

# SIGNAL DETECTION AND ANALYSIS

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## 27.1 GLOSSARY

$A$	dimensionless material constant for $1/f$ noise
$C$	capacitance (farads)
$I$	current (amps)
$I_{\text{shot noise}}$	shot noise current (amps)
$k$	Boltzmann's constant
$q$	electron charge (Coulombs)
$R$	resistance (ohms)
$S/N$	signal-to-noise ratio
$T$	temperature (Kelvin)
$V_{\text{Johnson, rms}}$	RMS Johnson noise voltage (V)
$\Delta f$	bandwidth (Hz)

## 27.2 INTRODUCTION

Many optical systems require a quantitative measurement of light. Applications range from the very simple, such as a light meter using a photocell and a d'Arsenval movement, to the complex, such as the measurement of a fluorescence lifetime using time-resolved photon counting.

Often, the signal of interest is obscured by noise. The noise may be fundamental to the process: photons are discrete quanta 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 pickup.

This chapter describes methods for making useful measurements of weak optical signals, even in the presence of large interfering sources. The chapter will emphasize the electronic aspects of the problem. Important details of optical systems and detectors used in signal recovery are covered in Chap. 24, "Photodetectors," by Paul R. Norton; Chap. 25, "Photodetection," by Abhay M.

Joshi and Gregory H. Olsen; and Chap. 26, “High-Speed Photodetectors,” by John E. Bowers and Yih G. Wey in this volume.

## 27.3 PROTOTYPE EXPERIMENT

Figure 1 details the elements of a typical measurement situation. We wish to measure light from the source of interest. This light may be obscured by light from background sources. The intensity of the source of interest, and the relative intensity of the interfering background, will determine whether some or all of the techniques shown in Fig. 1 should be used.

### Optics

An optical system is designed to pass photons from the source of interest and reject photons from background sources. The optical system may use spatial focusing, wavelength, or polarization selection to preferentially deliver photons from the source of interest to the detector.

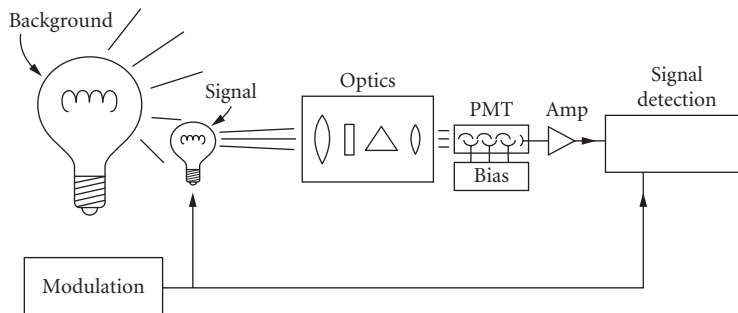
There are many trade-offs to consider when designing an optical system. For example, if the source is nearly monochromatic and the background is broadband, then a monochromator may be used to improve the signal-to-background ratio of the light reaching the detector. However, if the source of interest is an extended isotropic emitter, then a monochromator with narrow slits and high  $f$ -number will dramatically reduce the number of signal photons from the source which can be passed to the detector. In this case, the noise of the detector and amplifiers which follow the optical system may dominate the overall signal-to-noise ratio ( $S/N$ ).

### Photodetectors

There are many types of nonimaging photodetectors. Key criteria to select a photodetector for a particular application include sensitivity for the wavelength of interest, gain, noise, and speed. Important details of many detector types are given in other chapters in the *Handbook*. Operational details (such as bias circuits) of photomultipliers which are specific to boxcar integration and photon counting are discussed in Sec. 27.5.

### Amplifiers

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.



**FIGURE 1** Prototypical optical measurement.

## 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 modulation is inherent to the measurement. For example, when a pulsed laser is used to induce a fluorescence, 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 that modulation is unnecessary.

## 27.4 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 a photodetector may increase current noise by several orders of magnitude.<sup>1</sup>

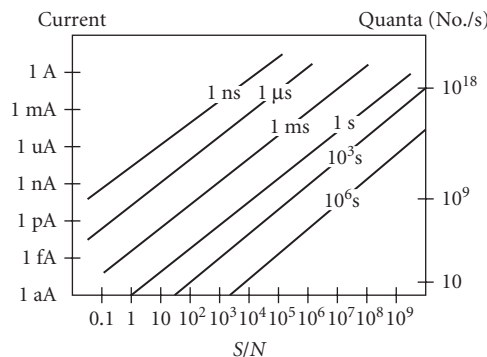
### Shot Noise

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, its 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.

“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

$$I_{\text{shot noise}} = \sqrt{(2qI\Delta f)} \quad (1)$$

where  $q = 1.6 \times 10^{-19}$  C



**FIGURE 2** Signal-to-noise versus flux and measurement time.

### 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  $R$  ohms, at a temperature of  $T$  Kelvin, in a bandwidth of  $\Delta f$  Hz is given by

$$V_{\text{Johnson,rms}} = \sqrt{4kTR\Delta f} \quad (2)$$

where  $k$  is Boltzmann's constant. The noise voltage in a 1-Hz bandwidth is given by

$$V_{\text{Johnson,rms}} (\text{per } \sqrt{\text{Hz}}) = 0.13 \text{ nV} \times \sqrt{R(\text{ohms})} \quad (3)$$

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- $\mu\text{V}$  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

$$I_{\text{Johnson,rms}} (\text{per } \sqrt{\text{Hz}}) = 130 \text{ pA} / \sqrt{R(\text{ohms})} \quad (4)$$

As 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

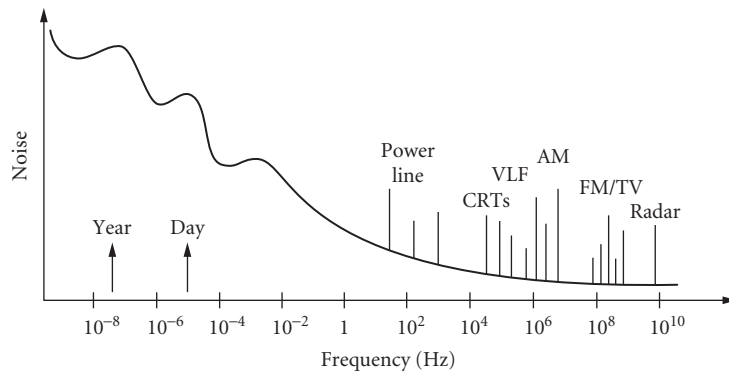
The voltage across a resistor carrying a constant current will fluctuate because the resistance of the material used in the resistor varies. The magnitude of the resistance fluctuation depends on the material used: carbon composition resistors are the worst, metal film resistors are better, and wire wound resistors provide the lowest 1/f noise. The rms value of this noise source for a resistance of  $R$  ohms, at a frequency of  $f$  Hz, in a bandwidth of  $\Delta f$  Hz is given by

$$V_{1f,\text{rms}} = IR \times \sqrt{A\Delta f/f} \quad (5)$$

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 ( $IR$ ), the  $S/N = 3 \times 10^5 \sqrt{(f/\Delta f)}$ . Often, this noise is a troublesome source of low-frequency noise in voltage amplifiers.

### Nonessential 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 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



**FIGURE 3** Simplified noise spectrum.

(AM, FM, VHF, and UHF TV), special services (cellular telephones, pagers, etc.), microwave ovens and communications, to RADAR and beyond.

Your best alternatives for avoiding these noise sources are

1. Shield to reduce pickup.
2. Use differential inputs to reject common mode noise.
3. Bandwidth limit the amplifier to match expected signal.
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 Fig. 4a to f.

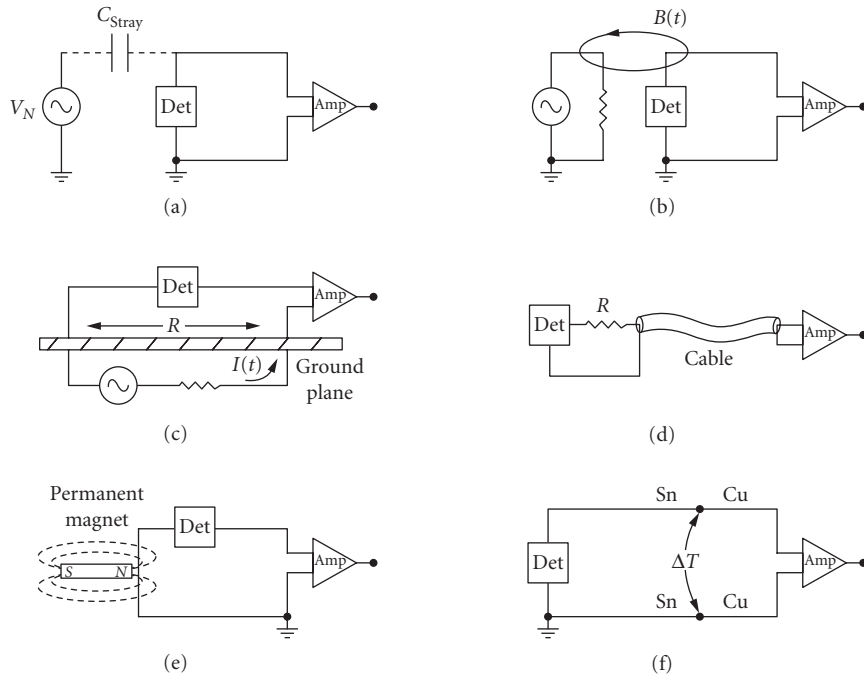
Noise may be injected via a stray capacitance as in Fig. 4a. The stray capacitance has an impedance of  $1/j\omega C$ . Substantial currents may be injected into low-impedance systems (such as transconductance inputs), or large voltages may appear at the input to high-impedance systems.

Inductive pickup is illustrated in Fig. 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 pickup 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 nonmagnetic 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 Fig. 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 Fig. 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 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 Fig. 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.



**FIGURE 4** Coupling of noise sources.

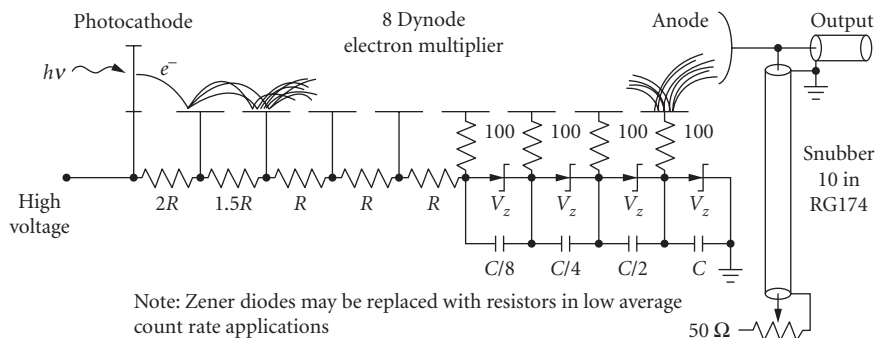
Unwanted thermocouple junctions are an important source of offset and drift. As shown in Fig. 4f, two thermocouple junctions are formed when a signal is connected to an amplifier. For typical interconnect materials (copper, tin) one sees about  $10 \mu\text{V}/^\circ\text{C}$  of offset. These extraneous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

## 27.5 APPLICATIONS USING PHOTOMULTIPLIERS

Photomultiplier tubes (PMTs) are used for detection of light from about 200 to 900 nm. Windowless PMTs can be used from the near UV through the x-ray region, and may also be used as particle detectors. Their low noise, high gain, wide bandwidth, and large dynamic range have placed them in many applications. They are the only detectors which may be recommended for low-noise photon counting applications.<sup>2,3</sup>

In this chapter, we are primarily concerned with the electrical characteristics of PMTs. Understanding these characteristics is important if we are to realize the many desirable features of these devices.

A schematic representation of a PMT, together with a typical bias circuit, is shown in Fig. 5. While the concepts depicted here are common to all PMTs, the particulars of biasing and termination will change between PMT types and applications. PMTs have a photocathode, several dynodes (6 to 14), and an anode. They are usually operated from a negative high voltage, with the cathode at the most negative potential, each successive dynode at a less negative potential, and the anode near ground. An incident photon may eject a single photoelectron from the photocathode which will strike the first dynode with an energy of a few hundred volts. A few (2– to 5) electrons will be ejected from the first dynode by the impact of the photoelectron: these electrons will in turn strike



**FIGURE 5** PMT base for photon counting or fast integration.

the second dynode, ejecting more electrons. The process continues at each dynode until all of the electrons are collected by the anode.

## Quantum Efficiency

The quantum efficiency (QE) of a PMT is a measure of the probability that a photon will eject a photoelectron at the photocathode. The QE depends on the type of material used in the cathode and the wavelength of light. QEs may be as high as 10 to 30 percent at their peak wavelength. The cathode material will also affect the dark count rate from the PMT: a cathode with good red sensitivity may have a high dark count rate.

## Gain

A PMT's gain depends on the number of dynodes, the dynode material, and voltage between the dynodes. PMT gains range from  $10^3$  to  $10^7$ . The anode output from the PMT will typically go to an electronic amplifier. To avoid having the system noise be dominated by the amplifier's noise, the PMT should be operated with enough gain so that the dark current times the gain is larger than the amplifier's input current noise.

## Bandwidth

The frequency response, speed, rise time, and pulse-pair resolution of PMTs depend on the structure of the dynode multiplier chain. The leading edges of the anode output have transition times from 2 to 20 ns. Trailing edges are usually about three times slower. Much faster PMTs, with rise times on order 100 ps, use microchannel plate multipliers.

When using gated integrators to measure PMT outputs, the pulse width of the anode signal should be less than the gate width so that timing information is not lost. For photon counting, the pulse width should be smaller than the pulse-pair resolution of the counter/discriminator to avoid saturation effects. When using lock-in amplifiers, pulse width is usually not important, since the slowest PMTs will have bandwidths well above the modulation frequency.

## Pulse Height

In pulsed experiments, the criterion for a detectable signal often depends on the electrical noise environment of the laboratory and the noise of the preamplifier. In laboratories with Q-switched lasers or pulsed discharges, it is difficult to reduce the noise on any coaxial cable below a few millivolts.

A good, wide bandwidth preamplifier will have about  $1.5 \text{ nV}/\sqrt{\text{Hz}}$ , or about  $25\text{-}\mu\text{V}$  rms over a 300-MHz bandwidth. Peak noise will be about 2.5 times the rms noise, so it is important that the PMT provide pulses of greater than 1-mV amplitude.

Use manufacturer's specifications for the current gain and rise time to estimate the pulse amplitude from the PMT:

$$\text{Amplitude (mV)} = 4 \times \text{gain (millions)} / \text{rise time (ns)} \quad (6)$$

This formula assumes that the electrons will enter a  $50\text{-}\Omega$  load in a square pulse whose duration is twice the rise time. (Since the rise time will be limited by the bandwidth of the preamplifier, use the larger of the amplifier or PMT rise times in this formula.)

If the PMT anode is connected via a  $50\text{-}\Omega$  cable to a large load resistance, then the pulse shape may be modeled by the lumped parameters of the cable capacitance (about 100 pF/meter for RG-58) and the termination resistance. All of the charge in the pulse is deposited on the cable capacitance in a few nanoseconds. The voltage on the load will be  $V = Q/C$  where  $C$  = cable capacitance. This voltage will decay exponentially with a time constant of  $RC$  where  $R$  is the load resistance in ohms. In this case, the pulse height will be

$$\text{Amplitude (mV)} = 160 \times \text{gain (millions)} / \text{cable } C \text{ (pF)} \quad (7)$$

The current gain of a PMT is a strong function of the high voltage applied to the PMT. Very often, PMTs will be operated well above the high voltage recommended by the manufacturer, and thus substantially higher current gains ( $10\times$  to  $100\times$  above specs). There are usually no detrimental effects to the PMT as long as the anode current is kept well below the rated value.

## Dark Counts

PMTs are the quietest detectors available. The primary noise source is thermionic emission of electrons from the photocathode and from the first few dynodes of the electron multiplier. PMT housings which cool the PMT to about  $-20^\circ\text{C}$  can dramatically reduce the dark counts (from a few kHz to a few Hz). The residual counts arise from radioactive decays of materials inside the PMT and from cosmic rays.

PMTs which are specifically designed for photon counting will specify their noise in terms of the rate of output pulses whose amplitudes exceed some fraction of a pulse from a single photon. More often, the noise is specified as an anode dark current. Assuming the primary source of dark current is thermionic emission from the photocathode, the dark count rate is given by

$$\text{Dark count (kHz)} = 6 \times \text{dark current (nA)} / \text{gain (millions)} \quad (8)$$

## PMT Base Design

PMT bases which are designed for general-purpose applications are not appropriate for photon counting or fast-gated integrator applications (gates  $< 10$  to  $20 \text{ ns}$ ). General-purpose bases will not allow high count rates, and often cause problems such as double counting and poor plateau characteristics. A PMT base with the proper high-voltage taper, bypassing, snubbing, and shielding is required for good time resolution and best photon counting performance.

**Dynode Biasing** A PMT base provides bias voltages to the PMTs photocathode and dynodes from a single, negative, high-voltage power supply. The simplest design consists of a resistive voltage divider. In this configuration the voltage between each dynode, and thus the current gain at each dynode, is the same. Typical current gains are three to five, so there will typically be four electrons leaving the first dynode, with a variance of about two electrons. This large relative variance (due to the small



number of ejected electrons) gives rise to large variations in the pulse height of the detected signal. Since statistical fluctuations in pulse height are dominated by the low gain of the first few stages of the multiplier chain, increasing the gain of these stages will reduce pulse-height variations and so improve the pulse-height distribution. This is important for both photon counting and analog detection. To increase the gain of the first few stages, the resistor values in the bias chain are increased to increase the voltage in the front end of the multiplier chain. The resistor values are tapered slowly so that the electrostatic focusing of electrons in the multiplier chain is not adversely affected.<sup>4</sup>

Current for the electron multiplier is provided by the bias network. Current drawn from the bias network will cause the dynode potentials to change, thus changing the tube gain. This problem is of special concern in lifetime measurements. The shape of exponential decay curves will be changed if the tube gain varies with count rate. To be certain that this is not a problem, lifetime measurements should be repeated at reduced intensity. The problem of gain variation with count rate is avoided if the current in the bias network is about 20 times the output current from the PMT's anode.

There are a few other methods to avoid this problem which do not require high bias currents. These methods depend on the fact that the majority of the output current is drawn from the last few dynodes of the multiplier:

1. Replace the last few resistors in the bias chain with Zener diodes. As long as there is some reverse current through a Zener, the voltage across the diodes is nearly constant. This will prevent the voltage on these stages from dropping as the output current is increased.
2. Use external power supplies for the last few dynodes in the multiplier chain. This approach dissipates the least amount of electrical power since the majority of the output current comes from lower-voltage power supplies. However, it is the most difficult to implement.
3. If the average count rate is low, but the peak count rate is high, then bypass capacitors on the last few stages may be used to prevent the dynode voltage from dropping (use  $20\times$  the average output current for the chain current). For a voltage drop of less than 1 percent, the stored charge on the last bypass capacitor should be  $100\times$  the charge output during the peak count rate. For example, the charge output during a 1-ms burst of a 100-MHz count rate, each with an amplitude of 10 mV into  $50\ \Omega$  and a pulse width of 5 ns, is  $0.1\ \mu\text{C}$ . If the voltage on the last dynode is 200 Vdc, then the bypass capacitor for the last dynode should have a value given by

$$C = 100Q/V = 100 \times 0.15\text{C}/200\text{V} = 0.05\ \mu\text{F} \quad (9)$$

The current from higher dynodes is smaller so the capacitors bypassing these stages may be smaller. Only the final four or five dynodes need to be bypassed, usually with a capacitor which has half the capacitance of the following stage. To reduce the voltage requirement for these capacitors, they are usually connected in series.

Bypassing the dynodes of a PMT may cause high-frequency ringing of the anode output signal. This can cause multiple counts for a single photon or poor time resolution in a gated integrator. The problem is significantly reduced by using small resistors between the dynodes and the bypass capacitors.

**Snubbing** Snubbing refers to the practice of adding a network to the anode of the PMT to improve the shape of the output pulse for photon counting or fast-gated integrator applications. This "network" is usually a short piece of  $50\text{-}\Omega$  coax cable which is terminated into a resistor of less than  $50\ \Omega$ . The snubber will delay, invert, and sum a small portion of the anode signal to itself. Snubbing should not be used when using a lock-in amplifier since the current conversion gain of a  $50\text{-}\Omega$  resistor is very small.

There are four important reasons for using a snubber network:

1. Without some dc resistive path between the anode and ground, anode dark current will charge the signal cable to a few hundred volts (last dynode potential). When the signal cable is connected to an amplifier, the stored charge on the cable may damage the front end of the instrument. PMT bases without a snubber network should include a  $100\text{-M}\Omega$  resistor between the anode and ground to protect the instruments.

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2. The leading edge of the output current pulse is often much faster than the trailing edge. A snubber network may be used to sharply increase the speed of the trailing edge, greatly improving the pulse pair resolution of the PMT. This is especially important in photon counting applications.
3. Ringing (with a few nanoseconds period) is very common on PMT outputs. A snubber network may be used to cancel these rings which can cause multiple counts from a single photon.
4. The snubber network will help to reverse terminate reflections from the input to the preamplifier.

The round-trip time in the snubber cable may be adjusted so that the reflected signal cancels anode signal ringing. This is done by using a cable length with a round-trip time equal to the period of the anode ringing.

**Cathode Shielding** Head-on PMTs have a semitransparent photocathode which is operated at negative high voltage. Use care so that no objects near ground potential contact the PMT near the photocathode.

**Magnetic Shielding** Electron trajectories inside the PMT will be affected by magnetic fields. A field strength of a few gauss can dramatically reduce the gain of a PMT. A magnetic shield made of a high permeability material should be used to shield the PMT.

#### *PMT Base Summary*

1. Taper voltage divider for higher gain in the first stages.
2. Bypass last few dynodes in pulsed applications.
3. Use a snubber circuit to shape the output pulse for photon counting or fast-gated integration.
4. Shield the tube from electrostatic and magnetic fields.

## 27.6 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.

### Voltage Amplifiers

**High Bandwidth** Photon 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 coaxial cables into their characteristic impedance. When PMTs (which are current sources) are connected to those 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 a lot of current noise.<sup>5</sup>

**High Input Impedance** 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 (FETs) 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.<sup>6</sup>

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.

**Moderate Input Impedance** 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  $1 \text{ nV}/\sqrt{\text{Hz}}$ . Bipolar transistors have larger input bias currents, hence larger shot noise current, and so should be used only with low-impedance ( $<1 \text{ k}\Omega$ ) sources.

**Transformer Inputs** 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.

**Low Offset Drift** Conventional bipolar and FET input amplifiers exhibit input offset drifts on the order of  $5 \text{ }\mu\text{V}/\text{C}$ . 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 measure their input offsets and subtract the measured offset from the signal. A similar approach is used to "autozero" the offset on the input to sensitive voltmeters. Chopper-stabilized amplifiers exhibit very low input offsets with virtually no input offset drift.

**Differential** 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, 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. These 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.<sup>7</sup>

A typical transconductance amplifier configuration is shown in Fig. 6. An 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

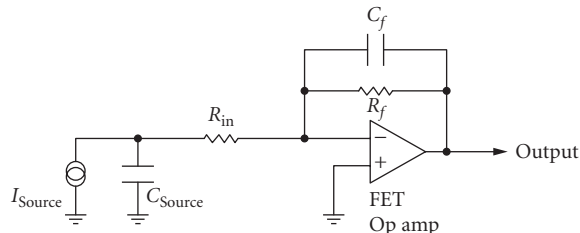


FIGURE 6 Typical transconductance amplifier.

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source capacitance at high frequencies, to assure stability.) Commercial transconductance amplifiers use  $R_s$ 's as large as  $10\text{ M}\Omega$ , with  $R_{in}$ 's, which are typically  $R_f/1000$ . A low-input impedance will ensure that current from the source will not accumulate on the input capacitance.

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  $1\text{ G}\Omega$ , the virtual null has an impedance of  $5\text{ M}\Omega$  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 noninverting 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 noninverting 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.

## 27.7 SIGNAL ANALYSIS

### Unmodulated Sources

For unmodulated sources, a strip-chart recorder, voltmeter,  $A/D$  converter, or oscilloscope may be used to measure the output of the amplifier or detector. In the case of low-light-level measurement, continuous photon counting would be the method of choice.

A variety of problems are avoided by modulating the signal source. 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.

### Modulated 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 scalars.

**Gated Integration** 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.<sup>8</sup>

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 submultiple 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.

**Boxcar Averaging** 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 7 shows a gated integrator with an exponential averaging circuit.

**Lock-In Amplifiers** Phase-sensitive synchronous detection is a powerful technique for the recovery of small signals which may be obscured by interference that 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.<sup>9</sup>

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

$$\cos(\omega_1 t + \Phi) * \cos(\omega_2 t) = \frac{1}{2} \{ \cos[(\omega_1 + \omega_2)t + \Phi] + \cos[(\omega_1 - \omega_2)t + \Phi] \} \quad (10)$$

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.

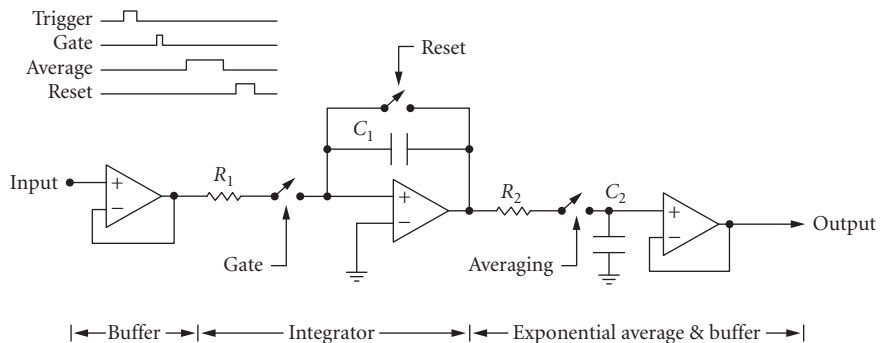
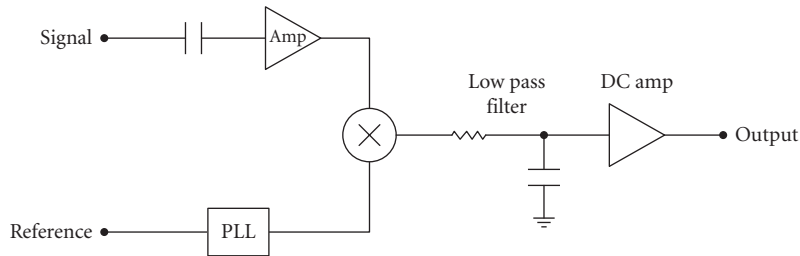


FIGURE 7 Gated integrator and exponential averager.



**FIGURE 8** Lock-in amplifier block diagram.

The simplified block diagram shown in Fig. 8 is for a “single-phase” lock-in amplifier, which measures the component of the signal at one set phase with respect to the reference. A dual-phase lock-in has another channel which measures the component of the signal at  $90^\circ$  relative to the first channel, which allows simultaneous measurement of the amplitude and phase of the signal.

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 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.<sup>10</sup>

**Photon Counting** Photon counting techniques 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.<sup>11</sup>

Key elements of a photon counting system include a high-gain PMT operated with sufficiently high voltage so that a single photoelectron 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.

**Transient Photon Counting** 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  $S/N$ .<sup>12</sup>

**Choosing the “Best” Technique** Which instrument is best suited for detecting signals from a photo-multiplier 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 percent 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 percent 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 of up to 10 kHz. The duty cycle of the signal is much less than 50 percent. 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 percent 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 multichannel scaler can greatly reduce the time required to make a measurement.

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