A transimpedance amplifier for excess noise measurements of high junction capacitance avalanche photodiodes

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Abstract. This paper reports a novel and versatile system for measuring excess noise and multiplication in avalanche photodiodes (APDs), using a bipolar junction transistor based transimpedance amplifier (TIA) front-end and based on phase-sensitive detection, which permits accurate measurement in the presence of a high dark current. The system can reliably measure the excess noise factor of devices with capacitance up to 5 nF. This system has been used to measure thin, large area Si pin APDs and the resulting data is in good agreement with measurements of the same devices obtained from a different noise measurement system which will be reported separately.


Keywords: Avalanche Photo-diode, Excess Noise, Noise Measurement, High Capacitance.
1. Introduction

Avalanche photodiodes (APDs) exhibit an internal gain mechanism whereby secondary carriers are generated by impact ionization [1]. Impact ionization is often exploited to enhance the signal-to-noise ratio of electro-optical systems in communications, medical and military applications [2–4]. A popular model that describes avalanche multiplication was proposed by McIntyre [5]. The electron ($\alpha$) and hole ($\beta$) ionization coefficients are usually reported as a function of electric field. $\alpha$ and $\beta$ are often extracted from measurements of the photo-multiplication and excess noise factor of a particular APD structure, fabricated from a certain material; for example in [6–8]. The ratio $k_{\text{eff}} = \frac{\alpha}{\beta}$ may be used as a relative figure of merit when comparing two or more competing material systems for use with a given device intrinsic width.

Several measurement systems have been reported which evaluate the excess noise associated with impact ionization mechanism [9–12]. Each of these systems has some limitations with respect to the measurement bandwidth, the minimum detectable photogenerated noise, and the maximum permissible device junction capacitance. The relative merits of each system are discussed in section 15.

The measurement system and front end reported herein enable measurements of photo-multiplication and excess noise on devices with junction capacitance, $c_{\text{j}}$, up to 5 nF. For the first time, HF region excess noise measurements on thin, large area devices including forward biased devices and photo-voltaic cells are possible. It has been suggested that the noise magnitude of a solar cell is proportional to the density of material defects [13]. The defect density is linked to cell quality. Noise measurements may be used to grade solar cells at the time of production [14].

2. Front-End Design Methodologies

Ignoring distributed techniques, amplifiers that interface electronic systems with electro-optical detectors can be split into three groups based on their input impedance [15].

- High impedance amplifier
- Impedance matched amplifier
- Transimpedance amplifier (TIA)

2.1. High Impedance Amplifier

A generic high impedance amplifier is shown in figure 1. A detector, such as a pin diode, is connected in series with a resistance, $R_1$, and is reverse biased. The device is illuminated; a photo-current flows in the diode and resistor combination. A voltage proportional to the photo-current appears across $R_1$; it is presented to a high input impedance amplifier. While this topology is potentially the simplest to design, it has considerable practical problems [16]. The bandwidth of the system is often limited by the time constant created by the resistor, $R_1$, and the parasitic capacitance associated
with the input node, $c_p$. The noise performance of the system is often dominated by the value of $R_1$.

2.2. Impedance Matched Amplifier

The impedance matched amplifier shown in figure 2 is designed with a specified input impedance, $z_{in}$, which is usually 50 Ω or 75 Ω. This topology is particularly suited to microwave noise measurements. The measurement system, and in some cases, the APD impedance, must be carefully designed to ensure that nearly all of the noise power generated in the detector enters the measurement system. Xie et al. have used a matched approach to noise measurements [10]. In their system, the diode is terminated for AC signals by $Z_T = 50 \, \Omega$ such that both ends of the microstrip line are terminated. Assuming the APD has a large source impedance, $r_s$, and terminating both ends of the transmission line, the necessity of tightly controlling the APD impedance is removed. The noise power generated by the diode is divided equally between the termination, $Z_T$, impedance and the measurement system input impedance, $z_{in}$.

In the context of the present work, a diode capacitance, $c_j$, of 1 nF and 50 Ω input impedance would exhibit an input time constant of 50 ns, which corresponds to a -3 dB frequency of 3.2 MHz. Attaining wide bandwidths from devices with high junction capacitance requires a different approach.
2.3. Transimpedance Amplifier

A transimpedance amplifier, shown in figure 3, presents a low input impedance and converts the diode current into a voltage. From DC to HF frequencies operational amplifiers are often suitable for producing an effective TIA, requiring only a few external components [17]. The Analog Devices AD9631 [18] and Texas Instruments OPA129 [19] are specifically designed for this application. Several transimpedance amplifier topologies are reviewed in [20]. TIAs appear frequently in optical communications literature [21,22] and physical measurement applications in single ended [16,23–27] and differential forms [28]. Excluding the optical communications TIAs, which generally make use of integrated circuit technology, most measurement TIAs are developed to fulfill a specific instrumentation requirement. Consequently, there is considerable application-dependent topological variation of TIAs in the literature.

At HF frequencies, operational amplifier based TIAs are not suitable for measuring APDs with high junction capacitances. It may be shown that an approximation of the input impedance of an opamp-based TIA, when the opamp open loop gain is much greater than unity, is $z_{in} = R_F/A_v$. Open loop gain, $A_v$, decreases with increasing frequency and so input impedance rises. Opamp based transimpedance amplifiers are also prone to instability when the input node is loaded with significant capacitance [9]. Consider the input impedance of a transimpedance amplifier designed using an Analog Devices AD9631 at 10 MHz with a 2.2 kΩ feedback resistor. The open loop gain at 10 MHz is approximately 22 dB = 12.59 [18]. The resulting input impedance is approximately 175 Ω. The system reported by Lau et al., which has similar bandwidth and gain requirements as the present work, is unstable with junction capacitance in excess of 50 pF [9]. This paper is concerned with a new transimpedance amplifier designed using a bipolar transistor based front end.

3. Diode Small Signal Model

A small signal model of a reverse biased pin or nip diode is shown in figure 4. The model is comprised of a Norton source ($i_{ph}$ and $r_s$) with a parallel capacitance ($c_j$). $r_s$ is the small signal dynamic resistance of the APD. $c_j$ is proportional to the depletion width,
the area of the APD and the relative permittivity of the material. This model is used throughout the present work. It has proved satisfactory from DC to HF frequencies.

4. Noise Measurement System

A block diagram of the noise measurement system is shown in figure 5. The laser is chopped by mechanical means at 180 Hz and is presented to the diode via a system of optics. The TIA is used to convert the diode current into a voltage. This voltage is amplified using a series of operational amplifiers with a total terminated gain of $\sim 13$. A precision stepped attenuator is used to vary the system gain permitting measurement of high and low noise devices. The noise signal is separated from the low frequency component of the photo-current by $C_3$ and by the Minicircuits SBP-10.7+ LC ladder filter. After filtration, the signal resembles an amplitude modulated noise waveform, where periods of diode illumination produce greater noise amplitude than periods of darkness. Voltage gain ($\sim 41$ V/V terminated) provided by operational amplifiers follows, prior to a wide band, 250 MHz, power meter which is a squaring and averaging circuit. The squared, averaged, signal is further amplified sixteen times. The output of the squaring and averaging circuit is an approximately square wave voltage signal with a fundamental frequency of 180 Hz. The amplitude is proportional to the noise power contained within the pass band of the SBP-10.7+ filter. The squaring circuit is
based on an Analogue Devices AD835 analogue multiplier and is described in section 11. The averaging circuit is a first order RC filter with a time constant of approximately 100 µs. The output from the squaring and averaging circuit is measured using a lock-in-amplifier. The photo-current signal is taken from the output of the TIA where the amplitude of the 180 Hz square wave is proportional to the photo-current (see figure 8). The photo-current signal is measured on a second lock-in-amplifier.

5. Transimpedance Amplifier

A simplified circuit diagram of the TIA first stage and biasing circuit is shown in figure 6. There are two transistor stages which are electrically similar and are constructed to the same physical layout. One transistor stage which is composed of T1–3, R1, R2, R3, R4 and C1, is ‘active’ and is presented with the APD. The other, comprising T4–6, R5, R6, R7, R8 and C2, is ‘passive’ and has only quiescent (DC) conditions. The objective of the active transistor stage is to present a low impedance to the APD and to convert the photo-current + noise signal into a voltage. It is preferable to maintain the input node at ground. To enable this, one of the power supply rails, which supplies both transistor stages, is controlled by a feedback system including R10, C3, A1 and T7. The advantages of this will be discussed in section 9. The outputs of the two transistor stages are subtracted using an opamp circuit to remove, as far as possible, the DC conditions of the transistor stages from the noisy photo-current signal (figure 8). Further gain is provided by several more opamp stages. Output DC offset adjustment circuitry, similar to an opamp offset null, is also included (R11).

5.1. Transistor Stages

The transimpedance function of the first stage is realized by a single transistor common base amplifier. The operation of this configuration can be thought of as a “low impedance current amplifier” [29]. It is the load resistor which converts the current to a voltage. The common base small signal current gain, α, is slightly less than unity; therefore, the small signal current flowing in the load resistor, R1, is approximately equal to the small signal input current. It can be shown without difficulty that the transimpedance gain in the mid band is set by the load resistor. R2, R3 and R4 bias the transistor in a way that provides voltage and thermal stability. The capacitances C1 and C2 provide ground at the base for AC signals. C5 is required to make the source measure unit (SMU) a signal ground from the viewpoint of the APD, and to significantly lessen the high frequency noise that the SMU injects into the measurement system. The SMU is unstable when its output is loaded with more than 20 nF. A small resistance, R12, is added in series to isolate the SMU from C5.

5.1.1. Design Specification

The design requirements for the transistor TIA are:

• Transimpedance gain of approximately 2200 V/A (1100 V/A terminated).
Figure 6. Simplified bipolar transistor transimpedance amplifier

-3 dB bandwidth of at least 10 MHz with maximum device junction capacitance of 5 nF.
- Lowest obtainable noise.
- DC coupling of the APD to the TIA.
- DC coupling of the output to the rest of the front end.
- Guaranteed stability, irrespective of APD capacitance.
- Operate from +/-15 Volt supplies.

The first three points are bound to the transistor operating conditions. The remainder are system considerations.

5.1.2. Front End Description  Using a simplified hybrid–π model, the common base stage may be represented with two real poles [20]. The lower frequency pole is formed at the emitter node by the APD junction capacitance, $c_j$, and the impedance looking into the emitter, $r_e$. The higher frequency pole is formed at the collector node by the load resistor, $R_1$, and all stray capacitances from the collector, including the input capacitance of the buffer opamp, which is approximately 5 pF for the Analog Devices AD829.

Assuming that the transistor stage can be approximated by a first order low pass system, composed of only the input pole, the required time constant $(1/(2\pi)) \times 10^{-7}$ is formed by $c_j = 10 \text{ nF}$ and the small signal resistance looking into the emitter, $r_e$. The
maximum permissible value of $r_e$ is approximately 1.6 Ω. $r_e$ does not appear explicitly in the hybrid–π model but it can be shown that $r_e \to 1/g_m$ as $\beta \to \infty$. The collector current required to achieve $r_e = 1.6$ Ω is given by $I_C = (kT) / (e r_e) \approx 16$ mA. Where $k$ is Boltzmann’s constant, $T$ is the absolute temperature and $e$ is the electron charge.

It is not possible to obtain all of the required gain from the first stage for two reasons. Firstly, the collector to base junction capacitance (0.3 pF) and the input capacitance of the opamp buffer (5 pF) form a pole with the load resistance at the collector node. Using a load resistance of 2.2 kΩ produces a pole at approximately 14 MHz. Stray capacitance will increase the datasheet capacitance values, further reducing the pole frequency. Secondly, given the collector current requirement, the DC voltage drop across 2.2 kΩ would require the use of a supply voltage greater than +/- 15 V. This is undesirable because the circuit is used in a measurement system where +/- 15 V is already available. A 500 Ω load resistance provides an acceptable frequency response, first stage gain and DC conditions at the collector. Additional voltage amplification is required, and is supplied using operational amplifiers after the transistor stage (figure 8). The transistor stage supplies approximately one quarter of the required gain.

6. TIA Characterization

The circuit in figure 7 is used to facilitate direct measurement of the transimpedance amplifier’s frequency response and to simulate the APD capacitance. The impedance of the resistor chain is constant within 3 dB from DC to 20 MHz. As frequency increases, the chain impedance falls as the resistors parasitic capacitance becomes more significant.

SPICE simulation and experimental measurements in which $c_j$ is stepped from 0 nF to 5 nF in 0.5 nF steps results in a resonance phenomenon in the amplitude response, near to the noise measurement frequency (figure 15). Grey et al. have noted the necessity of considering the effect of the base spreading resistance $r_b$ when the quiescent conditions cause the base emitter resistance, $r_{be}$, to become the same order of magnitude as $r_b$ [29]. The input and output poles are weakly interacting provided $r_{be} >> r_b$. While this is the case, the transistor’s internal base node is close to ground potential. If $r_{be}$ is approximately equal to $r_b$, the transistor’s internal base node, which joins $r_{be}$ and $r_b$, will deviate significantly from ground, and energy transfer between the emitter and base circuits will be possible. Further analysis of the frequency response and stability of the common base circuit, when $r_{be} \approx r_b$, is given in [30].

A result of the interaction of the input and output poles is a peaking effect in the
frequency response which can be seen in figure 15. It is advantageous to minimize the peaking effect caused by the change in \(c_j\). The available circuit variables that impact the magnitude of the peaking include,

- Base spreading resistance, \(r_b\).
- Collector–Base depletion capacitance, \(c_{cb}\).
- Base–Emitter depletion capacitance, \(c_{be}\).
- Base–Emitter resistance, \(r_{be}\).

Unfortunately none of these variables can be altered sufficiently to make the peaking effect negligible while maintaining the gain and bandwidth requirements. It is neither possible nor desirable to make \(r_b\) zero. All of the available parameters are ultimately bound to the transistor’s quiescent conditions. In this design problem, where \(r_b\) is a significant fraction of \(r_{be}\), the dependence of the frequency response on the APD junction capacitance appears to be unavoidable. All transistors have inter-electrode capacitances and non-zero \(r_b\), and will exhibit peaking. Any transistor with very low \(r_b\) would be unsuitable because it would oscillate. The necessity of \(r_b\) to maintain stability will be described shortly. A calibration procedure similar to that reported in [9] has been used to mitigate the peaking effect.

7. Stability

Unlike TIAs with feedback around more than one active device, there is no intentional negative feedback in the single transistor stage. Consequently, regeneration due to excessive phase shift, arising from useful components, is not a concern. However, any physical realization of a circuit possessing power gain may oscillate under sympathetic conditions. Both emitter follower and common base transistor circuits may suffer instability due to the impedance transforming nature of the transistor [16, 30, 31]. SPICE simulation of the input impedance of the TIA while connected to the APD small signal model in figure 4 provides an estimation of the likelihood of oscillation. For the common base transistor, oscillation is likely when the real part of the impedance looking into the emitter, \(\Re(z_{in})\), is negative for any frequency below the transistor’s transition frequency [30].

Two avenues of attack are open to combat oscillation problems, paralleling transistors and increasing the base resistance.

7.1. Parallel Transistors

Several transistors may be connected in parallel such that the total terminal currents are shared approximately equally between the devices. Each transistor then has lower transconductance than if a single device is used. Consequently, the impedance looking into each transistor’s emitter is greater than for an equivalently biased single transistor. The new operating conditions, under which each transistor has an increased \(r_e\), are
7.2. Increased Base Resistance

Secondly, a small resistance may be connected in series with the transistor base lead. A resistance in this location worsens the effect of gain peaking and increases the stage noise. If the resistance is made too large, it will affect the frequency response and frequency independent gain of the stage. Observation of the time constant, \( \tau \), and frequency independent gain, \( k \), of the transimpedance gain equation in Appendix A shows this. The use of both mechanisms provides a reasonable compromise between noise, stability, ease of construction and frequency response.

8. Physical Construction

The transistors \( T_{1-3} \) and \( T_{4-6} \) (SOT-23 package) are placed on top of each other. Vertically placed wires connect their electrodes together. The transistor closest to the circuit board (standard 1.6 mm FR4) is soldered to the copper tracks. No other sufficiently compact layout has been found. This layout also provides a degree of isothermal matching. Several board designs were developed in order to explore the layout options and noise reducing effect of paralleling transistors. Up to fifteen paralleled transistors were used in various layout implementations. The distance between emitter terminals in parallel-transistor front ends must be as small as is practically possible when the impedance looking into the input is low. Any parasitic components formed
between the individual transistor terminals have deleterious effects on the frequency response, rendering the layout unusable.

9. Transistor Stage Biasing Feedback System

The biasing feedback system is designed to permit DC coupling of the APD to the TIA by maintaining the active transistor stage emitter node at ground potential. The bias voltage displayed on the APD biasing equipment is then equal to the bias appearing across the device. A further benefit of DC coupling is that all of the noise current flows into the TIA (assuming $r_s \gg z_{in}$). Unlike AC coupling, no separate DC path is required through which noise power could pass undetected.

The feedback system allows the TIA to be DC coupled by modifying the lower supply rail voltage of both transistor stages in order to maintain the emitter nodes close to ground potential. This is implemented by using the second transistor stage which possesses, as closely as practically possible, the same biasing conditions as the first, but without an APD connected. The quiescent conditions of this passive stage are measured electronically and the lower supply rail of both transistor stages is adjusted in order to bring the emitter nodes to ground potential. The opamp, $A_1$, changes the voltage on the lower supply rail in order to maintain the inverting input and the non-inverting input at the same potential (ground). The opamp, $A_1$, and the RC network of $C_3$ and $R_{10}$ form an integrator with a time constant of a few seconds ($C_3$ is in the feedback loop of the opamp). $R_{10}$ is large in order to avoid loading the emitter node of $T_{4-6}$, and in order that a long time constant should be formed by a reasonably small value electrolytic capacitor. The potentiometer, $R_{11}$ may be used to introduce an extra DC current to the passive front end, modifying its DC conditions slightly in order to produce an offset null control.

10. Opamp Stages

It is preferable to DC couple the output of the common base stage as well as the input. This permits the use of a non-inverting amplifier to buffer the collector voltage without requiring a (large) resistance to provide a biasing current path. Furthermore, the DC output voltage of the front end will be proportional to the magnitude of the non-photogenerated current component in devices where several transport mechanisms contribute to current flow. Using the two common base stages, with nearly identical operating conditions, the output offset caused by DC coupling can be greatly lessened by subtracting one output from the other. This is accomplished with the operational amplifier based subtraction circuit shown in figure 8. The buffer opamp has a lower bandwidth than the noise measurement frequency. Allowing the transistor stage DC conditions to propagate through the opamp stages in figure 8 would quickly lead to saturation of the output against the power supply rail.
11. Power Meter

The power meter design shown in figure 9 was developed by B. K. Ng [33], and is based on the Analog Devices AD835 wide band analogue multiplier. The AD835 is shown in block diagram form in figure 9. The inputs of the multiplier are connected together in order that the output voltage is dominantly proportional to the square of the input. A small linear term and a DC offset remain. The DC offset is lessened by trimming the offset null of the TL071 and finally removed by AC coupling the power meter output to the lock-in-amplifier. The linear term is partially resultant from imperfection in the squaring process, but it is likely that the majority of the linear term is a result of regressing manually collected calibration data possessing a small uncertainty. The squared noise voltage signal is applied to a first order low pass filter, \( R_4 \) and \( C_1 \), with a time constant of approximately 100 \( \mu \)s. The filter is required to pass the 180 Hz chopping signal such that the LIA can differentiate between light and dark periods, but must filter the noise signal (around 10 MHz) to produce a signal representing the average value of the squared noise voltage.

The transfer function for the multiplier circuit is,

\[
V_o = 16 \cdot \frac{1.303}{1.05} \left[ \frac{1}{2} \left( \frac{V_i}{2} \right)^2 \right] = 2.48 V_i^2
\]

where \( V_o \) is the DC output voltage and \( V_i \) is the peak to peak voltage of a deterministic test input signal.

12. System Noise Analysis

Analytical and SPICE methods which complement each other are used to isolate the dominant noise source in the transistor stages. A first order analytical model, shown in figure A1, is developed by ignoring the base emitter capacitance, \( c_{be} \), and the base collector capacitance, \( c_{cb} \), and the load capacitance, \( C_L \). The error introduced by
the simplifications is shown not to significantly affect the results at the frequencies of interest. An introduction to noise analysis techniques can be found in [34] and [35]. The transistor stages are the dominant noise contributor in the front end. Figure 12 shows that the APD junction capacitance, $c_j$, acts to increase the noise contribution of the active transistor stage; therefore, under normal operating conditions, only the active transistor stage’s noise contribution is significant.

To assess the usefulness of the analytical noise model over the frequency range of interest, it was compared with a SPICE model using identical small signal parameters. The comparison is shown in figure 10. The deviation of the analytical model from the SPICE model occurs above 50 MHz. The deviation has been investigated by adding components to the analytical model and using it to produce numerical results. These results are compared with SPICE analyses. A purely analytical route quickly becomes impractical, due to the large number of terms appearing in the equations as the circuit becomes higher order. The high frequency discrepancy between SPICE and the first order analytical model is principally due to the effect of the collector – base junction capacitance, $c_{cb}$. The load capacitance, $C_L$, and the base emitter junction capacitance, $c_{be}$, also play some part, but their effect is less dominant in this particular design. Since the noise measurement is performed at 10 MHz, the first order approximation does not represent a significant loss of accuracy. The mid-band error between the SPICE transistor model and the first order analytical model is 84 pV/√Hz, with the analytical model producing an over-estimation of noise.

The noise contribution of each noise source in the transistor stage is shown in figure 11. The following conclusions may be drawn from the noise analysis,

- Around 10 MHz, the dominant noise contributor is the thermal noise associated...
High detector capacitance transimpedance amplifier

Figure 11. A plot of the noise contributions which appear at the collector of $T_{1-3}$ resulting from each noise generator in the analytical model. It can be seen that for the particular biasing conditions evaluated the base spreading resistance dominates the noise at the output. The small signal parameters for this analysis are $I_C = 15$ mA, $\beta = 105.8$, $r_b = 56.53 \, \Omega$, $R_E = 680 \, \Omega$ and $c_j = 1 \, \text{nF}$. The model used to generate this figure is outlined in Appendix A.

with the transistor’s base spreading resistance and the external base resistor, $E_n r_b$. The collector–emitter shot noise, $I_{nC}$, and collector–base shot noise, $I_{nB}$, are the second and third largest contributors respectively. Uncorrelated sources sum as the root of the contributions squared; the base spreading resistance thermal noise is responsible for almost all of the output noise.

- Paralleling transistors reduces the base spreading resistance thermal noise contribution. Noise can be reduced without limit using this method in almost all practical cases, because the amplifier noise is greater than the source noise [32]. The diode noise signal to front-end noise ratio is often negative (table 1). The practical problems of layout are the limiting factor in this respect.

- Increasing the value of the emitter biasing resistor, $R_E$, acts to reduce the noise gain of the emitter biasing resistor’s thermal noise, $E_n R_E$, to the collector node.

- Decreasing the time constant of $R_E$ and $c_j$ will increase the frequency at which the noise gain of the base spreading resistance, $E_n r_b$, the dominant noise source, begins to rise. However, decreasing the value of $R_E$ in order to move the zero up in frequency will increase the frequency independent gain, $k$, for the $E_n R_E$ generator by an amount which is sufficient to negate any improvement in noise performance. Observation of the noise gain equations in Appendix A will show this.
High detector capacitance transimpedance amplifier

Figure 12. Transistor TIA noise measured using a HP4396A spectrum analyzer and SPICE modelling of the TIA noise. In the measured data the TIA is connected in series with the SBP-10.7+ filter in order to show the extent of the pass band. \( c_j = 0 \text{ nF (lowest noise)} \) to \( 5 \text{ nF (highest noise)} \) in \( 0.5 \text{ nF steps} \). The relationship between capacitance and line style is identical to figure 15. The SPICE simulation results shown are the noise voltage at the ‘Noise O/P’ in figure 8 for the same set of capacitance as the measured data. The insertion loss of the SBP-10.7+ is not more than 1.5 dB [36]; the total error between SPICE and the measured performance is 2.5 dBV – approximately 13% – with SPICE underestimating the noise.

Figure 13. TIA effective noise power bandwidth as a function of device capacitance. The effective noise power bandwidth is used in the calibration process [9]. It is given by the normalized area under the transfer response within the bandwidth of the filter (see section 13).

13. Effective Noise Power Bandwidth

The effective noise power bandwidth (ENBW) is a function of APD capacitance due to the peaking effect that was described earlier. The ENBW is used in the calibration
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Figure 14. Reference data from two Si pin devices

Figure 15. A plot of frequency response showing the variation in relative response with increasing APD junction capacitance, $c_j$. $c_j = 0$ nF (solid), 0.5 nF (long dash), 1 nF (medium dash), 1.5 nF (short dash), 2 nF (dash dot dash), 2.5 nF (dash dot dot), 3 nF (medium medium dash), 3.5 nF (short short short dash), 4 nF (short short dash), 4.5 nF (short long dash), 5 nF (long short dash).

process to normalize the bandwidth of the measurement system when the reference device and the APD have differing junction capacitance [9]. The ENBW is given by the normalized area under the transfer response within the bandwidth of the SBP-10.7+ filter,

$$\text{ENBW}(c_j) = \frac{1}{A_0} \int_0^\infty |A(f, c_j)| \, df$$

where $A$ is the transimpedance gain at a frequency, $f$, with a particular value of $c_j$, $A_0$ is the transimpedance gain at the center frequency of the SBP-10.7+. In practice,
this integration is performed numerically on data obtained by connecting the circuit of figure 7 to the TIA and recording the transfer response using a vector network analyzer.

14. Noise Measurements

The photomultiplication is calculated by normalizing the measured photocurrent to an extrapolation of the primary photocurrent measured at low bias. The excess noise factor is obtained using,

\[ F = \frac{N_{\text{APD}}}{N_{\text{shot}} M^2} \frac{B_{\text{APD}}}{B_{\text{shot}}} \]  

where \( F \) is the excess noise factor, \( N_{\text{APD}} \) is the measured noise data. \( N_{\text{shot}} \) is the shot noise produced by the primary photocurrent measured on the reference device, \( M \) is the photomultiplication, and \( B_{\text{APD}} \) and \( B_{\text{shot}} \) normalize the measurement bandwidth for the experiment such that the differing effective noise power bandwidth of the APD noise measurement data and the reference noise measurement data is accounted for. The measurement is performed by first measuring the relationship between photocurrent and shot noise using a reference device (Perkin Elmer UV-BQ40) for a range of photocurrents (this data is shown in figure 14). The photocurrent is varied by altering the optical intensity falling on the device. The APD noise is then measured; a starting photocurrent is set, thereafter the optical intensity incident on the device is kept constant. Photocurrent and excess noise are recorded while increasing reverse bias voltage. \( B_{\text{APD}} \) and \( B_{\text{shot}} \) are the effective noise power bandwidth of the measurement system for the APD and the reference device respectively.
Figure 16 shows some example noise data measured using our new measurement system. This data is a set of five 1 mm$^2$ Si pin APDs with nominal intrinsic region widths ranging from 31 nm to 350 nm. The thinnest structure exhibits approximately 3 nF junction capacitance and tunneling current several orders of magnitude greater than the current due to avalanche multiplication of photo-generated carriers. This data was obtained with 633 nm light which elicits mixed carrier injection. This data is in good agreement with Tan et al. [37] for similar intrinsic width structures. This data has been confirmed using a second, novel high capacitance noise measurement system, which is also an original design of the authors’. This second measurement system will be reported separately. The implications of this data for impact ionization in thin Silicon structures will be discussed in a further forthcoming publication.

15. Comparison with other Excess Noise Measurement Systems

Several noise measurement systems have been reported in the literature and comparisons between this front end and those previously reported may be drawn. Several of the systems reported are reviewed in [38]. The figures of comparison are:

- The system signal to noise ratio (SNR), where the signal is defined as full shot noise exhibited by 1 µA.
- The maximum permissible junction capacitance. In the case of multi-frequency systems, such as the system reported by Xie et al. [10], the lowest available frequency is used; this produces the most favorable result. It is assumed that the system input impedance and diode junction capacitance form a first order low pass network.

The first reported noise measurements on photodiodes was by Baertsch [39]. Insufficient information is provided to estimate this system’s figures of merit so it is excluded from the comparison. Brain reports several results on commercial Silicon [40] and Germanium detectors [41–43] and photo-transistors [44]. Xie et al. [10] proposed a measurement system that was substantially similar to Toivonen et al. [45]. The Xie et al. system represents both. Bulman [12], Ando and Kanbe [11] and Lau et al. [9] presented systems based on phase sensitive detection. Xie, Toivonen and Brain used a DC approach.

Bulman’s report lacks some information regarding the front end amplifier. An Analog Devices AD9618 low noise opamp in non-inverting mode is used as a model. It achieves a gain of 100 V/V and a bandwidth of 80 MHz with 50 Ω input impedance. The equivalent input noise voltage is 1.94 nV/√Hz. Ando and Kanbe do not give information regarding the model numbers or manufactures of their system components. No noise specifications for the instrumentation are given. Consequently their system is assumed not to add any extra noise. Only the noise power available within the measurement bandwidth is compared with the signal. Because the input impedance of Brain’s system is not reported the maximum device capacitance is not calculable, however it may be expected to be quite large as the frequency range of Brain’s system was 1.6 kHz –
16. Conclusion

A new transimpedance amplifier has been presented for use in measurements of excess noise in avalanche photodiodes. This amplifier has been successfully used in situations where the device being measured has high junction capacitance, up to 5 nF, and where the device has lower than usual dynamic resistance $r_s$. The measurement of a much wider range of devices is now possible. This circuit design has been used in measurements of devices with particularly high dark currents. The system may also be employed to measure devices in modes of operation which increase the capacitance and decrease the dynamic resistance such as forward bias and photo-voltaic mode. Noise measurements may yield novel methods of grading the quality of solar cells.

Appendix A. Analytical Noise Equations for the Common Base Stage

This appendix lists the noise gain of each noise generator in the first order analytical noise model of the common base stage (figure A1). The noise gain of each noise contributor can be expressed as either a low pass or a pole–zero form, except for the contribution of the load resistor which appears directly at the output.

For the input current ($i_{in}$)

$$\frac{v_c}{i_{in}} = k \cdot \frac{1}{1 + s \cdot \tau}$$  \hspace{1cm} (A.1)

$$k = \frac{R_L \cdot r_{be} \cdot R_E \cdot g_m}{r_{be} + r_b + R_E + g_m \cdot r_{be} \cdot R_E}$$  \hspace{1cm} (A.2)

$$\tau = \frac{c_j \cdot R_E \cdot (r_{be} + r_b)}{r_{be} + r_b + R_E + g_m \cdot r_{be} \cdot R_E}$$  \hspace{1cm} (A.3)
Figure A1. Small signal noise model of common base amplifier. Definitions of the terms are given in table A1. The magnitudes of the noise generators are given in table A2.

Table A1. Table of analytical noise model components

<table>
<thead>
<tr>
<th>Name</th>
<th>Description</th>
<th>Component of</th>
</tr>
</thead>
<tbody>
<tr>
<td>$E_n r_b$</td>
<td>Voltage noise source</td>
<td>Base spreading resistance</td>
</tr>
<tr>
<td>$r_b$</td>
<td>Physical resistance</td>
<td>Transistor base – emitter junction</td>
</tr>
<tr>
<td>$I_n b$</td>
<td>Current noise source</td>
<td>Transistor base – collector junction</td>
</tr>
<tr>
<td>$I_n c$</td>
<td>Current noise source</td>
<td></td>
</tr>
<tr>
<td>$R_E$</td>
<td>Physical resistance</td>
<td>Emitter biasing resistance</td>
</tr>
<tr>
<td>$E_n R_E$</td>
<td>Voltage noise source</td>
<td>APD small signal model</td>
</tr>
<tr>
<td>$I_{in}$</td>
<td>AC signal current</td>
<td></td>
</tr>
<tr>
<td>$c_j$</td>
<td>Physical capacitance</td>
<td></td>
</tr>
<tr>
<td>$g_m v_{be}$</td>
<td>current source</td>
<td>The model of transistor action</td>
</tr>
<tr>
<td>$r_{be}$</td>
<td>Model resistance</td>
<td></td>
</tr>
<tr>
<td>$R_L$</td>
<td>Physical resistance</td>
<td></td>
</tr>
<tr>
<td>$E_n R_L$</td>
<td>Voltage noise source</td>
<td>Load resistance</td>
</tr>
</tbody>
</table>

For the load resistance ($E_n R_E$)

$$\frac{v_c}{E_n R_L} = 1$$  \hspace{1cm} (A.4)

For the base-collector shot noise ($I_n c$)

$$\frac{v_c}{I_n c} = k \cdot \frac{1 + s \tau_1}{1 + s \tau_2}$$  \hspace{1cm} (A.5)

$$k = \frac{R_L (R_E + r_{be} + r_b)}{R_E + g_m r_{be} R_E + r_{be} + r_b}$$  \hspace{1cm} (A.6)
Table A2. Table of noise source expressions

<table>
<thead>
<tr>
<th>Source</th>
<th>Description</th>
<th>Expression</th>
<th>$[\frac{V^2}{Hz} \cdot \frac{A^2}{Hz}]$</th>
</tr>
</thead>
<tbody>
<tr>
<td>$E_n r_b$</td>
<td>Thermal noise of the base spreading resistor</td>
<td>$4kT r_b$</td>
<td></td>
</tr>
<tr>
<td>$I_n b$</td>
<td>Shot noise in the b–c junction</td>
<td>$2e I_C$</td>
<td></td>
</tr>
<tr>
<td>$I_n c$</td>
<td>Shot noise in the b–e junction</td>
<td>$2e I_B$</td>
<td></td>
</tr>
<tr>
<td>$E_n R_E$</td>
<td>Thermal noise of the emitter biasing resistor</td>
<td>$4kT R_E$</td>
<td></td>
</tr>
<tr>
<td>$E_n R_L$</td>
<td>Thermal noise of the load resistor</td>
<td>$4kT R_L$</td>
<td></td>
</tr>
</tbody>
</table>

Where $k$ is Boltzmann’s constant, $e$ is the electron charge, $T$ is the absolute Temperature.

$$
\tau_1 = \frac{c_j R_E (r_{be} + r_b)}{R_E + r_{be} + r_b} \quad (A.7)
$$

$$
\tau_2 = \frac{R_l r_{be} g_m (r_b + r_{be})}{g_m r_{be} R_E + R_E + r_{be} + r_b} \quad (A.8)
$$

For the base–emitter shot noise ($I_n b$)

$$
\frac{v_c}{I_n b} = k \cdot \frac{1 + s \tau_1}{1 + s \tau_2} \quad (A.9)
$$

$$
k = - \frac{R_l r_{be} g_m (r_b + r_{be})}{g_m r_{be} R_E + R_E + r_{be} + r_b} \quad (A.10)
$$

$$
\tau_1 = \frac{c_j R_E r_b}{r_b + R_E} \quad (A.11)
$$

$$
\tau_2 = \frac{c_j R_E (r_{be} + r_b)}{g_m r_{be} R_E + R_E + r_{be} + r_b} \quad (A.12)
$$

For the base spreading resistance ($E_n r_b$)

$$
\frac{v_c}{E_n r_b} = k \cdot \frac{1 + s \tau_1}{1 + s \tau_2} \quad (A.13)
$$

$$
k = - \frac{R_l g_m r_{be}}{g_m r_{be} R_E + R_E + r_b + r_{be}} \quad (A.14)
$$

$$
\tau_1 = c_j R_E \quad (A.15)
$$

$$
\tau_2 = \frac{c_j R_E (r_{be} + r_b)}{g_m r_{be} R_E + R_E + r_{be} + r_b} \quad (A.16)
$$

For the emitter biasing resistance ($E_n R_E$)

$$
\frac{v_c}{E_n R_E} = k \cdot \frac{1}{1 + s \tau} \quad (A.17)
$$
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\[
\tau = \frac{g_m r_{be} R_E}{g_m r_{be} R_E + R_E + r_b + r_{be}} \quad (A.19)
\]

Acknowledgments

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References

High detector capacitance transimpedance amplifier

[36] Mini-Circuits, P.O. Box 350166, Brooklyn, NY 11235 U.S.A. *SBP-10.7+ Coaxial Bandpass Filter*.


