Transimpedance Considerations for High-Speed Amplifiers

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ABSTRACT

Designing high-resolution detection circuits using photodiodes presents considerable challenges because bandwidth, gain, and input-referred noise are coupled together. This application note reviews the basic issues of transimpedance design, provides a set of detailed design equations, explains those equations, and develops an approach to easily compare potential solutions.

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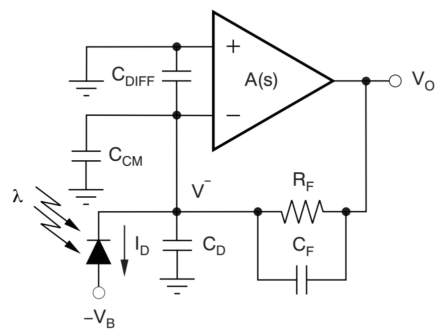
1 Introduction

The purpose of a transimpedance circuit is to convert an input current from a current source (typically a photodiode) into an output voltage. The simplest method to achieve this conversion is to use a resistor connected to ground. However, the achievable gain using this method is limited by the following factors:

1. The current source input impedance;
2. The load impedance; and
3. Desired bandwidth.

A closed-loop approach, using an operational amplifier, is normally beneficial to most applications as it has the potential of eliminating those issues. A typical circuit with all necessary components for this analysis is shown in Figure 1. In this circuit, the generator is a photodiode, whose role is to convert the photons into a current. This current is then amplified by the feedback resistor R_F. Working with an ideal amplifier for now, we can see that because no bias current is present, all the signal generated by the photodiode is going to be converted to the output. Note that the source impedance is not of interest here; it is isolated from the output. Note also that the output is low impedance, allowing a wide variety of load to be connected. The bandwidth is going to be a function of the source capacitance (C_S), the feedback capacitance (C_F), and the gain bandwidth product of the actual amplifier used.



$C_S = C_D + C_{CM} + C_{DIFF}$ $V_O = -V^- A(s)$

Figure 1. Transimpedance Circuit with Modeled Elements

The source capacitance (C_S) is the sum of the photodiode capacitance (C_D), the common-mode capacitance of the amplifier (C_{CM}), and the differential capacitance of the amplifier (C_{DIFF}). C_{CM} and C_{DIFF} include both the board layout and the op amp parasitic capacitance.

2 Stability Analysis

The first non-ideal op-amp characteristic we examine here is the non-infinite open-loop gain. The architecture for the operational amplifier used in the rest of this application report is a single pole op-amp model, as shown in Equation 1. This model allows us to analyze the resulting transimpedance design as a second-order, closed-loop transfer function.

$$A(s) = \frac{A_{OL} \cdot \omega_A}{s + \omega_A} \tag{1}$$

Expressing the Laplace transfer function in Bode analysis form yields the following equation.

$$\frac{V_O}{I_D} = \frac{-Z_F}{1 + \left(\dfrac{1 + \dfrac{Z_F}{Z_G}}{A(s)} \right)} \tag{2}$$

where:

$$Z_F = R_F \parallel \frac{1}{s \cdot C_F} = \frac{\frac{1}{C_F}}{s + \frac{1}{R_F \cdot C_F}} \tag{3}$$

$$Z_G = \frac{1}{s \cdot C_S} \tag{4}$$

At dc, and if the open-loop gain of the amplifier is infinite, the amplifier gain is set by the feedback resistor. This effect can be seen by setting s = 0 in Equation 3 and plugging the result into Equation 2 while setting A(s) = ∞.

$1 + \frac{Z_F}{Z_G}$ is the noise gain. The loop gain is the difference between the open-loop gain (A_{OL}) and the noise gain. The noise gain is plotted in Figure 2 alongside the single pole, open-loop gain model of the amplifier.

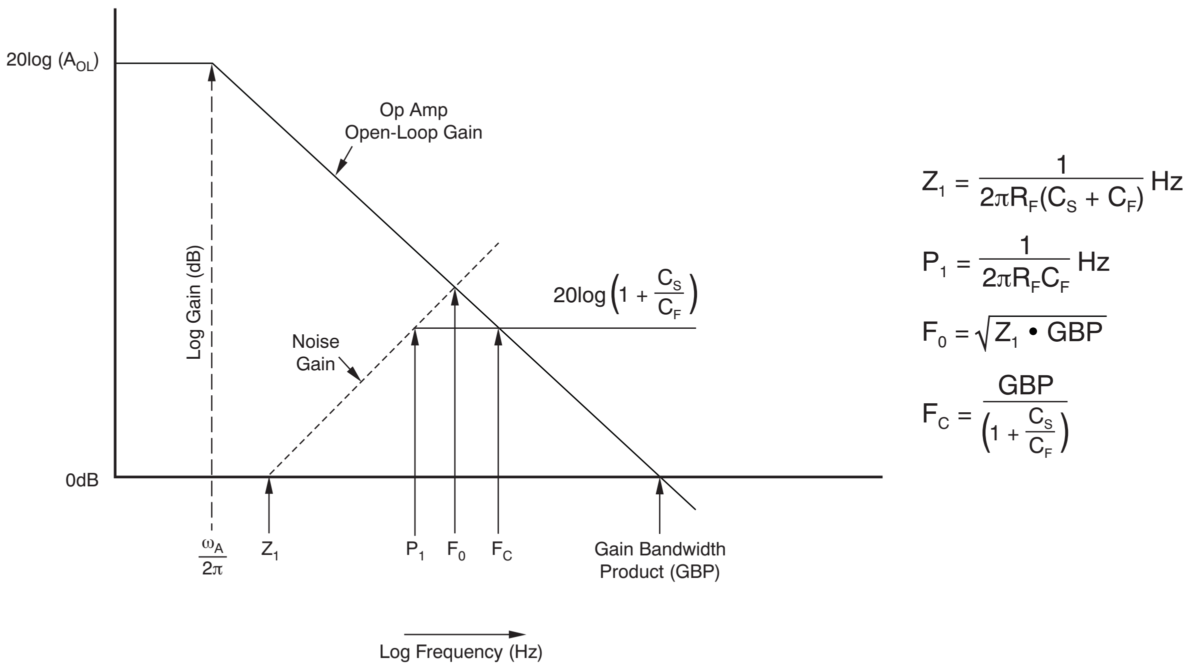


Figure 2. Bode Plot Magnitude

At dc, the gain is indeed the expected transimpedance gain R_F, while at high frequency, it is $1 + \frac{C_S}{C_F}$. Note that only the feedback capacitor (C_F) and the source capacitance (C_S) are used for stability; consequently, it can be noted that a unity-gain stable amplifier is not necessary for transimpedance applications. In fact, it is recommended to use a decompensated amplifier instead, because these decompensated amplifiers offer better voltage noise specifications and larger gain bandwidth products than any compensated version. This fact is easily illustrated by two families of devices: the OPA842/3/6/7 and the OPA656/7. The OPA842 is unity-gain stable and has 2.7nV/√Hz input voltage noise. The OPA843 is stable for gains greater than 3V/V and has a voltage noise of 2nV/√Hz. The OPA846 is stable for gains greater than 7V/V and has a 1.2nV/√Hz voltage noise. Finally, the OPA847 has a voltage noise of 0.85nV/√Hz and is stable for gains greater than 12V/V. For FET devices such as the OPA656 and OPA657, the OPA656 provides unity-gain stability with a 7nV/√Hz input voltage noise; the OPA657 has a 4.8nV/√Hz voltage noise, but is stable for gains greater than 7V/V.

Coming back to the analysis and expressing the full transfer function as a function of ω_o and Q.—given in Equation 5 with Equation 6 and Equation 7 expressing ω_o and Q with physical elements of the circuit —yields the following results:

$$\frac{V_O}{I_D} = R_F \bullet \frac{A_{OL}}{A_{OL} + 1} \bullet \frac{\omega_o^2}{s^2 + s \bullet \frac{\omega_o}{Q} + \omega_o^2} \tag{5}$$

$$\omega_o = 2\pi \bullet F_0 = \sqrt{\frac{(A_{OL} + 1) \bullet \omega_A}{R_F \bullet (C_S + C_F)}} \tag{6}$$

$$Q = \frac{\sqrt{\frac{(A_{OL} + 1) \bullet \omega_A}{R_F \bullet (C_S + C_F)}}}{\omega_A \bullet \left(1 + A_{OL} \bullet \frac{C_F}{C_S + C_F}\right) + \frac{1}{R_F \bullet (C_S + C_F)}} \tag{7}$$

Setting $Z_1 = \frac{1}{R_F \bullet (C_S + C_F)}$, and recognizing that the gain bandwidth product (GBP) is equal to $GBP = \frac{A_{OL} \bullet \omega_A}{2\pi}$ and using the following algebraic simplifications:

- $C_S >> C_F$ to simplify Z_1
- $(A_{OL} + 1) \bullet \omega_A \cong A_{OL} \bullet \omega_A = 2\pi \bullet GBP$
- $\left(1 + A_{OL} \bullet \frac{C_F}{C_S + C_F}\right) \cong A_{OL} \bullet \frac{C_F}{C_S + C_F}$

Leads to the following:

$$F_0 = \sqrt{Z_1 \bullet GBP} \tag{8}$$

$$Q = \frac{F_0}{Z_1 + F_C} \tag{9}$$

Further simplification on Q leads to these easier-to-use equations:

$$F_0 = \sqrt{Z_1 \bullet GBP}$$

$$Q = \frac{P_1}{F_0} \tag{10}$$

with:

$$P_1 = \frac{1}{2\pi \bullet R_F C_F} \tag{11}$$

3 Bandwidth Considerations

Maximum bandwidth will be achieved for a Butterworth response with Q = 0.707. This result is an interesting point because this is also equivalent to saying that the –3dB bandwidth is equal to F_0, and allows us to derive a design equation that proves very useful for amplifier selection.

$$GBP = 2\pi \bullet F^2_{-3dB} \bullet R_F C_S \quad (12)$$

Note that the –3dB bandwidth and the transimpedance gain are desired parameters and the source capacitance is known. Thus, Equation 12 allows you to do the following:

- Knowing the bandwidth required by the application, the photodiode capacitance, and the transimpedance gain specification, calculate the minimum GBP requirement for the amplifier
- Knowing the amplifier, the transimpedance gain, and the photodiode capacitance, calculate the maximum achievable bandwidth

We can use this last point to calculate the maximum achievable bandwidth for a few selected amplifiers. The OPA847 is a 3.9GHz GBP amplifier with 0.85nV/$\sqrt{Hz}$ input voltage noise. The OPA846 has a 1.75GHz GBP with a 1.2nV/$\sqrt{Hz}$ input voltage noise. The OPA843 is a 800MHz GBP device with a 2nV/$\sqrt{Hz}$ input voltage noise. All three amplifiers—OPA847, OPA846 and OPA843—are bipolar amplifiers. Finally, the OPA657 is a 1.65GHz GBP device with a 4.8nV/$\sqrt{Hz}$ input voltage noise. Unlike the other three devices, the OPA657 is a FET input amplifier.

Figure 3 shows an example of achievable bandwidth versus transimpedance gain for a 10pF source capacitance for various amplifiers.

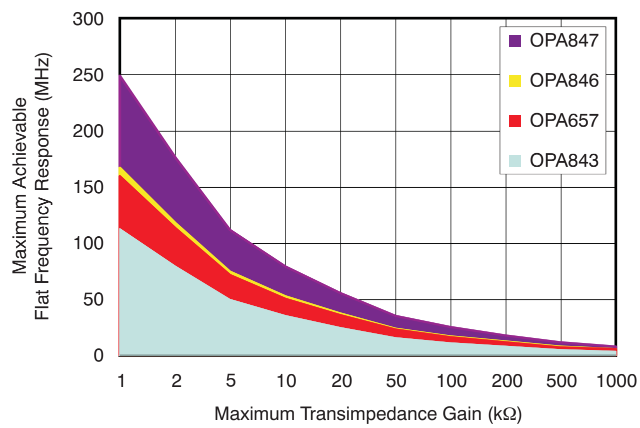


Figure 3. Maximum Achievable Bandwidth for Selected Op Amps (10pF Source Capacitance)

The relative performance for each amplifier is maintained as the source capacitance varies. Note that the source capacitance in this example considers the amplifier parasitic, both common-mode and differential capacitance. When multiple amplifiers can be selected to meet the desired performance—for example, for R_F = 20kΩ, f_{-3dB} = 10MHz—other considerations such as noise or power dissipation must be considered before making the amplifier selection.

At first glance, using a FET input amplifier does not present any advantages and has one major inconvenience: its high input voltage noise. The noise analysis will be developed next and then the relative performance of bipolar technology will be compared with FET technology.

4 Noise Considerations

The equivalent input current noise for a transimpedance amplifier makes the following assumptions:

- This evaluation is an integrated noise analysis that uses spot noise over frequency, and is not intended to be used as a spot noise equation for narrowband applications.
- The application is dc-coupled, pulse-oriented where the integrated noise is of interest.
- The final signal bandwidth for both the transimpedance design and any post-filtering is greater than 10 times the 1/f noise corner frequency of any of the amplifier noise terms. This consideration allows those effects to be neglected.
- The transimpedance bandwidth is set greater than the post-filtering.
- The current noise term on the noninverting is either negligible or made negligible by providing adequate bypassing.

With these assumptions at hand, the noise terms of interest are shown in Figure 4.

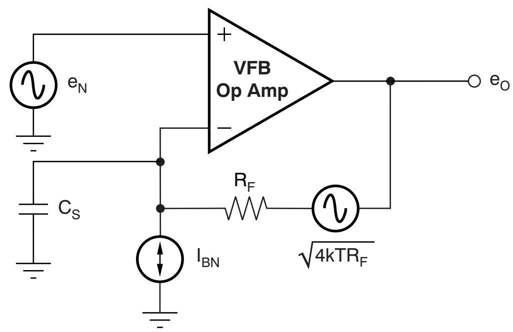


Figure 4. Noise Analysis Circuit

The equivalent input-referred noise current for wideband transimpedance design is provided in Equation 13.

$$i_{EQ} = \sqrt{i_B^2 + \frac{4kT}{R_F} + \left[\frac{e_N}{R_F}\right]^2 + \frac{(e_N 2\pi F C_S)^2}{3}} \tag{13}$$

where:

- I_B = inverting input spot current noise
- $4kT = 16 \times 10^{-21}$J at 290 degrees Kelvin
- R_F = feedback resistor
- e_N = noninverting input spot voltage noise
- C_S = inverting input total capacitance
- F = noise integration frequency limit

All these parameters except the noise integration frequency limit can be found in the respective operational amplifier product data sheet. The noise integration frequency limit (F) represents the equivalent *brick-wall* filter of a post-amplifier passive filter.

Using Equation 13 and comparing two amplifiers with very similar GBP—the OPA846 and the OPA657—we can then determine an appropriate transimpedance gain threshold. Below this threshold, it is preferable to use bipolar technology to achieve lower noise, while above the threshold it is better to use FET technology. For this comparison, we will consider that the noise integration frequency limit is the same for both amplifiers. This problem is expressed in Equation 14.

$$\sqrt{i_B^{\,2}{}_{(BIP)} + \frac{4kT}{R_F} + \left(\frac{e_{N(BIP)}}{R_F}\right)^2 + \frac{(e_{N(BIP)} \bullet 2\pi \bullet F \bullet C_{S(BIP)})^2}{3}} = \sqrt{i_B^{\,2}{}_{(FET)} + \frac{4kT}{R_F} + \left(\frac{e_{N(FET)}}{R_F}\right)^2 + \frac{(e_{N(FET)} \bullet 2\pi \bullet F \bullet C_{S(FET)})^2}{3}} \quad (14)$$

The solution for R_F is shown in Equation 15.

$$R_F = \sqrt{\frac{e_N^{\,2}{}_{(FET)} - e_N^{\,2}{}_{(BIP)}}{i_B^{\,2}{}_{(BIP)} - i_B^{\,2}{}_{(FET)} + \frac{2\pi \bullet F}{3} \bullet (C_{S(BIP)} \bullet e_N^{\,2}{}_{(BIP)} - C_{S(FET)} \bullet e_N^{\,2}{}_{(FET)})}} \quad (15)$$

For example, with a band-limit filter bandwidth set at 10MHz and a 10pF diode capacitance, all other parameters are set according to Table 1 (taken from the OPA657 and OPA846 product data sheets).

Table 1. Device Performance Comparison

Device	Element Specifications
OPA657	$e_{N(FET)}$ = 4.8nV/√Hz
	$I_{B(FET)}$ = 1.3fA/√Hz
	$C_{S(FET)}$ = 10pF + 5.2pF = 15.2pF
OPA846	$e_{N(BIP)}$ = 1.2nV/√Hz
	$I_{B(BIP)}$ = 2.8pA/√Hz
	$C_{S(BIP)}$ = 10pF + 3.8pF = 13.8pF

The result is that for a resistor lower than 2kΩ, the bipolar amplifier offers a noise advantage; we can therefore conclude that even though the voltage noise of the OPA657 is high, the total input-referred noise generated by the OPA657 FET amplifier will be lower than that of the OPA846 bipolar amplifier for any transimpedance gain greater than 2kΩ. If a larger post-amplifier filter bandwidth is necessary, the bipolar amplifier (such as the OPA846) will have the lowest noise contribution independent of the resistor.

In general (and from a noise perspective), FET input amplifiers such as the OPA657 are best for large or very large transimpedance gain with low-to-medium bandwidth because of the post-amplifier filter limitations, whereas bipolar amplifiers such as the OPA846 are best for medium-to-large transimpedance gain applications with high bandwidth as limited by the post-amplifier filter.

5 Example

From Figure 3, the maximum achievable bandwidth versus transimpedance gain, it is difficult to separate the OPA846 from the OPA657 simply by achievable bandwidth. In order to help assess the behavioral differences between these two amplifiers, first consider the following application circuit (Figure 5).

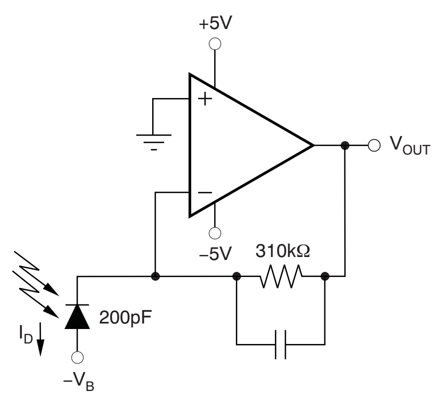
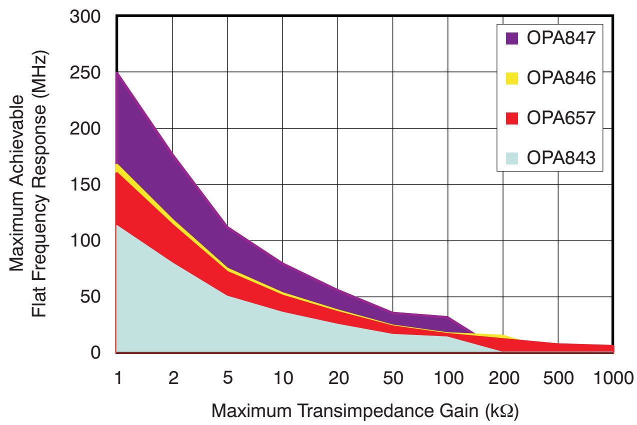


Figure 5. 310kΩ Transimpedance Gain with 200pF Source Capacitance Circuit

In this circuit, using Equation 12, the OPA846 achieves 2.1MHz while the OPA657 achieves 2MHz, making them equivalent on that specification. Looking at the noise, the OPA657 yields a 5pA/$\sqrt{Hz}$ equivalent input noise current if we assume the noise power bandwidth limit is set at 1.4MHz. This noise is dominated by the third term of the total noise equation, which is the effect of the rising portion of the noise gain curve times the relatively high 4.8nV/$\sqrt{Hz}$ input voltage noise for the OPA657. The OPA846, which has a much lower voltage noise but much higher current noise, actually yielded a lower equivalent input noise current of 3pA/$\sqrt{Hz}$. So why is the OPA846 not used in this application?

6 DC-Parameters Consideration

The input bias current of the OPA846, 19µA, generates an output offset voltage with the feedback resistor of 310kΩ of 5.89V. Because the OPA846 is operating on a ±5V power supply, this offset voltage sends the output into saturation. Adding a 310kΩ resistor on the noninverting input allows bias current cancellation but now puts 5.89V common-mode voltage on the input, exceeding the common-mode input range of the OPA846. Figure 6 shows the maximum achievable bandwidth as a function of the transimpedance gain for a few operational amplifiers when taking into consideration dc parameters. The difference between bipolar technology and FET-input technology is now easy to establish. FET-input operational amplifiers, such as the OPA657, are capable of higher transimpedance, where decompensated bipolar operational amplifiers are capable of much higher bandwidth but are limited in gain range.



This measurement takes into consideration a dc parameter for a 10pF source capacitance.

Figure 6. Maximum Achievable Bandwidth vs Transimpedance Gain for Selected Op Amps

7 Conclusion

Although all operational amplifiers can be used in transimpedance applications, the limit in performance is always limited by the transimpedance gain, the bandwidth, and the noise. Many potential problems have been discussed along with suggestions and solutions. This application report developed an emphasis on the methodology to approach transimpedance designs as well as component selection.