# Effective Amplifier Noise for an Optical Receiver Based on Linear Mode Avalanche Photodiodes

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*The rms noise charge induced by the amplifier for an optical receiver based on the linear-mode avalanche photodiode (APD) was analyzed. It is shown that for an amplifier with a 1-pF capacitor and a noise temperature of 100 K, the rms noise charge due to the amplifier is about 300. Since the noise charge must be small compared to the signal gain, APD gains on the order of 1000 will be required to operate the receiver in the linear mode.*

## I. Introduction

The use of silicon avalanche photodiodes (APDs) in photon detection has been reported by several authors [1-4]. Compared to photomultiplier tubes (PMTs), the APDs offer smaller size and higher quantum efficiency. Since it is a solid state device, the APD is also inherently more reliable than the PMT. Additionally, the APD can be operated at a moderate supply voltage.

In order to apply the APD to photon-counting operation, the APD can be operated in either the Geiger mode [1-4] or the linear mode [5]. Most of the research in APD-based photon counters has been carried out with Geiger-mode APDs. In the Geiger-mode operation, the detector is biased beyond the breakdown voltage and cooled to reduce the number of thermally excited carriers. Since the device is biased beyond its normal breakdown voltage, an easily observable avalanche event will be initiated when a photon is absorbed by the device. Detector gain on the order of $10^5$–$10^6$ can be achieved. The principal disadvantage of the Geiger-mode operation is the requirement to quench the detector after each photon event to stop the avalanche. Both passive [1, 2] and active [3, 4] quenching schemes have been investigated. Since the quenching operation requires a direct feedback from the detector, the operating bandwidth is limited. Device speeds on the order of tens of megacounts per second have been reported. Another disadvantage associated with the Geiger mode operation is the large amount of after-pulsing observed. This is caused by the release of carriers that are trapped by the shallow potential wells during the avalanche. Since the receiver cannot differentiate between a photon event and after-pulsing, the after-pulsing effect must be minimized for reliable photon-counting operation.

An alternative to operating the APD in the Geiger mode is to operate the device in the linear mode, under normal bias condition. Under this mode, the detector is biased at just below the breakdown voltage so that the detection gain is large but finite. Since the detector is biased at below the breakdown voltage, the avalanche action does not persist and

the diode recovers quickly for the next photon event. Further-more, since the number of carriers flowing through the junction is relatively small compared to the Geiger mode, the effect of after-pulsing is practically negligible. The low detection gain, however, poses a significant problem for the post-detection amplifier design. The amplifier must be designed to have an equivalent noise that is small compared to the total charge developed by the APD. The purpose of this article is to perform the noise analysis of such a detector.

## II. Simplified Noise Model

The APD can be modeled as a current source which outputs $G$ electrons for each detected photon. The avalanche gain, $G$, is random and its statistics depend on the bias voltage. An amplifier is needed to convert the current signal from the APD into a voltage signal. The two amplifier configurations that are most commonly used are the trans-impedance or shunt-feedback configuration, and the high impedance or integrating configuration. The effects of amplifier noise in these configurations are considered separately.

### A. Shunt Feedback Configuration

The block diagram for a shunt-feedback receiver circuit is shown in Fig. 1. The output current (charge) from the APD is integrated by the amplifier, which consists of an op-amp and a feedback capacitor. The reset switch with a low resistive load is included so that the charge accumulated by the capacitor can be quickly discharged, thereby improving the response time of the circuit.

The photodiode can be modeled as an ideal current source with a shunt resistor $R_D$, a shunt capacitor $C_D$, and a series resistance $R_s$. The shot noise and the excess noise of the diode are summarized by the current source $I_s$. The amplifier can be characterized by an input noise voltage $v_n$, an input noise current $i_n$, an input impedance $Z_i$, and an effective gain $A$. Both the amplifier gain and the input impedance are assumed to be large. Because of the large input impedance of the amplifier, charge flowing from the detector will tend to accumulate over the feedback capacitor $C_F$, resulting in a voltage signal at the inverting input given by

$$v_- - v_o = \frac{Q_F}{C_F} \qquad (1)$$

where $Q_F$ is the amount of charge accumulated across the feedback capacitor, and is related to the amount of charge accumulating over the diode capacitance, $Q_D$, by

$$(A + 1) \times \frac{Q_D}{C_D} = \frac{Q_F}{C_F} \qquad (2)$$

In order to achieve effective charge transfer into the integrating capacitor, the capacitance of the diode must be small compared to the effective feedback capacitance, $(A + 1) C_F$. Current state-of-the-art APDs have capacitance on the order of 1–2 pF. Future devices may have capacitance less than 1 pF. The total charge generated by the APD, or $Q$, is simply the sum of $Q_F$ and $Q_D$

$$Q = Q_F + Q_D \qquad (3)$$

For a sufficiently large $A$, it is reasonable to assume that most of the charge developed across the diode is accumulated across the feedback capacitor. In other words, the signal charge accumulated over the integrating capacitor is equal to the avalanche gain, $G$, of the detector.

The effect of noise on the integrating amplifier can be seen from the noise-equivalent circuit shown in Fig. 2. The rms voltage noise at the input to the amplifier, $v_n$, will induce an effective charge over the capacitor, $Q_{n,v}$. For an integrating amplifier with integration period $T_s$, the total noise charge is given by

$$Q_{n,v} = C_F v_n = C_F v_n / \sqrt{T_s} \qquad (4)$$

Note that the amplifier noise voltage, $v_n$, which is generally specified in volts per root hertz, should be multiplied by the root bandwidth $(1 / \sqrt{T_s})$ to provide the rms input voltage.

The effect of current noise can be similarly analyzed. The shunt resistance of the diode, $R_D$, is typically on the order of several hundred megohms to a few gigohms. Because of its large value, the thermal noise generated by the diode can be practically ignored, and the dominating source of current noise is the amplifier noise current $i_n$. In general, since the noise current is shunted by $C_D$ and $R_D$, the effective current noise flowing into $C_F$ will be smaller than $i_n$. However, since the feedback capacitor is much larger than the diode capacitance for effective charge transfer, most of the noise current will flow across $C_F$. As a result, the effect of diode shunt is small and will be ignored in subsequent calculations. For a white current noise with power spectral density $i_n^2$, the rms charge accumulated during the integrating period can be calculated as

$$Q_{n,i} = \sqrt{\langle \iint i_n(t) i_n(t)\, dt\, dt \rangle} = \sqrt{i_n^2 T_s} \qquad (5)$$

where the notation $\langle x \rangle$ denotes the expectation of $x$. By assuming that the current noise and the voltage noise are

uncorrelated, the total rms noise charge sensed by the integrating amplifier is therefore

$$Q_n = \sqrt{\frac{(C_F v_n)^2}{T_s} + i_n^2 T_s} \qquad (6)$$

Note that the effective noise charge depends on the noise current and noise voltage of the amplifier. For an amplifier operating at an equivalent noise temperature $T_n$, the noise charge can be optimized by choosing the noise impedance, $R_n$, equal to

$$R_n = \frac{v_n}{i_n} = \frac{T_s}{C_F} \qquad (7)$$

The resulting optimal amplifier noise charge is given by

$$Q_n = C_F v_n \sqrt{\frac{2}{T_s}} = \sqrt{2kT_n C_F} \qquad (8)$$

where we have used the fact that

$$v_n i_n = kT_n \qquad (9)$$

In an optical communication receiver using an APD, the detector output should be integrated over the slot time. With a 10-nsec slot time, a 1-pF feedback capacitor, and an amplifier noise temperature of 100 K, the optimum effective noise charge as seen by the amplifier input is equal to 330 electrons. In order to effectively distinguish the detector output from system noise, the total signal charge must be several times the noise charge. In other words, the APD must provide a gain on the order of 1000. In practice, the noise voltage and noise current of an amplifier will not be optimized, and an even larger detector gain will be required.

## B. High Impedance Configuration

The block diagram of the high impedance integrating amplifier is shown in Fig. 3. The configuration is simpler than the shunt feedback amplifier in that no feedback impedance is employed to develop the voltage signal. Instead, a high input impedance, $Z_i$, is used to convert the current signal from the APD into a voltage signal. Since the amplifier does not rely on the closed-loop feedback, the frequency response for the high Z configuration is generally better than that of the shunt feedback configuration.

The noise analysis for this configuration is straightforward. Again, the effect of diode series resistance $R_s$ shall be ignored. The effect of amplifier noises, $i_n$ and $v_n$, can be analyzed in a similar way to the shunt feedback case. The results are

$$Q_{n,v} = C_D v_n = C_D v_n / \sqrt{T_s} \qquad (10)$$

$$Q_{n,i} = \sqrt{i_n^2 T_s} \qquad (11)$$

Note that for the high impedance configuration, the device capacitance $C_D$ dominates the voltage noise contribution. In addition to the amplifier current and voltage noises, the thermal noise of the input impedance $Z_i$, is also important. For a resistive load of $R_i$, the rms noise charge at the diode capacitor due to the thermal noise can be calculated by substituting into Eq. (5) the spectral density of the thermal noise, $i_T^2 = 4kT/R_i$. The result is

$$Q_{n,T} = \sqrt{\frac{4kT}{R_i} T_s} \qquad (12)$$

where $T$ is the operating temperature of the amplifier. Note that $T$ is in general different from the noise equivalent temperature $T_n$. The total rms noise charge at the amplifier input is the root mean square sum of $Q_{n,i}$, $Q_{n,v}$, and $Q_{n,T}$. For an input impedance of 100 kilohms and an operating temperature of 300 K, the equivalent noise charge is about 250. This number can be reduced significantly, of course, by increasing the amplifier input impedance $R_i$. For a 10-megohm input impedance, the noise count due to $R_i$ is only about 30. By choosing the high input impedance, therefore, the effect of thermal noise can be ignored compared to the effects of amplifier current and voltage noises, and the optimal rms noise charge can again be approximated by Eq. (8).

## III. Limitations

Practical limitations on the application of linear-mode APDs include the following: non-ideal amplifier noise impedance, gain-bandwidth limitation of the amplifier, and stray (parasitic) capacitance and dynamic resistance.

The effect of amplifier noise impedance can be seen as follows: For a system operating with a 1-pF capacitance and 10-nsec slot time, the optimal noise impedance is 10 kilohms. With a 10:1 mismatch, the equivalent noise charge is about 740 electrons rms. This is large compared to the 300 electrons rms of a matched amplifier. The gain-bandwidth product is another factor that limits the application of linear mode APDs. For the shunt feedback configuration, the amplifier gain must be high enough to ensure effective charge transfer from the diode, and yet must be low enough to provide a reasonable bandwidth. In order to achieve an effective integration period of 10-nsec, a closed-loop bandwidth in excess of 1 GHz will be required. The high input impedance configuration, on the

other hand, does not rely on the active feedback to achieve high trans-impedance gain. Consequently, it is easier to achieve the desired frequency response with the high impedance configuration. Finally, the effect of stray capacitance is also important. Excessive capacitance at the input of the amplifier can reduce the charge transfer and increase the effective noise charge. And any stray capacitance at the output can reduce the frequency response of the circuit.

## IV. Conclusions

Operating the APD under linear mode removes the need for quenching the detector after each photon event. The large, but finite, charge gain resulting from the avalanche can be integrated by either a shunt-feedback or an integrating amplifier to generate a detectable signal at the output. The rms noise charge introduced by the amplifier was analyzed. It is shown that for a receiver operating with a 1-pF capacitance and an equivalent amplifier noise temperature of 100 K, APD gain in excess of 300 is required to provide a detectable signal (SNR = 1). The required detection gain can be even higher if the amplifier noise impedance is not perfectly matched. In order to provide a detection threshold that is several times higher than the noise level, a detection gain on the order of 1000 will be required for a matched system. It should be noted, however, that this analysis was carried out only for the two basic amplifier configurations shown in Figs. 1 and 3. Further reduction in amplifier noise at the same noise temperature is possible with other circuit configurations, and more studies are needed to fully understand the lower limit on the amplifier noise.

## References

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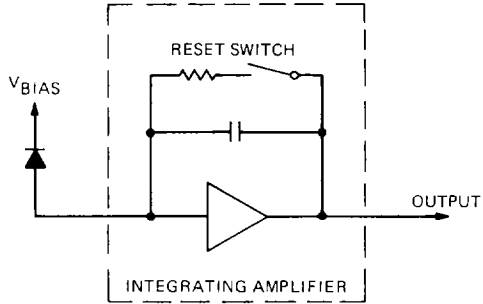
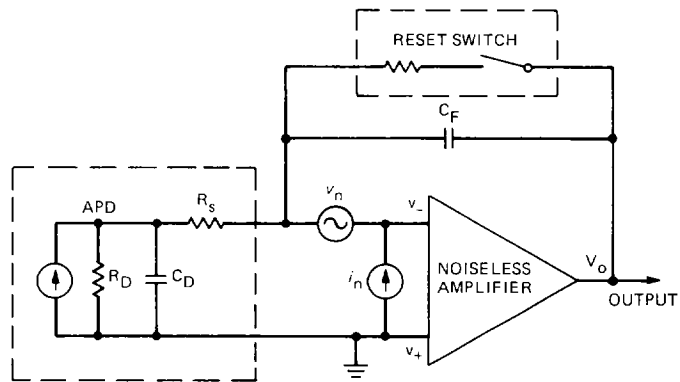**Fig. 1. Configuration of a shunt-feedback integrating amplifier.**



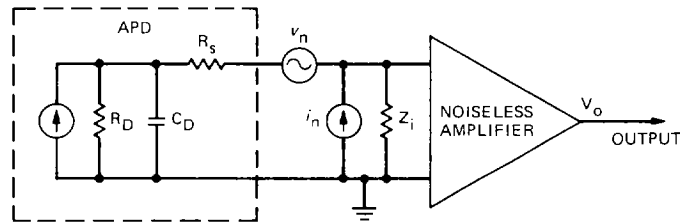**Fig. 2. Noise-equivalent circuit of the integrating amplifier.**



**Fig. 3. Noise-equivalent circuit of high impedance trans-impedance amplifier.**

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