Detector Preamp Frequency: the interface problem

A guide to detector/preamp connections for the electro-optical systems designer

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The high-frequency response characteristics of solid-state photodetectors are determined by transit times of carriers in the photodetector and by the electrical impedance parameters that describe the detector/preamp interface. Since PIN photodiodes with transit times of less than 1 nanosecond are now available to the e-o systems designer, the preamplifier input circuit usually establishes the maximum frequency of operation.

Alongside the growing optical communications field is a need for higher data rates. That means the systems designer will have to understand the detector/preamp interface. Here, we describe the two most commonly used detector/preamp connections in enough detail for the designer to establish operating limits and make the necessary design trade-offs to optimize his particular system.

The photodiode equivalent circuit

This analysis must begin with an accurate representation of photodetector impedance. The small-signal, a.c. equivalent circuit illustrated in Figure 1a can be derived by considering the physics of the photodetector.[1] The current generator represents the photo-current induced by the optical signal. The leakage current of the diode is not included since this is a d.c. current.

R_p is the dynamic resistance of the junction at the operating bias point of the photodiode. Empirically, this can be determined by measuring the slope of the voltage versus current curve through the operating point.

C_p is the junction capacitance of the photodiode at the bias point. C_p is inversely proportional to the square root of applied plus "built-in" reverse voltages.

R_s is the series resistance of the photodiode and contains contributions from contact resistance, diffused layer spreading resistance and substrate resistance. Sometimes this parameter is ignored in photodetector analysis, but it can have a profound influence on high-frequency performance and should not be overlooked.

Figure 1a represents the small-signal a.c. equivalent circuit of the photodetector in either the photovoltaic or photoconductive modes (unbiased or inverse biased). We need only to insert the appropriate parameters measured at the operating point. Noise generators have not been included in the equivalent circuit since we do not treat noise in this discussion.

Finally, it is important to stress that the equivalent circuit shown is a "small-signal" representation. A silicon photodiode is limited in the amount of current it can deliver. Figure 1a assumes operation well below photodetector saturation.

While the figure is helpful in understanding the physical parameters that affect photodiode operation, other equivalent circuit representations are easier to use in the circuit analysis. The two equivalent circuits we will use are illustrated in Figures 1b and 1c. These are Norton's equivalent circuit and Thevenin's equivalent circuit, respectively.[2] The time constants shown in both figures are defined through the following general expression:

$$\tau_{Re} \equiv R_R C_e \qquad (1)$$

Series load configuration

Figure 2a shows one common way to interface a photodiode and its preamplifier. This circuit, which we will refer to as the "series load" connection, is most commonly associated with high-frequency applications where reverse bias is applied to the photodetector and a wide bandwidth voltage amplifier is used.

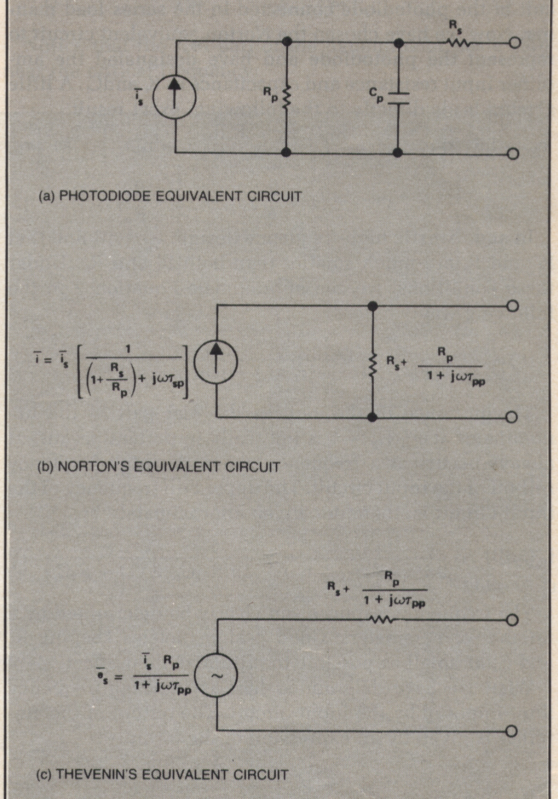

Figure 1.
A REPRESENTATION OF THE PHOTODIODE EQUIVALENT CIRCUIT.

Figure 2.
SERIES LOAD CIRCUIT CONFIGURATION.

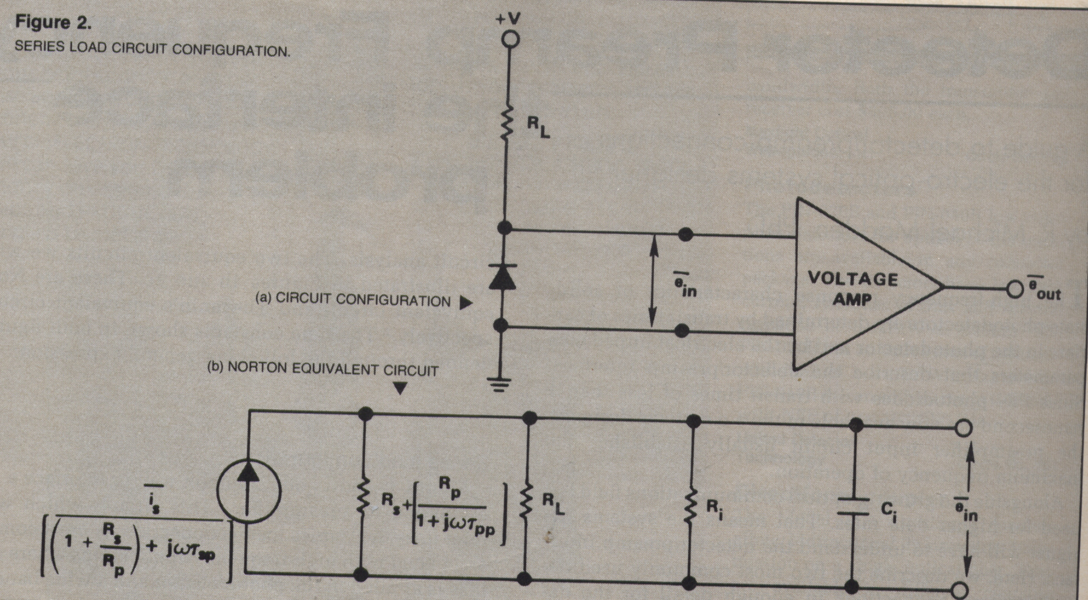

Figure 2b illustrates the small-signal a.c. equivalent circuit of the photodiode connected in the series load manner. Now we have chosen the Norton equivalent circuit to represent the photodiode and have designated the amplifier input resistance and capacitance as R_i and C_i. A little algebra leads directly to the following, exact result:

$$\mathbf{e}_{in} = \mathbf{i}_s R_L \Big/ \left\{\left[1 + \frac{R_L}{R_p} + \frac{R_L}{R_i} + \frac{R_s}{R_p} + \frac{R_s R_L}{R_p R_i} + \omega^2 \tau_{sp}\tau_{Li}\right] + j\omega \left[\tau_{Lp} + \left(1 + \frac{R_L}{R_i}\right)\tau_{sp} + \left(1 + \frac{R_s}{R_p}\right)\tau_{Li}\right]\right\} \quad (2)$$

In any realistic high-frequency design, we will find that R_L and R_s are much smaller than both R_i and R_p. Under these conditions, we can approximate Equation 2 in the following way:

$$\underset{(R_L, R_s << R_p, R_i)}{\mathbf{e}_{in}} \approx \frac{\mathbf{i}_s \mathbf{R}_L}{(1 - \omega^2 \tau_{Li}\tau_{sp}) + j\omega(\tau_{Lp} + \tau_{Li} + \tau_{sp})} \quad (3)$$

The presence of the zero in the real part of the denominator at $\omega^2{}_{Li}\tau_{sp} = 1$ is not really important because it always occurs at a frequency above the high-frequency roll-off of the input circuit. The upper 3dB frequency of the input circuit is, therefore, given approximately by:

$$\underset{(R_L, R_s << R_p, R_i)}{f_{in}(3dB)} \equiv \frac{1}{2\pi(\tau_{Lp} + \tau_{Li} + \tau_{sp})} \approx \frac{1}{2\pi[R_L(C_p + C_i) + R_s C_p]} \quad (4)$$

The frequency response can clearly be determined by a number of parameters. Good design dictates keeping R_s and C_i as small as possible to minimize their effect.

What we have determined here is the high-frequency limit imposed by the input circuit. The output of the amplifier is given by an expression of the form:

$$\mathbf{e}_{out} = \mathbf{e}_{in} \left\{\frac{A_{oo}}{1 + j\omega\tau_o}\right\} \quad (5)$$

Therefore, the inherent bandwidth of the amplifier is:

$$f_o(3dB) = \frac{1}{2\pi\tau_o} \quad (6)$$

The overall high-frequency limit will be determined by the smaller of f_{in} and f_o.

Expressions for the phase and amplitude can be derived by rationalizing Equations 2, 3 and 5.

Transimpedance configuration

In the transimpedance configuration, the photodiode drives the inverting input of an operational amplifier using negative feedback. This circuit connection is shown in Figure 3a for the photovoltaic mode (detector unbiased) and Figure 3b for the photoconductive mode (detector reverse biased). The a.c. equivalent circuit has the same form for both modes and is shown in Figure 3c. In this equivalent circuit, the op-amp must be shown since feedback plays a primary role in the functioning of the circuit.

The basic formula describing the operation amplifier is as follows:[3]

$$\mathbf{e}_{in} = \left[\frac{\mathbf{Y}}{\mathbf{Y} + \mathbf{Y}_L + \mathbf{Y}_i}\right] \mathbf{e}_s + \left[\frac{\mathbf{Y}_L}{\mathbf{Y} + \mathbf{Y}_L + \mathbf{Y}_i}\right] \mathbf{e}_{out} \quad (7)$$

Y is the complex immittance of the photodiode; $\mathbf{Y}_L$ is the complex immittance of the feedback loop; and $\mathbf{Y}_i$ is the complex immittance of the amplifier input. The coefficient of $\mathbf{e}_s$ in Equation 7 is the attenuation factor the signal voltage undergoes before reaching the amplifier input. It is often expressed as:[3]

$$\alpha \equiv \frac{\mathbf{Y}}{\mathbf{Y} + \mathbf{Y}_L + \mathbf{Y}_i} \quad (8)$$

The coefficient of $\mathbf{e}_{out}$ in Equation 7 is referred to as the feedback factor in circuit theory and is usually represented as:[3]

$$\beta \equiv \frac{\mathbf{Y}_L}{\mathbf{Y} + \mathbf{Y}_L + \mathbf{Y}_i} \quad (9)$$

The generalized formula for the closed-loop gain of an operational amplifier is given by the well known expression:

$$A \equiv \frac{e_{out}}{e_{in}} = \frac{\alpha A_o}{1 - \beta A_o} \qquad (10)$$

A_o is the open-loop gain of the op-amp and can be expressed as:

$$A = \frac{-A_{oo}}{1 + j\omega\tau_o} \qquad (11)$$

A_{oo} is the magnitude of the open-loop d.c. gain, which is usually very large. τ_o is the effective integration time of the open-circuited op-amp.

Equations 7 through 11 can be used to express $\mathbf{e}_{out}$ in terms of the photodiode signal current $\mathbf{i}_s$:

$$\mathbf{e}_{out} = \frac{-\mathbf{i}_s R_L}{(1 + j\omega\tau_{LL})} \left[\frac{1}{(1 + \frac{R_s}{R_p}) + j\omega\tau_{sp}} \right] \times 1/\left\{1 + \frac{1}{A_{oo}} \qquad (12)\right.$$

$$\left.\left[\frac{1 + j\omega\tau_o}{1 + j\omega\tau_{LL}}\right] \left[(1 + j\omega\tau_{LL}) + (1 + j\omega\tau_{Li}) + \frac{(1 + j\omega\tau_{Lp})}{(1 + \frac{R_s}{R_p}) + j\omega\tau_{sp}}\right]\right\}$$

Equation 12 is an exact expression that can be used to determine the effects of each of the circuit elements of Figure 3. Rationalizing this function will provide both amplitude and phase information.

It is instructive to consider some limiting cases of Equation 12. Suppose, for instance, that the open-loop gain of the op-amp is infinite. Under this condition, Equation 12 reduces to:

$$\lim_{A_{oo} \to \infty} \mathbf{e}_{out} \approx \frac{-\mathbf{i}_s R_L}{(1 + j\omega\tau_{LL})} \left[\frac{1}{(1 + \frac{R_s}{R_p}) + j\omega\tau_{sp}} \right] \qquad (13)$$

The coefficient in Equation 13 is the first order result we would expect if the op-amp were ideal and the parasitic photodiode resistance, R_s, were zero. Notice how R_s modifies the result, even with $A_{oo} \to \infty$.

Another important limiting case arises when $\tau_{LL} \to 0$, $\tau_{sp} \to 0$, and $R_s \to 0$. In this instance, Equation 12 reduces to:

$$\lim_{\tau_{LL}, \tau_{sp}, R_s \to 0} \mathbf{e}_{out} \approx \qquad (14)$$

$$\left\{ \frac{-\mathbf{i}_s R_L}{1 + \frac{1}{A_{oo}} (1 + j\omega\tau_o) \left[1 + (1 + j\omega\tau_{Li}) + (1 + j\omega\tau_{Lp}) \right]} \right\}$$

This expression describes the performance of the circuit when the feedback capacitance has been removed for highest frequency operation and the series resistance of the photodiode is negligible.

By rearranging terms in the denominator of Equation 14, we separate the real and imaginary terms. This yields:

$$\lim_{\tau_{LL}, \tau_{sp}, R_s \to 0} \mathbf{e}_{out} \approx \bar{i}_s R_L / \left[\left[1 + \frac{3 - \omega^2\tau_o (\tau_{Li} + \tau_{Lp})}{A_{oo}} \right] + \frac{j\omega}{A_{oo}} \left[\tau_{Li} + \tau_{Lp} + 3\tau_o \right] \right] \qquad (15)$$

Notice that the real part of the denominator has a zero near the critical frequency:

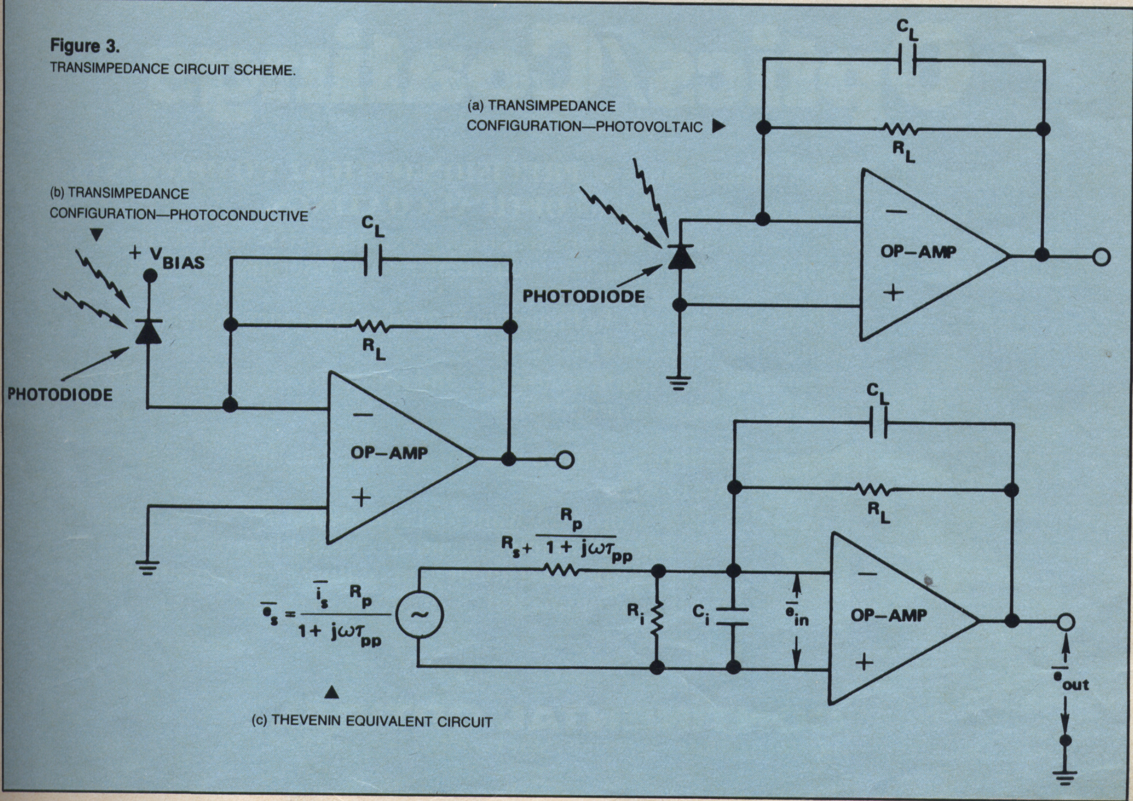

Figure 3.
TRANSIMPEDANCE CIRCUIT SCHEME.

$$\omega_c = \sqrt{\frac{A_{oo}}{\tau_o(\tau_{Li} + \tau_{Lp})}} \qquad (16)$$

Since the imaginary term of the denominator is usually much less than unity at this frequency, the gain peaks and operation become unstable. It then becomes necessary to add enough shunt capacitance in the feedback loop to roll the frequency off below ω_c. We may therefore consider ω_c as the maximum frequency operation of a photodiode in the transimpedance configuration. The frequency, f_c, associated with the radial frequency, ω_c, is given by:

$$f_c = \sqrt{\frac{f(GBW)}{2\pi R_L(C_i + C_p)}} \qquad (17)$$

f(GBW) is the small-signal gain bandwidth product of the operational amplifier.

The advantage of the transimpedance configuration is evident when Equations 4 and 17 are compared. For a given load resistor higher frequency is possible with a transimpedance amplifier. Setting $R_s = 0$ in Equation 4 permits us to write the magnitude of the improvement as:

$$\frac{f(3dB, \text{transimpedance})}{f(3dB, \text{series load})} \approx_{\lim R_s \to 0} \sqrt{\frac{f(GBW)}{f(3dB, \text{series load})}} \qquad (18)$$

The larger the gain bandwidth product, the greater the improvement in frequency response.

For a given bandwidth requirement, a larger load resistance can be used in the transimpedance configuration, yielding a greater output signal, concurrent with lower noise than that achieved with a series load connection.

In very high frequency applications, it is difficult to find operational amplifiers with gain bandwidth products that are large enough to be useful. From Equation 18, we see that, unless the gain bandwidth product exceeds the required upper 3dB frequency, the transimpedance configuration actually degrades high-frequency performance.

Operational amplifiers that give gain bandwidth products greater than about 20 or 30 megahertz are not generally available. Consequently, the series load configuration has usually been utilized above this frequency. There are companies, however, that are now developing op-amps that exhibit gain bandwidth products of several hundred megahertz. Once these are available commercially, they will undoubtedly supplant series load circuits throughout their frequency range, unless their prices are prohibitively high. □

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Meet the author

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