Analysis of Op Amp Based Transimpedance Photo Receivers

A Comprehensive Practical Approach

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***Abstract*—A widely used optical wireless receiver topology is that of a transimpedance amplifier built around an operational amplifier. Although its principle operation is uncomplicated and widely understood, its analysis is rather involved. Moreover, existing literature on the matter includes only part of the analysis, and does not include an account of the type of Operational Amplifier (Op Amp) that should be employed. This paper aims to fill the gap by including a complete, comprehensible assessment of the Op Amp based Transimpedance Amplifier, as well as a rational of the type of operational amplifier these designs should incorporate. In an effort to make the presented investigation accessible also to researchers with limited circuit analysis skills, a detailed practical step-by-step approach has been adopted.**

Keywords—Optical Wireless Communication; Photodiode; Transimpedance Amplifier; Signal Analysis; Noise Analysis; Stability

I. INTRODUCTION

The most effective method, to date, of converting the photocurrent of a large area photodiode into a suitable processing voltage is the Transimpedance Amplifier (TIA). Employing an Operational Amplifier (Op Amp) offers the benefit of reduced complexity and limited component count. Moreover, Op Amp technology has matured such that low noise, high bandwidth devices are available at low cost. For designs of limited complexity and/or designs where budgets for the high development cost of dedicated on-chip solutions are not available, or cannot be covered sufficiently by the anticipated revenue, TIA designs with Op Amps on Printed Circuit Board (PCB) offer a worthy alternative. Low-cost Optical Wireless Communication (OWC) receivers for a specific market niche fit that description.

Although the design of a TIA becomes less complex when implemented with Op Amps, it is often underestimated and not always sufficiently understood. Moreover, design resources such as [1] until [4] as well as a textbook dedicated to this type of circuits [5] do not cover the full analysis picture. This paper aims to provide for a complete and comprehensible account of Op Amp based transimpedance photo receivers, including a thorough noise and stability analysis. The presented material is of particular benefit to developers of OWC receivers who are not analogue circuit designers by trade. Various types of Op Amps are available, and choosing an inappropriate type or inappropriate specifications may lead to designs with an unfavorable outcome. An overview of contemporary Op Amp types, including a rational of the type that needs to be selected for photo transimpedance receivers, is also incorporated.

Section II starts a step-by-step analysis approach by calculating the received signal's current-to-voltage conversion while ignoring noise and assuming ideal Op Amp characteristics. The influence of noise, and how to combat it, is presented in Section III. In Section IV, attention is shifted to the choice of Op Amp, and Sections V and VI incorporate the non-ideal Op Amp input characteristics. Finally, findings of the presented work are summarized in Section VI.

II. SIGNAL ANALYSIS

In order to develop electronics that convert the photo current of a photodiode into a suitable voltage, the photodiode needs to be modeled by its electrical equivalent. A simple, yet accurate and widely used model is the one shown in the highlighted part of Fig. 1. It models the photodiode as a current source that produces the photocurrent i_d [6]. Parallel to this current source are the dark current resistance R_d and the junction capacitance C_d. Between these and the connecting leads of the device is the contact resistance R_c. The current generated by the photodiode needs to be converted into a voltage for further processing. The simplest way to convert a current into a voltage is the employment of a resistor, and to take direct advantage of Ohm's law. Fig. 1 shows the schematic, and its signal equivalent, of a photodiode connected to a resistor. Note that a biasing voltage, which equates to the negative supply voltage of the Op Amp is usually added to reduce C_d. This is explained by the fact that a voltage over the photodiode creates an electrical field, which increases the speed of the photo-generated carriers. This, in turn, equates to a reduction of the capacitance.

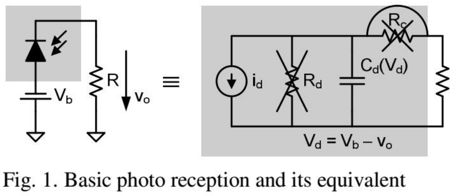

Fig. 1. Basic photo reception and its equivalent

Note also that the value of R_d is very large in comparison with the current-to-voltage converting resistor (R) and, therefore, can be ignored by removing it. Likewise, R_c is very small in comparison with R and can be ignored by shortening it. With R_d and R_s eliminated from the equivalent circuitry, it is observed that the photo current i_d experiences a low pass filter that is formed by C_d and R. Note that a reverse bias voltage V_b reduces the junction capacitance of the photodiode though that the voltage drop over R limits this reduction.

The transimpedance, which equates to the ratio of the output voltage v_o and the photocurrent i_d, is calculated by employing Kirchhoff's current law

$$i_d = \frac{v_o}{R} + \frac{v_o}{\frac{1}{j\omega C_d}} \tag{1}$$

$$\frac{v_o}{i_d} = \frac{R}{1 + j\omega RC_d} \tag{2}$$

Equating the real and imaginary parts of the transimpedance's pole, the achievable bandwidth B is calculated as

$$1 = \omega RC_d \tag{3}$$

$$B = \frac{\omega}{2\pi} = \frac{1}{2\pi RC_d} \tag{4}$$

Since the photocurrent is low, R needs to be sufficiently high to yield a sufficiently high output voltage. In practice, 100 kΩ is a typical resistor value, and a typical value for the junction capacitance of a large area photodiode is at least 50 pF. Substituting these values in (4) yields a bandwidth of a mere 318 Hz. A better current-to-voltage converter is the Transimpedance Amplifier (TIA), and Fig. 2 shows its configuration with an Op Amp. The virtual ground at the non-inverting input of the Op Amp shorts the capacitor, while the high impedance of the Op Amp's inverting input directs the photocurrent to the feedback resistance R_f. The current-to-voltage converting resistor R_f, therefore, is isolated from C_d and the output voltage v_o equals

$$v_o = i_d R_f \tag{5}$$

The bandwidth now depends on the bandwidth of the Op Amp, which is substantially higher than the bandwidth of a low pass filter formed by a resistor and the junction capacitance of a photodiode. Another advantage is that the TIA has low output impedance, which means that a wide variety of loads can be connected. Note also that, due to the virtual ground at the non-inverting input of the Op Amp, the negative bias voltage v_b is placed entirely over C_d, reducing its value. This increased reduction of C_d is important with regard to noise, which follows next.

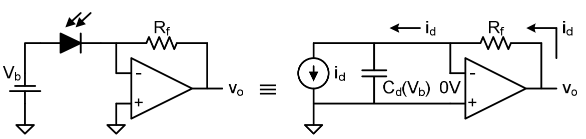

Fig. 2. Transimpedance photo reception its equivalent

III. Noise Analysis

The noise in a TIA is dominated by two types of noise, namely shot noise and thermal noise [7]. Shot noise is caused by the statistical nature of two physical phenomena, namely the emission of light, which is a random emission of discrete energy quanta called *photons*, and the recombination of electrons in semiconductor devices. Thermal noise, on the other hand, is caused by the random movement of electrons under the influence of heat. It is also known as *Johnson* noise or *Johnson-Nyquist* noise, after the two scientists who first researched the phenomenon. Shot noise and thermal noise are the only types of Gaussian noise [7]. The Gaussian character is due to their statistical nature. Within practical bandwidth-limited systems, Gaussian noise exhibits a constant density. In reference to white light, which contains all wavelengths of the visible spectrum, it is also known as *white noise*. Since the impact of Gaussian noise is frequency dependent, it is specified as a spectral density. The spectral density is either a noise voltage specified in $V/Hz^{1/2}$, or a noise current specified in $A/Hz^{1/2}$. Before addressing the influence of noise in the TIA as a whole, the noise in its individual components will be addressed. Sections III.A until III.C discuss the noise – and how it is modelled – of resistors, photodiodes and Op Amps, respectively. Section III.D incorporates the respective noise models in the TIA circuit and assesses its noise performance. Since noise increases with bandwidth, excessive bandwidth needs to be reduced. How to accomplish this is covered in Section III.E.

A. Noise in Resistors

Resistors are affected by thermal noise. The Root Mean Square (RMS) value of the voltage noise V_n – which equates to the standard deviation of its Gaussian distribution – is calculated as [7]

$$V_n = \sqrt{4kTRB} \tag{6}$$

where k is Boltzman's constant (1.38×10^{-23}), T the temperature in Kelvin, R the resistance in Ohms, and B the Bandwidth in Hertz.

Observe from (6) that, when the voltage over a resistor increases with R, its noise increases with $R^{1/2}$. Equation (5), on the other hand, shows that the signal increases with R, hence, the Signal-to-Noise Ratio (SNR) of a resistor increases with $R^{1/2}$. The RMS value of the noise current I_n is derived as

$$I_n = \frac{V_n}{R} = \sqrt{\frac{4kTB}{R}} \tag{7}$$

Since resistor noise can be specified in terms of voltage and current, there are two possible noise equivalents for a resistor. Both are displayed in Fig. 3. Since noise current only has influence when it flows through a resistor, and is converted into a noise voltage, the noise voltage equivalent is normally used. Noise current equivalents are used when they influence a signal current, e.g. the photocurrent of photodiodes.

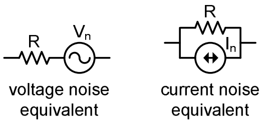

Fig. 3. Resistor noise equivalents

B. Noise in Photodiodes

Photodiodes are affected by shot noise and thermal noise. The thermal noise is generated by R_d and R_c. Their influences, however, are negligible and will be ignored. The shot noise is due to a statistical stream of photons being converted into a statistical stream of charge carriers (electrons). The RMS value of the shot noise current I_{sn} is calculated as [7]

$$I_{sn} = \sqrt{2qI_d B} \tag{8}$$

where q is the electron charge (1.602×10^{-19} C), I_d is the average photocurrent in Amperes and B the bandwidth in Hertz. The corresponding noise equivalent of a photodiode is displayed in Fig. 4.

C. Noise in Op Amps

Op Amps contain semiconductors and resistors. Hence, they are affected by thermal noise and shot noise. They are also affected by other types of noise, most notably $1/f$ or flicker noise, which is also known as pink noise. In wide band applications, however, these other types of noise are negligible compared to thermal and shot noise. For a detailed account of Op Amp noise sources, the reader is referred to Chapter 12 of [8]. Op Amp manufacturers use the noise equivalent of Fig. 5. This model, however, has two flaws. Firstly, the voltage noise is referred to the non-inverting input only and, secondly, the noise sources are assumed to be uncorrelated [9]. In spite of this shortcoming, this equivalent remains widely used for the simple reason that Op Amp datasheets do not specify more accurate noise parameters. As such, the equivalent of Fig. 5 will be used in the noise analyses of this work as well.

D. Noise in Transimpedance Amplifier

Replacing the components of the TIA of Fig. 2 with the noise equivalent circuits of Fig. 3, Fig. 4 and Fig. 5 yields the noise equivalent circuit of the TIA depicted in Fig. 6.

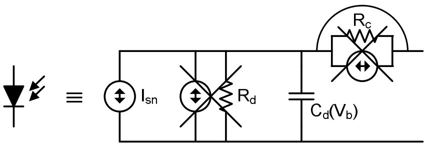

Fig. 4. Photodiode noise equivalent

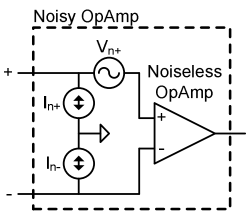

Fig. 5. Op Amp noise equivalent

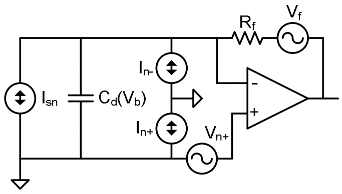

Fig. 6. TIA noise equivalent

The influence of the various noise sources is calculated by applying the principle of superposition. To assess the influence of an individual noise source:

- Replace the resistance of that noise source with a wire in case of a voltage noise source, or remove the resistance in case of a current noise source, and
- Replace the other noise sources with a wire in case of voltage noise sources, and remove the noise source in case of current noise sources.

From Fig. 6, observe that I_{n+} is shorted by the signal ground and, as such, has no influence. The noise current sources I_{nd} and I_{n-} are connected in parallel and can be considered jointly. Their influence is calculated from the equivalent circuit of Fig. 7, which is similar to the equivalent circuit of Fig. 2. As such, the influence of I_{sn} and I_{n-} on v_o is identical to the influence of the signal i_d. It is calculated as

$$\begin{aligned} v_o\left(I_{sn}, I_{n-}\right) &= \sqrt{\left(I_{sn}^2 + I_{n-}^2 \cdot B\right)} \cdot R_f \\ &= \sqrt{\left(2qI_d + I_{n-}^2\right)B} \cdot R_f \end{aligned} \tag{9}$$

where I_{n-} is specified in Op Amp data sheets. Note that I_{sn} is an RMS value whereas I_{n-} is a spectral density.

The influence of the noise voltage source V_f is equally straightforward. From Fig. 8, it is observed that, due to the virtual ground at the non-inverting input, its value is directly transferred to the output, i.e.

$$V_o\left(V_f\right) = V_f = \sqrt{4kTR_f B} \tag{10}$$

The noise voltage source V_{n+}, however, sees an integrator. This is shown in Fig. 9 from which the influence of V_{n+} on the output voltage v_o is calculated as

$$\sqrt{B} \cdot V_{n+} = \frac{V_o\left(V_{n+}\right)}{\frac{1}{j\omega C_d} + R_f} \cdot \frac{1}{j\omega C_d} \tag{11}$$

$$\frac{V_o\left(V_{n+}\right)}{\sqrt{B} \cdot V_{n+}} = 1 + j\omega R_f C_d \tag{12}$$

Note that V_{n+} in (11) and (12) above is a spectral noise density. Hence, it needs to be multiplied by $B^{1/2}$ to obtain a noise voltage.

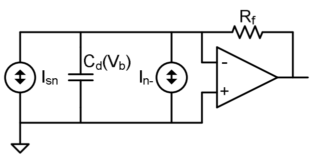

Fig. 7. Influence of input noise currents

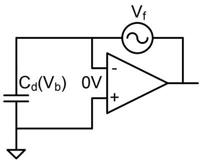

Fig. 8. Influence of feedback noise voltage

Equation (12) shows that $V_o(V_{n+})$ is frequency dependent. From the Bode plot of Fig. 9, it is observed that $V_o(V_{n+})/B^{1/2}V_{n+}$ increases with frequency until its plot intersects with the plot of the Op Amp's open loop gain $A_{OL}(f)$. From this point onwards, the rising slope of $V_o(V_{n+})/B^{1/2}V_{n+}$ is cancelled out by the falling slope of $A_{OL}(f)$. The frequency f_n determines the +3 dB point of the $V_o(V_{n+})/B^{1/2}V_{n+}$ plot, i.e. the point where $V_o(V_{n+}) = (2B)^{1/2}V_{n+}$. It is calculated by equating the real and imaginary parts of $V_o(V_{n+})/B^{1/2}V_{n+}$, i.e.

$$1 = \omega_n R_f C_d \quad (13)$$

$$f_n = \frac{\omega_n}{2\pi} = \frac{1}{2\pi R_f C_d} \quad (14)$$

If $R_f = 100\ \mathrm{k\Omega}$ and $C_d = 50$ pF, f_n equals 318 Hz, which indicates that noise starts increasing well below the bandwidth of wide and medium band receivers. Observe that, if C_d decreases, f_n increases. This, in turn, moves the plot to the right (indicated by the dashed line in the Bode plot of Fig. 9), and reduces the impact of V_{n+}. As mentioned in Section II, it is possible to reduce C_d by employing a negative bias voltage (V_b in Fig. 2).

E. Limitting bandwidth to Reduce Noise

The bandwidth of commercially available Op Amps is usually somewhat higher than the requirement of the target design. Since noise in a TIA increases with bandwidth, it is therefore necessary to limit the bandwidth of the TIA to the target specification in order to improve noise performance. This is accomplished by placing a capacitor C_f in parallel over the feedback resistance R_f (see Fig. 10).

The addition of C_f means that the signal current i_d sees the equivalent circuit of Fig. 11, which is a low pass filter that reduces the bandwidth to

$$B = \frac{1}{2\pi R_f C_f} \quad (15)$$

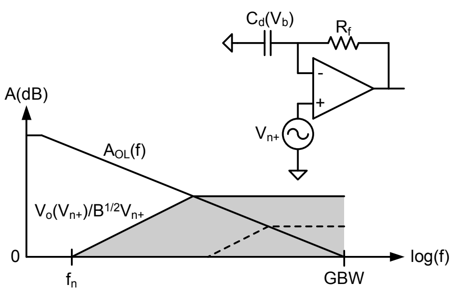

Fig. 9. Influence of input noise voltage

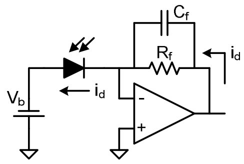

Fig. 10. Limitting the bandwidth of a TIA

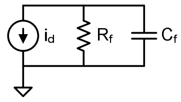

Fig. 11. Feedback low pass filter

IV. Choice of Op Amp

At present, two varieties of Op Amp exist with very different internal circuitry, namely Voltage Feedback (VFB) and Current Feedback (CFB) types. The former type has two high impedance inputs that are balanced, and employs voltage amplification. The latter type is a transimpedance amplifier and its inverting input is low impedance, which means that the inputs are not balanced. Another difference is that the feedback impedance of a CFB Op Amp must be a resistor that is fixed for optimum performance. A detailed description of the internal workings of the two types, which falls beyond the scope of this paper, is available in [8]. Most Op Amps are VFB, which is often referred to as the *standard type of Op Amp*. CFB types offer an increased slew rate and trade speed for precision. To the author's best knowledge, a CFB has not yet been considered for use in TIA designs for photo reception. Analysis presented in this work is under the assumption of a VFB type of Op Amp.

Since the input photocurrent is very low, it is important to increase the sensitivity by ensuring that the TIA produces little noise. As mentioned in Section III.C, noise in Op Amps is specified by both current and voltage noise sources. BJT low-noise Op Amps used to have a better voltage noise performance than their MOSFET counterparts, which have a superior current noise performance. Contemporary low noise MOSFET Op Amps, however, have noise voltages that are equal to these of low noise BJT Op Amps. Present day low noise devices reach levels below 10 nV/Hz$^{1/2}$. With regard to current noise, MOSFET types offer levels of 100s fs/Hz$^{1/2}$, whereas BJT types can reach only 100s ps/Hz$^{1/2}$. As such, MOSFET Op Amps offer a superior noise performance.

Finally, for stability reasons detailed in Section VI, it is important to select a unity gain stable Op Amp. This means that the transfer characteristic of $A_{OL}(f)$ needs to be first order, i.e. there may only be one pole before the $A_{OL}(f)$ curve crosses the 0 dB line. Following an in-depth argumentation, it is apparent that, for photodiode TIA designs, one needs to select a *low noise, unity gain stable MOSFET VFB* type of Op Amp.

V. Influence of Op Amp Input Impedances

Up to this point, the presented analysis has ignored the input impedances of the Op Amp. An Op Amp has two inputs, and there are three input impedances. The first two impedances are the so-called *common mode* impedances, which are the impedances between the respective inputs and the signal ground. Since VFB Op Amps have balanced inputs, the common mode impedances have equal values. For this reason, data sheets usually specify the common mode impedance as a single value. The third impedance is the *differential* impedance, which is situated between the two inputs. The input impedances of Op Amps consist of a resistor in parallel with a capacitor. In data sheets, they are usually specified as such. The input resistances of low-noise MOSFET VFB Op Amps are extremely high (terra Ohms) and can be ignored. The input capacitances, however, are pico farads and cannot be ignored.

Fig. 12 shows an Op Amp with its input capacitances. The common mode capacitances at non-inverting and inverting inputs are denoted as C_{cm+} and C_{cm-}, respectively. Note that these can also include stray capacitances from the Printed Circuit Board (PCB) and surrounding components. The differential capacitance is symbolized by C_{diff}.

Let's now incorporate Fig. 12 in previous equivalent circuits and examine the influence of the Op Amp's input capacitances on the signal and noise of the TIA of Fig. 2. Since the non-inverting input of the Op Amp in the TIA is connected to ground, C_{cm+} is shorted and has no influence. Replacing the Op Amp in the equivalent circuit of Fig. 2 with Fig. 12 yields Fig. 13. From this figure it is observed that, with regard to the signal i_d, C_{cm-} and C_{diff} are placed in parallel with the photodiode's junction capacitance C_d. A similar deduction is made for the noise currents I_{sn} and I_{n-}. Since, through the virtual ground at the inverting input of the Op Amp, these capacitances are shorted, they do not influence the impact of I_{sn} and I_{n-}. Likewise, the influence of V_f does not change. Therefore, (9) and (10) remain valid.

Replacing the Op Amp in the equivalent circuit of Fig. 9 with Fig. 12 yields Fig. 14, from which the effect of C_{cm-} and C_{diff} on the influence of V_{n+} can be determined.

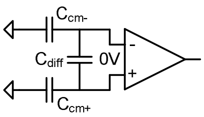

Fig. 12. Input capacitances of an Op Amp

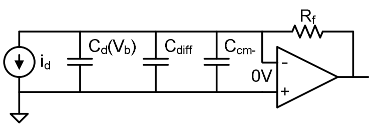

Fig. 13. Influence of input capacitances on the signal

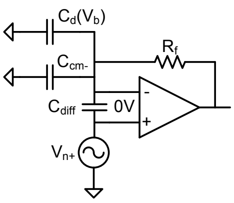

Fig. 14. Influence of input capacitances on the input noise voltage

Observe from Fig. 14 that, since the inputs of the Op Amp have the same potential, C_{diff} is shorted and has no influence. The capacitance C_{cm-} (and the stray capacitances that surround it) is parallel to C_d and, as such, contributes to the amplification at higher frequencies. In analytical terms, this translates to a term change in (12) and (14) as C_d needs to be replaced by $C_d + C_{m+}$. The equations for $V_o(V_{n+})$ and f_n, hence, become

$$V_o\left(V_{n+}\right) = \left(1 + j\omega R_f\left(C_d + C_{m+}\right)\right)\cdot\sqrt{B}\cdot V_{n+} \quad (16)$$

$$f_n = 1/2\pi R_f\left(C_d + C_{m+}\right) \quad (17)$$

VI. INFLUENCE OF OP AMP ERROR VOLTAGE

Finite open loop gain leaves a residual signal voltage between the inputs of a practical Op Amp. Albeit small, the value of this error voltage v_e does not equal zero. This section examines the influence of this non-ideal property.

Consider Fig. 13, which is the equivalent circuit of the TIA without bandwidth limiting feedback capacitor C_f. Note that the inclusion of C_f will be considered later in this section. Simplifying this circuit by considering a single capacitance C_{in} that equals

$$C_{in} = C_d // C_{diff} // C_{cm-} = C_d + C_{diff} + C_{cm-} \quad (18)$$

yields Fig. 15.a. Since $v_e \neq 0$, there is current flowing through C_{in} which drains part of the signal current i_d away from R_f. This, of course, reduces the bandwidth. A lower capacitance, hence, would not only reduce noise. It will also preserve bandwidth. Moreover, the current feedback path is no longer resistive. In addition to R_f, it now includes a capacitance that is placed in series with R_f. At the non-inverting input, which is at v_e, the feedback path forms a low pass filter. This is illustrated in Fig. 15.b. Since the Op Amp is unity gain stable, it can have a phase shift of up to 90 degrees. The low pass filter feedback path could add another 90 degrees, making the design less stable. Since v_e is very small, it has no significant influence on the generated noise, hence (9), (10), (16) and (17) remain valid.

Taking a more analytical approach, and including C_f again, consider Fig. 15.c which shows what is fed back from the output of the Op Amp to its input. Hence, v_e is calculated as

$$v_e = \frac{v_o}{\left(R_f // \frac{1}{j\omega C_f}\right) + \frac{1}{j\omega C_{in}}}\cdot\frac{1}{j\omega C_{in}} \quad (19)$$

After some mathematical manipulation, the so-called feedback factor β is calculated as

$$\beta = \frac{v_e}{v_o} = \frac{j\omega R_f C_f + 1}{j\omega R_f\left(C_{in} + C_f\right) + 1} \quad (20)$$

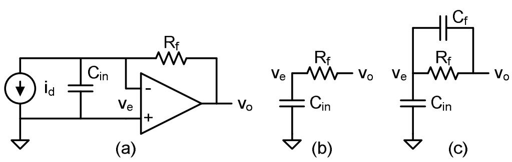

Fig. 15. Influence of Op Amp error voltage

The reciprocal of (20), i.e. 1/β, and its plot in Fig. 16 show that stability issues arise by the presence of a zero at f_z, and that these are countered by the presence of a pole at f_p. Note that setting f_p at the intersect of 1/β with $A_{OL}(f)$ reduces the feedback phase shift by 45°, which is commonly considered as good compromise between stability and achievable bandwidth.

A widely used parameter in the characterization of Op Amps is the Gain Bandwidth Product *GBW*, which is defined as

$$GBW = A_{OL}\left(f\right)\cdot f \ . \tag{21}$$

Since $A_{OL}(f)$ is a dimensionless ratio, *GBW* is in fact a frequency, more specifically the frequency where $A_{OL}(f)$ equals unity, i.e. 0 dB. From Fig. 16, observe that, before crossing the 0 dB axis, the $A_{OL}(f)$ curve should have a single pole to ensure stability. This is due to the fact that additional poles introduce extra 90° shifts. As mentioned previously, devices that exhibit such single pole curve are called *unity gain stable Op Amps*.

From AC circuit theory, it is observed that f_p is the frequency where $1/\omega C_f$ equals R_f. From this equality, f_p is calculated as

$$f_p = \frac{1}{2\pi R_f C_f} \ , \tag{22}$$

which, from (20), shows that

$$\frac{1}{\beta\left(f >> f_p\right)} = \frac{C_f + C_{in}}{C_f} \ . \tag{23}$$

Taking the asymptotical approximation of Fig. 16, observe that $A_{OL}(f_p) \approx 1/\beta(f_p)$. Hence, from (21) and (23)

$$\frac{GBW}{f_p} \approx \frac{C_f + C_{in}}{C_f} \ . \tag{24}$$

Substitution of (22) in (24) yields

$$2\pi R_f C_f^2 GBW - C_f - C_{in} = 0 \ , \tag{25}$$

which, with regard to C_f, is a quadratic equation with one real positive root, namely

$$C_f = \frac{1}{2\pi R_f GBW}\left(1 + \sqrt{1 + 8\pi R_f C_{in} GBW}\right) . \tag{26}$$

Assuming that $8\pi R_f C_{in} GBW >> 1$, (26) simplifies to

$$C_f = \sqrt{\frac{C_{in}}{2\pi R_f GBW}} \tag{27}$$

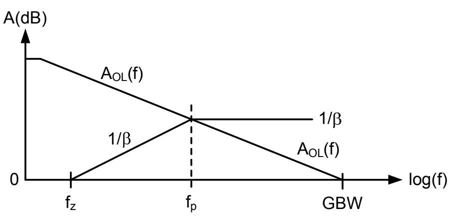

Fig. 16. Bode plots of stabilized TIA

Note that two equations have been derived to calculate C_f, namely (15) and (27). The equation that yields the highest value should be adopted. Equation (27) is often quoted in design guides and the application section of high speed Op Amps. To illustrate its validity for wide band TIAs, let's assume that the selected Op Amp has a *GBW* of 10 MHz, that $R_f = 100$ kΩ and that $C_{in} = 50$ pF. In that case, the term $8\pi R_f C_{in} GBW$ equals 1257 which, indeed, is much greater than one. Substituting (27) in (22), the attainable bandwidth *B* of the design with compensating feedback capacitance C_f is calculated as

$$B = f_p = \sqrt{\frac{GBW}{2\pi R_f C_{in}}} \ . \tag{28}$$

This result indicates that the resulting bandwidth is inversely proportional to the square root of the current-to-voltage converting feedback resistor R_f. If bandwidth is a critical requirement, the best approach may be to limit the gain of TIA and to follow it by a broadband voltage amplifier.

VII. Summary

Through a step-by-step approach, a detailed account of the design and analysis of a TIA photo receiver, built around an Op Amp, has been provided. It was deducted that low noise, unity gain stable MOSFET VFB types of Op Amp should be used. Formulae to assess noise performance were also derived. A noteworthy observation was that the junction capacitance of the photodiode, as well as input capacitances of the Op Amp, cause amplification of the Op Amp's noise voltage, and provide for a feedback path that causes instability. An effective measure of guaranteeing stability is to include a feedback capacitance that also limits the noise generated by the TIA. Formulae to calculate the value of this capacitance were also obtained.

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