

# Application Note #6

## Signal Enhancement

### Introduction

Many experimental techniques rely on the quantitative measurement of charged particles. Applications range from the very simple, such as the measurement of a DC anode current by a picoammeter, to the complex, such as the measurement of atomic state lifetimes using time-resolved particle counting.

Often, the signal of interest is obscured by noise. The noise may be fundamental to the process: discrete charges are governed by Poisson statistics which gives rise to shot noise. Or, the noise may be from more mundane sources, such as microphonics, thermal emf's, or inductive pick-up.

This article will describe methods for making useful measurements of weak signals, even in the presence of large interfering sources, emphasizing the electronic aspects of the measurement.

Figure 1 details the elements of a typical measurement situation. In this experiment the photo ionization cross section for a gas will be determined by passing a laser through the gas and by measuring the number of ions which are created.

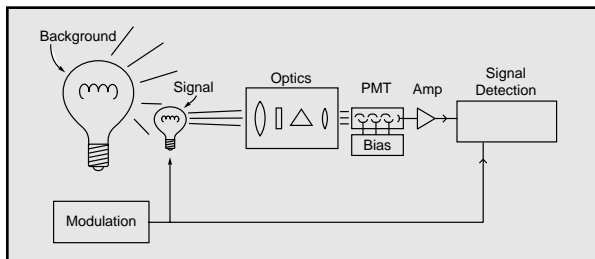


Figure 1: Prototype Experiment

In many applications, the output of the detector must be amplified or converted from a current to a voltage before the signal may be analyzed. Selection criteria for amplifiers include type (voltage or transconductance), gain, bandwidth, and noise. Often, the amplifier noise will be the limiting factor in determining the S/N ratio in a measurement, especially in situations where the charge detector has no gain.

### Signal Analysis

There are two broad categories of signal analysis, depending on whether or not the source is modulated. Modulating the source allows the signal to be distinguished from the background. Often, source modula-

tion is inherent to the measurement. For example, when a pulsed laser is used to induce a current, the signal of interest is present only after the laser fires. Other times, the modulation is "arranged", as when a CW source is chopped. Sometimes the source cannot be modulated, or the source is so dominant over the background as to make modulation unnecessary.

### Noise Sources

An understanding of noise sources in a measurement is critical to achieving signal-to-noise performance near theoretical limits. The quality of a measurement may be substantially degraded by a trivial error. For example, a poor choice of termination resistance for an electron multiplier may increase current noise by several orders of magnitude.

### Shot Noise

Light and electrical charge are quantized, and so the number of photons or electrons which pass a point during a period of time are subject to statistical fluctuations. If the signal mean is  $M$  photons, the standard deviation (noise) will be  $\sqrt{M}$ , hence the  $S/N = M/\sqrt{M} = \sqrt{M}$ . The mean,  $M$ , may be increased if the rate is higher or the integration time is longer. Short integration times or small signal levels will yield poor S/N values. Figure 2 shows the S/N which may be expected as a function of current level and integration time for a shot-noise limited signal.

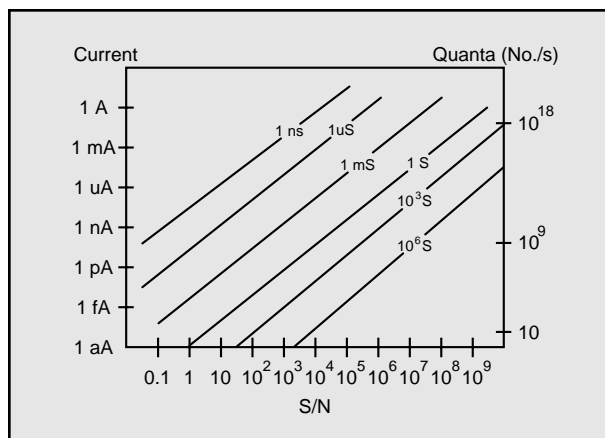


Figure 2: S/N vs. Flux and Time

"Integration Time" is a convenient parameter when using time domain signal recovery techniques. "Bandwidth" is a better choice when using frequency domain techniques. The rms noise current in the bandwidth

$\Delta f$  Hz due to a "constant" current, I amps, is given by:  
 $I_{\text{shot noise}} = \sqrt{(2qI\Delta f)}$  where  $q=1.6 \times 10^{-19}$  C

### Johnson Noise

The electrons which allow current conduction in a resistor are subject to random motion which increases with temperature. This fluctuation of electron density will generate a noise voltage at the terminals of the resistor. The rms value of this noise voltage for a resistor of R Ohm, at a temperature of T degrees Kelvin, in a bandwidth of  $\Delta f$  Hz is given by:

$$V_{\text{rms}_{\text{johnson}}} = \sqrt{(4KTR\Delta f)}$$

where K is Boltzman's constant. The noise voltage in a 1 Hz bandwidth is given by:

$$V_{\text{rms}_{\text{johnson}}}/\sqrt{\text{Hz}} = 0.13\text{nV} \times \sqrt{(R(\text{ohm}))}$$

Since the Johnson noise voltage increases with resistance, large value series resistors should be avoided in voltage amplifiers. For example, a 1 k $\Omega$  resistor has a Johnson voltage of about 4.1 nV/ $\sqrt{\text{Hz}}$ . If detected with a 100 MHz bandwidth, the resistor will show a noise of 41  $\mu\text{V}_{\text{rms}}$ , which has a peak-to-peak value of about 200  $\mu\text{V}$ .

When a resistor is used to terminate a current source, or as a feedback element in a current-to-voltage converter, it will contribute a noise current equal to the Johnson noise voltage divided by the resistance. Here, the noise current in a 1 Hz bandwidth is given by:

$$I_{\text{rms}_{\text{johnson}}}/\sqrt{\text{Hz}} = 130 \text{ pA} / \sqrt{R(\text{ohm})}$$

Since the Johnson noise current increases as R decreases, small value resistors should be avoided when terminating current sources. Unfortunately, small terminating resistors are required to maintain a wide frequency response. If a 1 k $\Omega$  resistor is used to terminate a current source, the resistor will contribute a noise current of about 4.1 pA/ $\sqrt{\text{Hz}}$ , which is about 1000 $\times$  worse than the noise current of an ordinary FET input operational amplifier.

### 1/f Noise

The voltage across a resistor carrying a constant current will fluctuate because the resistance of the material used in the resistor varies. The magnitude of the resistance fluctuation depends on the material used: carbon composition resistors are the worst, metal film resistors are better, and wire wound resistors provide the lowest 1/f noise. The rms value of this noise source for a resis-

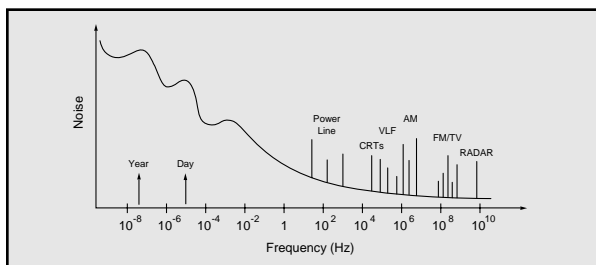
tance of R ohm, at a frequency of f Hz, in a bandwidth of  $\Delta f$  Hz is given by:

$$V_{\text{rms}_{1/f}} = IR \times \sqrt{(A \Delta f/f)}$$

where the dimensionless constant A has a value of about  $10^{-11}$  for carbon. In a measurement in which the signal is the voltage across the resistor (IR), then the  $S/N = 3 \times 10^5 \sqrt{(f/\Delta f)}$ . Often, this noise source is a troublesome source of low frequency noise in voltage amplifiers.

### Non-Essential Noise Sources

There are many discrete noise sources which must be avoided in order to make reliable low level light measurements. Figure 3 shows a simplified noise spectrum on log-log scales. The key features in this noise spectra are frequencies worth avoiding: diurnal drifts (often



**Figure 3: Simplified Noise Spectrum**

seen via input offset drifts with temperature), low frequency (1/f) noise, power line frequencies and their harmonics, switching power supply and CRT display frequencies, commercial broadcast stations (AM, FM, VHF and UHF TV), special services (cellular telephones, pagers, etc.), microwave ovens and communications, to RADAR and beyond.

The best alternatives for avoiding these noise sources are:

- 1) Shield to reduce pick-up.
- 2) Use differential inputs to reject common mode noise.
- 3) Bandwidth limit the amplifier to match the expected signal bandwidth.
- 4) Choose a quiet frequency for signal modulation when using a frequency domain detection technique.
- 5) Trigger synchronously with interfering source when using a time domain detection technique.

Common ways for extraneous signals to interfere with a measurement are illustrated in Figure 4. Noise may be injected via a stray capacitance as in Figure 4a. The

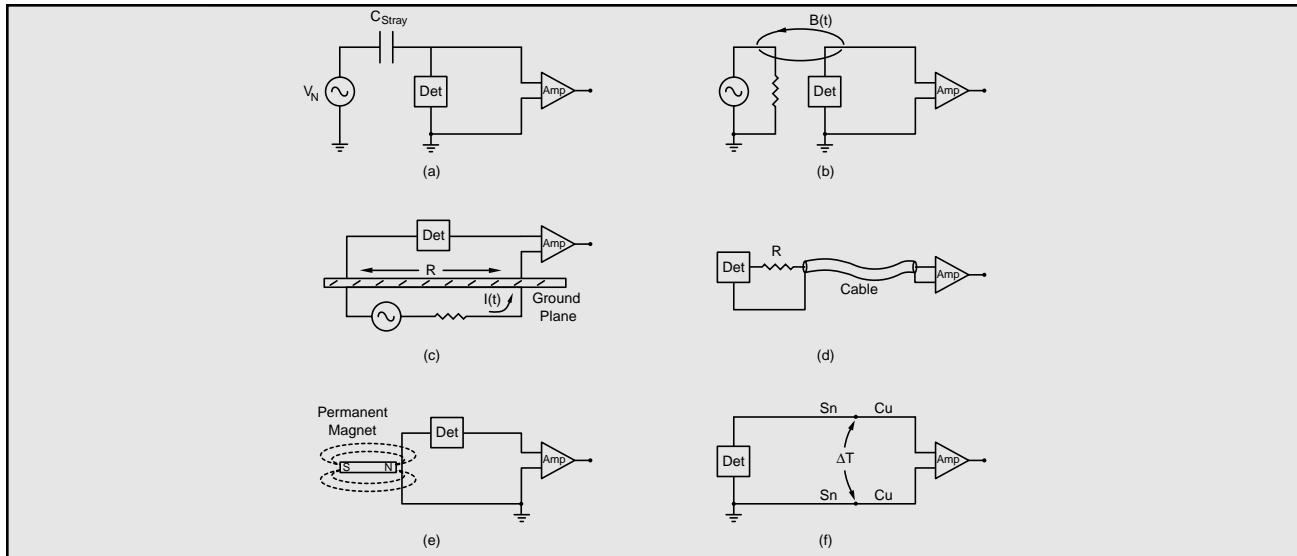


Figure 4: Coupling of Noise Sources

stray capacitance has an impedance of  $1/j\omega C$ . Substantial currents may be injected into low-impedance systems (such as transconductance inputs), or large voltages may appear at the input to high-impedance systems.

Inductive pick-up is illustrated in Figure 4b. The current circulating in the loop on the left will produce a magnetic field which in turn induces an emf in the loop on the right. Inductive noise pick-up may be reduced by reducing the areas of the two loops (by using twisted pairs, for example), by increasing the distance between the two loops, or by shielding. Small skin depths at high frequencies allow non-magnetic metals to be effective shields, however high- $\mu$  materials must be used to shield from low frequency magnetic fields.

Resistive coupling, or a "ground loop", is shown in Figure 4c. Here, the detector senses the output of the experiment plus the IR voltage drop from another circuit which passes current through the same ground plane. Cures for ground loop pickup include: grounding everything to the same point, using a heavier ground plane, providing separate ground return paths for large interfering currents, and using a differential connection between the signal source and amplifier.

Mechanical vibrations can create electrical signals (microphonics) as shown in Figure 4d. Here, a coaxial cable is charged by a battery through a large resistance. The voltage on the cable is  $V=Q/C$ . Any deformation of the cable will modulate the cable's capacitance. If the period of the vibration which causes the deformation is short compared to the RC time constant, then the stored charge on the cable,  $Q$ , will remain constant. In this case, a 1 ppm modulation of the cable capacitance will generate an ac signal with an amplitude of 1 ppm of the DC bias on the cable, which may be larger than the signal of interest.

The case of magnetic microphonics is illustrated in Figure 4e. Here, a DC magnetic field (the Earth's field or the field from a permanent magnet in a latching relay, for example) induces an emf in the signal path when the magnetic flux through the detection loop is modulated by mechanical motion.

Unwanted thermocouple junctions are an important source of offset and drift. As shown in Figure 4f, two thermocouple junctions are formed when a signal is connected to an amplifier. For typical interconnect materials (copper, tin) one sees about  $10 \mu\text{V}/^\circ\text{C}$  of offset. These extraneous junctions occur throughout instruments and systems: their impact may be eliminated by making ac measurements.

### Amplifiers

Several considerations are involved in choosing the correct amplifier for a particular application. Often, these considerations are not independent, and compromises will be necessary. The best choice for an amplifier depends on the electrical characteristics of the detector, and on the desired gain, bandwidth, and noise performance of the system.

Charge counting and fast gated integration require amplifiers with wide bandwidth. A 350 MHz bandwidth is required to preserve a 1 ns rise time. The input impedance to these amplifiers is usually  $50 \Omega$  in order to terminate coax cables into their characteristic impedance. When PMT's (which are current sources) are connected to these amplifiers, the  $50 \Omega$  input impedance serves as the current to voltage converter for the PMT anode signal. Unfortunately, the small termination resistance and wide bandwidth yield lots of current noise.

It is important to choose an amplifier with a very high input impedance and low input bias current when amplifying a signal from a source with a large equivalent resistance. Commercial amplifiers designed for such applications typically have a 100 M $\Omega$  input impedance. This large input impedance will minimize attenuation of the input signal and reduce the Johnson noise current drawn through the source resistance, which can be an important noise source. Field Effect Transistors (FET's) are used in these amplifiers to reduce the input bias current to the amplifiers. Shot noise on the input bias current can be an important noise component, and temperature drift of the input bias current is a source of drift in dc measurements.

The bandwidth of a high input impedance amplifier is often determined by the RC time constant of the source, cable, and termination resistance. For example, a PMT with 1 meter of RG-58 coax (about 100 pF) terminated into a 1 M $\Omega$  resistor will have a bandwidth of about 1600 Hz. A smaller resistance would improve the bandwidth, but increase the Johnson noise current

Bipolar transistors offer an input noise voltage which may be several times smaller than the FET inputs of high input impedance amplifiers, as low as 1 nV/ $\sqrt{\text{Hz}}$ . Bipolar transistors have larger input bias currents, hence larger shot noise current, and so should be used only with low impedance (<1 k $\Omega$ ) sources.

When AC signals from very low source impedances are to be measured, transformer coupling offers very quiet inputs. The transformer is used to step-up the input voltage by its turns-ratio. The transformer's secondary is connected to the input of a bipolar transistor amplifier.

Conventional bipolar and FET input amplifiers exhibit input offset drifts on the order of 5  $\mu\text{V}/^\circ\text{C}$ . In the case where the detector signal is a small dc voltage, such as from a bolometer, this offset drift may be the dominant noise source. A different amplifier configuration, chopper stabilized amplifiers, essentially measures the input offsets and subtracts the measured offset from the signal. A similar approach is used to "auto-zero" the offset on the input to sensitive voltmeters. Chopper stabilized amplifiers exhibit very low input offsets with virtually no input offset drift.

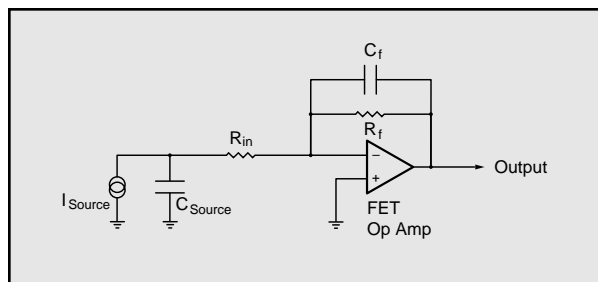
The use of "true-differential" or "instrumentation" amplifiers is advised to provide common mode rejection to interfering noise, or to overcome the difference in grounds between the voltage source and the amplifier. This amplifier configuration amplifies the difference

between two inputs, unlike a single-ended amplifier, which amplifies the difference between the signal input and the amplifier ground. In high frequency applications, where good differential amplifiers are not available or are difficult to use, a balun or common mode choke may be used to isolate disparate grounds.

### Transconductance Amplifiers

When the detector is a current source (or has a large equivalent resistance) then a transconductance amplifier should be considered. Transconductance amplifiers (current-to-voltage converters) offer the potential of lower noise and wider bandwidth than a termination resistor and a voltage amplifier, however, some care is required in their application.

A typical transconductance amplifier configuration is shown in Figure 5. A FET input op amp would be used for its low input bias current. (Op amps with input bias currents as low as 50 fA are readily available.) The detector is a current source,  $I_o$ . Assuming an ideal op amp, the transconductance gain is  $A = V_{\text{out}}/I_{\text{in}} = R_f$ , and, the input impedance of the circuit is  $R_{\text{in}}$  to the op amp's virtual null. ( $R_{\text{in}}$  allows negative feedback, which would have been phase shifted and attenuated by the source capacitance at high frequencies, to assure stability.) Commercial transconductance amplifiers use  $R_f$ 's as large as 10 M $\Omega$ , with  $R_{\text{in}}$ 's which are typically  $R_f/1000$ . A low input impedance will insure that current from the source will not accumulate on the input capacitance.



**Fig. 5: Typical Transconductance Amplifier**

This widely used configuration has several important limitations which will degrade its gain, bandwidth and noise performance. The overall performance of the circuit depends critically on the source capacitance, including that of the cable connecting the source to the amplifier input. Limitations include:

- 1) The "virtual null" at the inverting input to the op amp is approximately  $R_f/A_v$  where  $A_v$  is the op amp's open

loop gain at the frequency of interest. While op amps have very high gain at frequencies below 10 Hz (typically a few million), these devices have gains of only a few hundred at 1 kHz. With an  $R_f$  of 1 G $\Omega$ , the virtual null has an impedance of 5 M $\Omega$  at 1 kHz, hardly a virtual null. If the impedance of the source capacitance is less than the input impedance, then most of the ac input current will go to charging this capacitance, thereby reducing the gain.

2) The configuration provides high gain for the voltage noise at the non-inverting input of the op amp. At high frequencies, where the impedance of the source capacitance is small compared to  $R_{in}$ , the voltage gain for noise at the non-inverting input is  $R_f/R_{in}$ , typically about 1000. As FET input op amps with very low bias currents tend to have high input voltage noise, this term can dominate the noise performance of the design.

3) Large  $R_f$ 's are desired to reduce the Johnson noise current, however large  $R_f$ 's degrade the bandwidth. If low values of  $R_f$  are used, the Johnson noise current can dominate the noise performance of the design.

4) To maintain a flat frequency response, the size of the feedback capacitance must be adjusted to compensate for different source capacitances.

As many undesirable characteristics of the transconductance amplifier can be traced to the source capacitance, a system may benefit from integrating the amplifier into the detector, thereby eliminating interconnect capacitance. This approach is followed in many applications, from microphones to CCD imagers.

### Signal Analysis

A variety of noise sources are avoided by AC measurement of the signal. When making DC measurements, the signal must compete with large low frequency noise sources. However, when the source is modulated, the signal may be measured at the modulation frequency, away from these large noise sources.

When the source is modulated, one may choose from gated integration, boxcar averaging, transient digitizers, lock-in amplifiers, spectrum analyzers, gated photon counters or multichannel scalers.

A measurement of the integral of a signal during a period of time can be made with a gated integrator. Commercial devices allow gates from about 100 ps to several milliseconds. A gated integrator is typically used in a pulsed laser measurement. The device can provide

shot-by-shot data which is often recorded by a computer via an A/D converter. The gated integrator is recommended in situations where the signal has a very low duty cycle, low pulse repetition rate, and high instantaneous count rates.

The noise bandwidth of the gated integrator depends on the gate width: short gates will have wide bandwidths, and so will be noisy. This would suggest that longer gates would be preferred, however, the signal of interest may be very short lived, and using a gate which is much wider than the signal will not improve the S/N.

The gated integrator also behaves as a filter: the output of the gated integrator is proportional to the average of the input signal during the gate, so frequency components of the input signal which have an integral number of cycles during the gate will average to zero. This characteristic may be used to 'notch out' specific interfering signals.

It is often desirable to make gated integration measurements synchronously with an interfering source. (This is the case with time-domain signal detection techniques, and not the case with frequency domain techniques such as lock-in detection.) For example, by locking the pulse repetition rate to the power-line frequency ( or to any sub-multiple of this frequency ) the integral of the line interference during the short gate will be the same from shot-to-shot, which will appear as a fixed offset at the output of the gated integrator.

Shot-by-shot data from a gated integrator may be averaged to improve the S/N. Commercial boxcar averagers provide linear or exponential averaging. The averaged output from the boxcar may be recorded by a computer or used to drive a strip chart recorder. Figure 6 shows a gated integrator with an exponential averaging circuit.

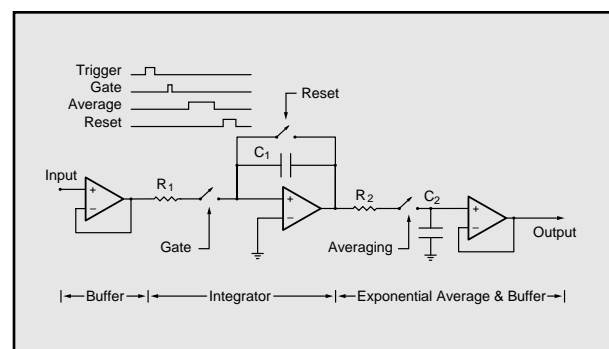


Fig. 6: Gated Integrator and Boxcar Averager

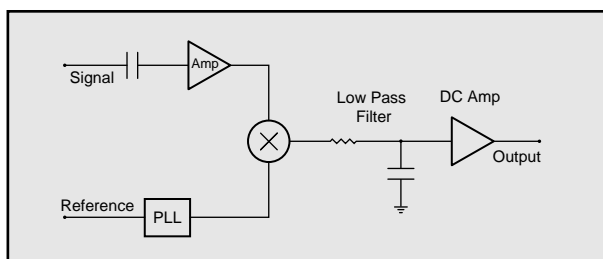
## Lock-In Amplifiers

Phase sensitive synchronous detection is a powerful technique for the recovery of small signals which may be obscured by interference which is much larger than the signal of interest. In a typical application, a CW laser which induces the signal of interest will be modulated by an optical chopper. The lock-in amplifier is used to measure the amplitude and phase of the signal of interest relative to a reference output from the chopper.

Figure 8 shows a simplified block diagram for a lock-in amplifier. The input signal is ac coupled to an amplifier whose output is mixed (multiplied by) the output of a phase-locked loop which is locked to the reference input. The operation of the mixer may be understood through the trigonometric identity:

$$A\cos(\omega_1 t + \phi) \times B\cos(\omega_2 t) = 1/2AB[\cos((\omega_1 + \omega_2)t + \phi) + \cos((\omega_1 - \omega_2)t + \phi)]$$

When  $\omega_1 = \omega_2$  there is a DC component of the mixer output,  $\cos(\phi)$ . The output of the mixer is passed through a low pass filter to remove the sum frequency component. The time constant of the filter is selected to reduce the equivalent noise bandwidth: selecting longer time constants will improve the S/N at the expense of longer response times.



**Figure 8: Lock-In Amplifier Block Diagram**

The simplified block diagram shown in Figure 8 is for a "single phase" lock-in amplifier, which measures the component of the signal at a single phase with respect to the reference. A dual phase lock-in has another channel which measures the component of the signal at 90 degrees relative to the first channel, which allows simultaneous measurement of the amplitude and phase of the signal.

## Digital Signal Processing Techniques

Digital signal processing (DSP) techniques are rapidly

replacing the older analog techniques for the synchronous detection of the signal. In these instruments, the input signal is digitized by a fast, high resolution A/D converter, and the signal's amplitude and phase are determined by high speed computations in a digital signal processor. To maintain the 100 kHz bandwidth of the analog designs, the DSP designs must complete a quarter million 16-bit A/D conversions and 20 million multiply-and-accumulate operations each second. Many of artifacts of the analog designs are eliminated by the DSP approach; for example, the output drift and dynamic range of the instruments are dramatically improved.

## Photon Counting

Photon counting techniques ( which are readily applicable to particle counting measurements) offer several advantages in the measurement of light: very high sensitivity (count rates as low as 1 per minute can be a usable signal level), large dynamic range (signal levels as high as 100 MHz can be counted, allowing a 195 dB dynamic range), discrimination against low level noise (analog noise below the discriminator thresholds will not be counted), and ability to operate over widely varying duty cycles.

Key elements of a photon counting system include: a high gain PMT (or charge multiplier) operated with sufficient high voltage so that a single photoelectron (or charged particle) will generate an anode pulse of several millivolts into a 50  $\Omega$  load, a fast discriminator to generate logic pulses from anode signals which exceed a set threshold, and fast gated counters to integrate the counts.

In situations where the time evolution of a light signal must be measured (LIDAR, lifetime measurements, chemical kinetics, etc.) transient photon counters allow the entire signal to be recorded for each event. In these instruments, the discriminated photon pulses are summed into different bins depending on their timing with respect to a trigger pulse. Commercial instruments offer 5 ns resolution with zero dead-time between bins. The time records from many events may be summed together in order to improve the S/N.

Which instrument is best suited for detecting signals from a photomultiplier tube? The answer is based on many factors, including the signal intensity, the signal's time and frequency distribution, the various noise sources and their time dependence and frequency distribution. In general, the choice between boxcar averaging (gated integration) and lock-in detection (phase

sensitive detection) is based on the time behavior of the signal. If the signal is fixed in frequency and has a 50 % duty cycle, lock-in detection is best suited. This type of experiment commonly uses an optical chopper to modulate the signal at some low frequency. Signal photons occur at random times during the 'open' phase of the chopper. The lock-in detects the average difference between the signal during the 'open' phase and the background during the 'closed' phase.

To use a boxcar averager in the same experiment would require the use of very long, 50 % duty cycle gates since the photons can arrive anywhere during the 'open' phase. Since the gated integrator is collecting noise during this entire gate, the signal is easily swamped by the noise. To correct for this, baseline subtraction can be used where an equal gate is used to measure the background during the 'closed' phase of the chopper and subtracted from the 'open' signal. This is then identical to lock-in detection. However, lock-in amplifiers are much better suited to this, especially at

low frequencies (long gates) and low signal intensities.

If the signal is confined to a very short amount of time, then gated integration is usually the best choice for signal recovery. A typical experiment might be a pulsed laser excitation where the signal lasts for only a short time (100 ps to 1  $\mu$ s) at a repetition rate up to 10 kHz. The duty cycle of the signal is much less than 50 %. By using a narrow gate to detect signal only when it is present, noise which occurs at all other times is rejected. If a longer gate is used, no more signal is measured but the detected noise will increase. Thus, a 50 % duty cycle gate would not recover the signal well and lock-in detection is not suitable.

Photon counting can be used in either the lock-in or the gated mode. Using a photon counter is usually required at very low signal intensities or when the use of a pulse height discriminator to reject noise results in an improved S/N. If the evolution of a weak light signal is to be measured, a transient photon counter or multi-channel scaler can greatly reduce the time required to make a measurement.