isolating a solution for } C_f. \text{ Given this fact, you determine the } [\text{psi}]_a \text{ of Equation 5a independently of } C_f. \text{ However, } C_f \text{ controls the } Q, \text{ as Equation 5b shows. This equation shows that the root-loci-vs-}C_f\text{ plot is approximately a circle with a radius that is the geometric mean of GBP and } Z_1 \text{ sets. You can simplify Equation 5b by recognizing that } [C_f/(C_f+C_s)] = [Z_1/P_1]. \text{ Using this equation, Equation 5b becomes}
\]

\[
Now, [\text{psi}]_a \text{ is always the geometric mean of the op-amp GBP and the zero that the source capacitance of the diode and the desired transimpedance gain ([psi]_a = [square root](GBPxZ_1)) form. Typically, you set the high-frequency pole } P_1 = (1/R_F C_f), \text{ at some fraction } [\text{alpha}] \text{ of } [\text{psi}]_a. \text{ Substituting } [\text{alpha}][\text{square root}(GBP3Z_1) \text{ for } P_1 \text{ into Equation 6 yields}
\]

\[
Now, test the impact of various choices for [\text{alpha}]. You often place } P_1 \text{ exactly at the geometric mean of GBP and } Z_1, \text{ which places } P_1 \text{ at the intersection of the noise gain and the open-loop response (Figure 3) ([\text{alpha}] = 1). Alternatively, you can place } P_1 \text{ at one-half of this geometric mean ([\text{alpha}] = 0.5). If you desire maximum bandwidth with no peaking, then use } Q = [\text{alpha}] = 0.707. \text{ Table 1 shows the effect of selecting these different as.}
\]

<table>
<thead>
<tr>
<th>Desired Q (a)</th>
<th>Peaking (dB)</th>
<th>-3-dB bandwidth</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>1.25</td>
<td>1.27 [psi]0</td>
</tr>
<tr>
<td>0.5</td>
<td>None</td>
<td>0.644 [psi]0</td>
</tr>
<tr>
<td>0.707</td>
<td>None</td>
<td>[psi]0</td>
</tr>
</tbody>
</table>

Setting } Q = 0.5 \text{ by setting } [1/R_F C_f] = 0.53[square root]GBPxZ_1 \text{ yields repeated real poles at } [\text{psi}]_a \text{ for the closed-loop response. This approach is unnecessarily band-limited for a given op amp and } Z_1. \text{ You can get a 36\% increase in bandwidth for the same op amp with no peaking by simply setting
\[1/R_{\text{f}}C_{\text{f}}\] = 0.707 \times \text{square root} \text{GBP} \times Z_{\text{i}}.

The following simple result makes several assumptions along the way. **Equation 8** shows an exact solution for \(1/P_{\text{i}} = R_{\text{f}}C_{\text{f}}\) to achieve a desired \(Q\), given \(C_{\text{s}}\), \(R_{\text{f}}\), and GBP. Solve for \(x = R_{\text{f}}C_{\text{f}}\) to set \(C_{\text{f}}\).

\[x(8)\]

where \(x = R_{\text{f}}C_{\text{f}}\), \(y = R_{\text{f}}C_{\text{s}}\), \(K_{\text{o}} = \text{dc open-loop gain for op amp}, [\psi]A = \text{dominant open-loop pole for op amp in radians}, \) and \(\text{GBP} = K_{\text{o}}[\psi]A\).

By setting \(P_{\text{i}}\) at 0.707 of the geometric mean of GBP and \(Z_{\text{i}}\), the resulting closed-loop bandwidth offers a maximally flat, second-order response with a -3-dB bandwidth equal to that geometric mean. Using this assumption on setting \(P_{\text{i}}\), the minimum required GBP to achieve a desired -3-dB bandwidth is simply \(\text{GBP} = 2\pi \times (F - 3\text{ dB})/Z_{\text{i}}\) (with \(F - 3\text{ dB}\) in hertz, \(Z_{\text{i}}\) in radians, and GBP in hertz). From an op-amp-selection standpoint, this point is a good start for setting a minimum required amplifier gain bandwidth to achieve a desired -3-dB bandwidth, given \(Z_{\text{i}} = [1/R_{\text{f}}C_{\text{s}}]\). This approach assumes that you set \(P_{\text{i}} = [1/R_{\text{f}}C_{\text{f}}]\) to get a maximally flat, Butterworth, second-order response \((P_{\text{i}} = 0.707 \times \text{square root} \text{GBP} \times Z_{\text{i}} = \text{square root} \text{GBP} \times Z_{\text{i}}/2)\).

**Transimpedance noise analysis**

A combination of three noise sources dominates the total output noise for the transimpedance amplifier. These sources are the inverting input-current noise for the op amp, the noninverting input-voltage noise for the op amp, and the noise of the feedback resistor itself. Each source has a different gain to the output, which you must consider in computing the total output noise (**Figure 4**). Also, for low-frequency systems, consider the [1/f] char.

You can develop a relatively simple means of determining which op amp yields the lowest integrated noise at the output by comparing a computed equivalent input-current noise. You can place this noise, a white (constant power over frequency) current-noise source, at the input of a noiseless transimpedance stage to produce the same integrated output noise over a certain noise-power bandwidth (NPB) as that of the actual amplifier. NPB is that rectangular frequency span that encloses the same power as the actual frequency response of the system (see box, "Analyze noise sources").

The analysis to set the compensation implies a well-defined GBP for the op amp. Most voltage-feedback op amps have a GBP that is much more tightly controlled from part to part than is \(K_{\text{o}}\). However, this analysis is approximate. Most systems have some filtering after the gain stages to set the system signal and noise bandwidth. Because some variation exists in the signal bandwidth you computed by setting \(P_{\text{i}}\) (**Figure 3**), subsequent filtering should always be controlling the system bandwidth to less than \(P_{\text{i}}\). Even though each of the noise terms has a different frequency response to the output of the amplifier, this narrower system bandwidth sets the NPB for equivalent input-current noise-calculation purposes.

To simplify this analysis, consider the output noise of each of these terms only as high as a frequency less than \(P_{\text{i}}\). Because this approach limits you to the flat portion of the signal’s frequency response, referring the total output voltage noise to input is simply a matter of dividing by the feedback resistor, \(R_{\text{f}}\). Finally, for broadband (greater than 1 MHz) applications, you can neglect the contribution of noise in the [1/f] region (**Reference 2**).
Now, compute the total output-noise voltage up to $F$.

1. The op amp's input-current noise, $I_n$, has a gain of $R_F$ to the output. The feedback resistor noise appears as $\sqrt{4kTR_F}$ as an output-voltage-noise contribution. The noninverting input-voltage noise, $E_{vN}$, starts out with a unity gain to the output but then has an increasing gain at frequencies higher than $Z_1$ (Figure 3). Assume an NPB less than $P_1$, and integrate just the differentiator part of the response up to $F=NPB$. Equation 9a shows the voltage noise at the output for this term, and Equation 9b shows just the magnitude for this gain. (Phase would be meaningless in combining uncorrelated noise sources.)

\[(9)\]

Next, find an average value over the range 0 to $F$ (in hertz) for this squared output-noise voltage. Equation 10 shows this integral in hertz instead of radians:

\[(10)\]

Now, combine all three noise contributions at the output, as $E_0$ in Equation 11 shows.

\[(11)\]

You then input-refer this total equivalent output-noise voltage dividing by the gain, $R_F$.

\[(12)\]

This equivalent spot input-noise current is the combined total effect of all three terms considered up through a NPB=$F$, where $F$ is less than the pole set by $[1/R_0C_p]=P_1$. Equation 12 also implies that $P_1$ has been set to limit peaking in the frequency response and that the signal gain up to $F$ is simply $R_F$, and it then rolls off steeply at $F$ by filtering after this transimpedance stage. The equation shows that the last term under the radical can be dominant for broadband transimpedance amplifiers. The noise power that this term contributes exceeds that of the first three terms for NPBs greater than that shown in Equation 13.

\[(13)\]

where $Z_1$ is the zero in Figure 13.

**FET or bipolar for best noise?**

Both voltage and current noise for the op amp contribute to the total equivalent input spot-noise current (Equation 12). Which of the two dominates depends on their relative values, the circuit configuration, and the NPB=$F$. You can determine a minimum GBP and the required $C_p$ set using the compensation analysis, given $C_\mu$, a desired $R_F$, and a maximum desired bandwidth. With this information, you can select actual op amps. Use Equation 12 to determine which of a likely group of op amps yields the lowest integrated noise when $F$ sets the NPB. However, it's useful to plot iso-
input spot current-noise curves, given a starting point for one amplifier, as a means of illustrating the trade-off of op-amp noise voltage and current. These curves would be of equal total equivalent input-current noise, trading off the current and voltage noise for the op amp itself to hold the evaluation of Equation 12 constant.

**Some design examples**

As an example, look at four wideband, voltage-feedback op amps, representing both FET and bipolar input stages, for transimpedance designs (Table 2).

<table>
<thead>
<tr>
<th>Part no.</th>
<th>Type</th>
<th>GBP (MHz)</th>
<th>$E_N$ (nV/\text{[square root/Hz]}</th>
<th>$I_N$ (pA/\text{[square root kHz]}</th>
<th>$C_{\text{DIFF}}$ (pF)</th>
<th>$C_{\text{CM}}$ (pF)</th>
<th>Minimum stable gain</th>
</tr>
</thead>
<tbody>
<tr>
<td>OPA637</td>
<td>FET</td>
<td>80</td>
<td>4.5</td>
<td>0.0016</td>
<td>8</td>
<td>7</td>
<td>5</td>
</tr>
<tr>
<td>OPA655</td>
<td>FET</td>
<td>240</td>
<td>6</td>
<td>0.0013</td>
<td>1.2</td>
<td>1</td>
<td>1</td>
</tr>
<tr>
<td>OPA651</td>
<td>Bipolar</td>
<td>500</td>
<td>4.6</td>
<td>1.1</td>
<td>2</td>
<td>1</td>
<td>2</td>
</tr>
<tr>
<td>OPA643</td>
<td>Bipolar</td>
<td>950</td>
<td>1.8</td>
<td>2.4</td>
<td>1.1</td>
<td>1</td>
<td>2</td>
</tr>
</tbody>
</table>

For example, consider a large-area detector with a high desired transimpedance gain and bandwidth, with $C_D=40$ pF, $R_F=200$ k(ohm), and $F_{-3\text{dB}}>2$ MHz. If you set $C_F$ following the compensation discussion to get a $Q=0.707$, the required minimum gain bandwidth is OP-AMP GBP=2p(2 MHz)$C_D=201$ MHz. (14)

The last three candidate parts can deliver the bandwidth. Table 3 shows the $C_F$ required to maintain a maximally flat, Butterworth response, including op-amp input parasitic capacitances. $Z_1$ is approximately 20 kHz, if you neglect the input-parasitic C for the op amp. The approximate solution for $C_F$ is nearly equal to the more exact solution using Equation 8. Set the high-frequency pole, $P_1$, in the noise gain to achieve a closed-loop, Butterworth ($Q=0.707$) response. The typical closed-loop transimpedance -3-dB bandwidth is 1.414 $P_1$ in this case. The high-frequency noise gains are the $1+[C_S/C_F]$ factor in the compensation discussion. Clearly, these parts need not be unity-gain-stable, because they never operate near unity-noise gain at loop-gain crossover. The OPA655, the only FET-input part here, is just meeting the bandwidth requirement.

<table>
<thead>
<tr>
<th>Part no.</th>
<th>Approximate $C_F$ (pF)</th>
<th>Exact $C_F$ (pF)</th>
<th>Approximate $P_1$ (MHz)</th>
<th>-3-dB bandwidth (MHz)</th>
<th>Approximate high-frequency noise gain $(1+C_S/C_F)$</th>
</tr>
</thead>
<tbody>
<tr>
<td>OPA655</td>
<td>0.53</td>
<td>0.48</td>
<td>1.5</td>
<td>2.1</td>
<td>81</td>
</tr>
<tr>
<td>OPA651</td>
<td>0.37</td>
<td>0.35</td>
<td>2.2</td>
<td>3.1</td>
<td>114</td>
</tr>
<tr>
<td>OPA643</td>
<td>0.27</td>
<td>0.26</td>
<td>3.0</td>
<td>4.2</td>
<td>157</td>
</tr>
</tbody>
</table>

Use the OPA655’s noise numbers to set a baseline ($I_n$ on Figure 5), decrease the op-amp voltage noise, and increase the current noise to hold the total equivalent input-noise current constant at this.
baseline (Figure 5). This graph is parametric on NPB, set by F in the integrated-noise analysis. The circles represent noise voltage and current combinations available in other parts. Circles below an iso-input noise curve give lower integrated noise for that NPB, compared with the OPA655.

At the full 2-MHz bandwidth you desire, the bipolar input OPA651 is comparable in noise to the OPA655, because the OPA651's voltage noise is much lower than that of the OPA655. The OPA643, with an input-current noise of 2.4 pA, cannot get below the 1.8-pA target that the OPA655 establishes. The OPA637 appears on the graph and would be a superior selection if you want $F = 0.5$ MHz. If you drop the target bandwidth to 500 kHz, the GBP must be $>50$ MHz, which is within the capability of the OPA637.

As another example, consider a smaller area detector for faster, smaller transimpedance gains, with $C_D = 10$ pF, $R_F = 10$ k(ohm), and $F$-3 dB $> 10$ MHz. Then, if you set $C_F$ to get a $Q = 0.707$, the required minimum GBP is

$$\text{OP-AMP GBP} = 2p(10 \text{ MHz})^2 CD R_F = 62 \text{ MHz. (15)}$$

Although all four parts of Table 4 should satisfy the bandwidth requirement, their input parasitic capacitance may limit their performance. Table 4 shows the achievable results, including the input parasitic capacitance for each part, where you set $C_F$ to get a maximally flat, Butterworth response.

<table>
<thead>
<tr>
<th>Part no.</th>
<th>Approximate $C_F$ (pF)</th>
<th>Exact $C_F$ (pF)</th>
<th>Approximate $P_1$ (MHz)</th>
<th>-3-dB bandwidth (MHz)</th>
<th>Approximate high-frequency noise gain $(1+C_S/C_F)$</th>
</tr>
</thead>
<tbody>
<tr>
<td>OPA637</td>
<td>3.16</td>
<td>3.11</td>
<td>5.0</td>
<td>7.1</td>
<td>8.9</td>
</tr>
<tr>
<td>OPA655</td>
<td>1.27</td>
<td>1.26</td>
<td>12.5</td>
<td>17.7</td>
<td>14.8</td>
</tr>
<tr>
<td>OPA651</td>
<td>0.87</td>
<td>0.86</td>
<td>18.2</td>
<td>25.8</td>
<td>20.1</td>
</tr>
<tr>
<td>OPA643</td>
<td>0.63</td>
<td>0.63</td>
<td>25.1</td>
<td>35.6</td>
<td>21.0</td>
</tr>
</tbody>
</table>

The OPA637 adds 15 pF to the 10-pF source-diode capacitance. This addition limits the bandwidth to less than the desired 10 MHz. Including that 15 pF for the OPA637 shows that it can achieve a 7-MHz transimpedance bandwidth in this application. The other three parts easily satisfy the required bandwidth, as the location of $P_1$ shows. Place $P_1$ here to achieve a maximally flat, Butterworth response in the closed-loop transimpedance response. Also, the high-frequency noise gain is high enough to hold all of these amplifiers stable. Starting with the OPA655 as a baseline design at a 10-MHz NPB, you can generate curves of equal equivalent input-noise current ($I_{N\text{ in}}$ in Figure 6). Figure 6 shows the results of this analysis, where you add 2 pF of input parasitic capacitance to the 10-pF diode source for noise analysis.

With their combination of voltage and current noise, both the OPA651 and OPA643 bipolar input-stage amplifiers provide less integrated noise over a 10-MHz NPB than the FET-input OPA655. The OPA651 performs better down to about 6 MHz NPB because of the dominant contribution of the last term under the radical of Equation 12. The OPA637 may be the lowest noise choice if the NPB is lower than 7 MHz. However, you must include the OPA637’s higher input parasitic capacitance in the integrated-noise analysis. Evaluating Equation 12 for the OPA637 at NPB=5 MHz yields 2.4 pA. This result is higher than those of either the OPA655 or the OPA651. In this case, therefore, a FET part with lower voltage noise than the OPA655 has higher integrated noise due to the part’s lower
in general, higher source-capacitance, lower bandwidth designs benefit from the low noise current of a FET input op amp. As the diode capacitance decreases and the desired bandwidth increases, the low input-voltage noise of a bipolar input gives better results. Many dedicated transimpedance amplifier components do not separately describe their input-voltage noise. As you increase the diode source capacitance from what the data sheets specify, you often see significantly increased output noise from this unspecified noise contribution. Direct evaluation of Equation 12 provides an easy means of comparison between possible choices.

Michael Steffes is a strategic marketer for high-speed signal-processing components at Burr-Brown Corp, Tucson, AZ. He has a BSEE from the University of Kansas (Lawrence, KS) and an MBA from Colorado State University (Fort Collins, CO) and has helped develop numerous amplifier ICs. His spare-time interests include history, classic literature, travel, and running.

References


Analyze noise sources

It's useful to review the key features of noise analysis. You can combine independent noise sources at the output by adding their powers. After combining, compute an equivalent total-noise voltage by taking the square root of the sum of squared contributing noise voltages. (In other words, uncorrelated noise powers add by superposition.) Integrated noise converts from the noise power in a 1-Hz bandwidth (sometimes called spot noise) to the noise power over some bandwidth of interest. If a voltage describes the noise and if the noise is constant over frequency, integrated noise is the spot noise multiplied by the square root of the NPB.
If the noise voltage is not flat over frequency, you derive an equivalent flat-noise voltage by computing the average of the noise voltage squared over the frequency band of interest. Although you often compute the total noise at the output, an equivalent input noise is often of more interest in deriving a S/N ratio. You do this by input-referring the total output noise. Divide the total output noise by the gain from the desired input signal point to the output (Reference 2).

To apply these ideas to a transimpedance amplifier, consider the following. The amplifier's input-current noise, \( I_{n} \), has the same frequency response to the output as the input signal in the second-order response. The op amp's input-voltage noise, \( E_{n} \), has a gain to the output that at first follows the noise-gain part of Figure 3 until the gain intersects the open-loop roll-off of the amplifier. Intuitively, the gain then rolls off with a one-pole characteristic. The feedback-resistor noise appears directly at the output and is band-limited by the \( 1/R_{f}C_{f} \) pole with a single-pole roll-off. Each of these terms has a slightly different frequency response to the output. A complete analysis would separately consider frequency-response shapes in computing integrated noise (Reference 1).