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Optimal Design of High Speed Analog APD Receivers

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I. INTRODUCTION

In order to provide optical to electrical conversion of high speed analog signals, avalanche photodiode receivers must be optimized for several parameters. Among these parameters are receiver amplifier noise, APD noise, and bandwidth.

Optimization of these parameters requires a theoretical understanding of each of the parameters and their interactions. In the first part of this paper the theoretical design of an avalanche photodiode optical receiver will be discussed with attention to the parameters mentioned. Later, a comparison between calculated and measured values for a receiver design will be presented.

II. RECEIVER AMPLIFIER

Applications employing analog transmission of information typically place a premium on the dynamic range of the receiver. For this reason, the high impedance amplifier design, with its high input resistance and subsequently limited dynamic range, will not be considered in this analysis. Instead the transimpedance amplifier design shown in figure 1 will be the subject of the analysis.

Figure 1. The transimpedance receiver.

As will be discussed later, the sensitivity of a transimpedance receiver is increased by increasing the value of the feedback resistance. The feedback resistance, R_f , can be made as large as possible for a given bandwidth by maximizing the open loop gain of the amplifier. The open loop gain of the amplifier is limited by the phase shift due to the amplifier. That is to say that the feedback ceases to be negative and the amplifier becomes unstable. This instability can

manifest itself as pulse distortion such as overshoot and ringing.

In order to achieve the maximum open loop gain with a minimum number of stages, thereby reducing the phase shift in the amplifier, most current high gain, high speed analog receivers employ microwave bipolar transistors as a gain element. The useful transconductance of present day microwave bipolar transistors is from 6 to 10 times that of commercially available FET's. Various topologies using microwave BJT's have been reported ^{1,2}. A typical high speed analog transimpedance receiver is shown in figure 2.

Figure 2. A typical BJT transimpedance receiver.

III. NOISE DUE TO THE RECEIVER AMPLIFIER

The variance of the input equivalent noise of a bipolar transistor transimpedance amplifier is given by ^{3,4,5}

$$\begin{aligned} \langle i_n^2 \rangle = & 4kTB/R_f + 2qI_bB + 4kTr_{bb'}(2\pi C_p)^2B^3 \\ & + (2qI_c/g_m^2)(2\pi C_T)^2B^3 \end{aligned} \quad (1)$$

where

I_b is the base bias current,
 I_c is the collector bias current,
 R_f is the feedback resistance,
 B is the noise bandwidth,
 $r_{bb'}$ is the base spreading resistance,
 k is Boltzmann constant,
 T is the absolute temperature,
 q is the charge on an electron, and
 g_m is the transconductance of the BJT, given by

$$g_m = I_c/V_t$$

where $V_t = kT/q$.

The capacitance C_T is given by

$$C_T = C_{b'e} + C_{b'c} + C_p \quad (2)$$

where C_p is the sum of the detector and feedback resistor parasitic capacitance. The capacitances $C_{b'e}$ and $C_{b'c}$ are the small-signal hybrid- π -model capacitances of the transistor.

The capacitance $C_{b'e}$ is dependent on the collector current of the transistor due to the fact that the total emitter capacitance is the sum of a space-charge capacitance, C_{je} , and a diffusion capacitance term.

$$C_{b'e} = C_{je} + I_C / 2\pi V_t f_t \quad (3)$$

where f_t is the short circuit gain-bandwidth product.

The total input capacitance can now be expressed as the sum of a zero bias condition term and the bias dependent emitter diffusion capacitance term, such that

$$C_T = C_0 + I_C / 2\pi V_t f_t \quad (4)$$

where

$$C_0 = C_{b'c} + C_{je} + C_p \quad (5)$$

Substituting equation (4) into equation (1) and setting the derivative of $\langle i_n^2 \rangle$ with respect to I_C to zero, an optimum collector bias current can be found such that ⁴

$$I_C = 2\pi C_0 f_t V_t (1 + f_t^2 / \beta B^2)^{-1/2} \quad (6)$$

where β is the DC forward current gain of the transistor.

Each term of equation (1) denotes one of the principal noise sources in the BJT transimpedance amplifier. In order of their occurrence in the equation these noise sources are thermal noise due to the feedback resistor, base current shot noise, thermal noise due to the base spreading resistance, and collector current shot noise.

It can be seen that for high speed applications the noise contributions of the base spreading resistance, the collector shot noise, and associated capacitances dominate. Therefore, in order to minimize these noise effects, a device should be chosen with the highest possible f_t achievable at the lowest possible I_C . The device should also have the lowest possible $r_{bb'}$. This is not easily achieved since the small geometries required to produce high f_t devices often results in high values of $r_{bb'}$, particularly in monolithically integrated amplifiers. In addition, all of the terms of C_0 (5) must be minimized.

Some parameters of currently available microwave Si bipolar transistors are

$$\begin{aligned}
\beta &= 100 \\
f_t &= 10 \text{ GHz} \\
r_{bb'} &= 20 \, \Omega \\
C_{b'c} &= 0.2 \text{ pF} \\
C_{je} &= 0.8 \text{ pF} .
\end{aligned}$$

It should be noted that although increasing the value of R_f decreases the thermal noise contribution of the feedback resistance, it also decreases the maximum allowable input level for linear operation of the amplifier, thereby limiting the dynamic range of the receiver. Increased values of R_f also reduce the bandwidth of the receiver as will be discussed later.

IV. NOISE DUE TO THE APD

The photocurrent gain of an avalanche photodiode is created by impact ionization or ionizing collisions with the lattice in a region of high electric field strength in the p-n junction of the APD. These collisions produce an avalanche of exciton pairs thereby producing the gain. This gain process is probabilistic in that probability that a carrier will have an ionizing collision is a strong function of the electric field strength.

The gain or multiplication factor, $\langle M \rangle$, of the APD is the average gain of the total injected photocurrent. Due to the probabilistic nature of the gain there is a multiplication noise given by the gain distribution of all photogenerated carriers.

Given this distribution of gains, then

$$\langle M^2 \rangle > \langle M \rangle^2 \quad (7)$$

This excess noise factor, F , is then given by

$$F(M) = \langle M^2 \rangle / \langle M \rangle^2 . \quad (8)$$

The variance of the noise current of the APD can be expressed as ⁶

$$\langle i_{napd}^2 \rangle = 2q(I_{ds} + (I_{db}M^2 + P_0RM^2)F)B \quad (9)$$

where

P_0 is the incident optical power,
 M is the gain of the APD, and
 R is the responsivity of the APD.

There are two components of the dark current in an APD. The surface dark current, I_{ds} , is not multiplied by the avalanche process, but the bulk dark current, I_{db} , is multiplied by the gain of the photodiode.

Expressions for the excess noise factor have been derived for both electron and hole injection ⁶. For electron injection the excess noise

factor is given by

$$F = kM + (1-k)(2-(1/M)) \quad (10)$$

where k is ratio of the ionization coefficients of holes to electrons. For hole injection k would be replaced by $1/k$.

When carrier ionization rates are markedly different as in Si APD's, the result is low excess noise. However, when the carriers in a material have comparable ionization rates as is the case in Ge or III-V compounds, then the gain process is noisier. A typical value of k for Si is 0.02, while Ge is 0.7 to 1.0 and InGaAs APD's exhibit k values of 0.3 to 0.5. These high excess noise factors for the long wavelength APD's limit their usefulness in wide dynamic range analog receivers.

It is important to notice that from (9) the noise generated by an APD is a function of the input power. So that for a given minimum signal to noise ratio there is an optimum APD gain.

It should also be noted that this treatment of the gain induced noise of the APD assumes that the gain of the APD does not vary as a function of position on the surface of the APD. This is not always a valid assumption, and care must be taken to mitigate any position sensitive gain effects. In general this none uniform gain distribution as a function of position is most pronounced at higher APD gains.

V. TOTAL RECEIVER NOISE

Given the receiver noise contributions expressed in (1) and (9), an expression can be written for the total noise equivalent power at the input of the receiver, such that

$$P_{ntot} = (P_{napd}^2 + P_{namp}^2)^{1/2} \quad (11)$$

Substituting (1) and (9) into (11), and solving in terms of the signal to noise ratio, Q , gives

$$P_{ntot} = [(\langle i_n \rangle Q / (RM))^2 + (2qBFQ^2/R)^2]^{1/2} \quad (12)$$

From this expression the sensitivity of the receiver as a function of the minimum usable signal to noise ratio can be determined with respect to the other design parameters such as APD gain and feedback resistance.

VI. BANDWIDTH

The bandwidth of the receiver is the convolution of the bandwidth of the amplifier and the bandwidth of the APD. The bandwidth of the amplifier is approximated by

$$f_{-3dB \text{ amp}} \approx 1/(2\pi R_{ct} C_{ct}) \quad (13)$$

where R_{ct} is the total resistance at the collector of the first transistor, and C_{ct} is the total capacitance at the same point.

In the gain bandwidth limited case the bandwidth of the APD can be expressed as ⁷

$$f_{-3dB} \text{ apd} = 1/(2\pi M\tau) \quad (14)$$

where τ is the effective carrier transit time of the APD.

VII. A DESIGN EXAMPLE

Using the expressions presented in this paper, an analog APD receiver with the following specifications was designed.

f_{-3dB}	500 MHz
t_r	<700 ps
gain	>10 ⁴ volts/watt

In addition, the receiver was to be optimized for maximum dynamic range with a minimum usable signal to noise ratio of 20 dB.

The receiver used an RCA type C30902 APD and microwave BJT's as described previously.

Applying equation (6), a optimum bias condition for the amplifier was found. In this case value of C_0 was found to be 6.9 pF, and the resulting bias condition was $I_C = 5$ mA.

This condition was used in the solution of the combined noise expression (12) with $Q=10$. The results of that solution as a function of APD gain and feedback resistance is shown in figure 3.

Figure 3. Noise equivalent power as a function of APD gain at a signal to noise ratio of 20 dB.

As can be seen from the curve, the optimum APD gain is defined by the broad region around $M=50$ and is essentially independent of the value of R_f . Since the design specification only called for a 10 kV/W transfer characteristic, a lower than optimum gain of $M=30$ was chosen in order to produce a larger linear dynamic range from the APD.

The feedback resistance was chosen to be $1k\Omega$ by considering the effect of R_{ct} on the bandwidth (13).

Good agreement between the calculated and measured performance was achieved. A comparison of calculated and measured values for the receiver is presented in table 1.

	calculated	measured
Output noise with zero signal input	230 μV	210 $\pm$ 30 μV
Output noise with 10 μW input power	800 μV	800 $\pm$ 100 μV
Rise time (from bandwidth)	680 ps	700 $\pm$ 50 ps

Table 1.

VIII. CONCLUSIONS

Several theoretical expressions have been presented that allow the designer to optimize the performance of APD based transimpedance optical receivers. It should be noted that there is no one optimum set of design conditions for analog APD receivers. For example, a receiver with maximum sensitivity may not provide the maximum dynamic range. Yet, with the tools presented here the designer can better understand the interplay of design parameters necessary to achieve the design goal.

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