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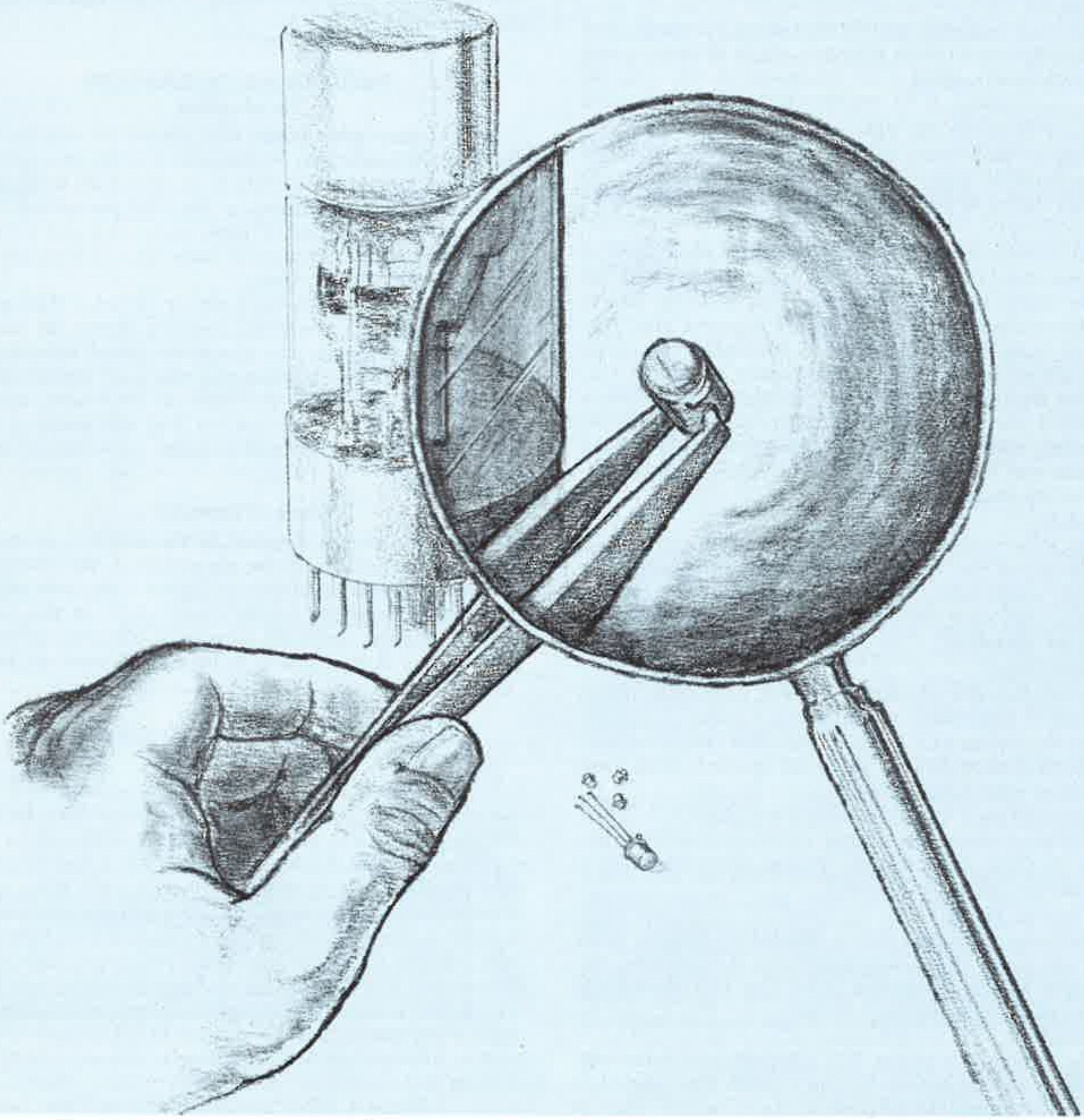
APPLICATION NOTE 915

THRESHOLD DETECTION OF VISIBLE AND INFRARED RADIATION WITH PIN PHOTODIODES



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Printed in U.S.A.

Traditionally, the detection and demodulation of extremely low level optical signals has been performed with multiplier phototubes. Because of this tradition, solid-state photodetectors are often overlooked even though they have a number of clear functional advantages and in some applications provide superior performance as well. Some of these advantages are summarized below and become even more apparent in the following discussion.

ADVANTAGES OF PIN PHOTODIODES VERSUS MULTIPLIER PHOTOTUBES

1. Size and weight:
   PIN photodiodes are approximately three orders of magnitude smaller and lighter. This greatly simplifies and reduces the cost of mounting.
2. Power Supply:
   Multiplier phototubes require more than 1000 volts, which must be precisely regulated and divided among the dynodes. By comparison, PIN photodiodes and associated amplifiers operate stably on less than 20 volts, which does not require precise regulation.
3. Cost:
   The cost, including that of the necessary amplifier, is lower for the PIN photodiode because of lower power supply requirements.
4. Spectral Response:
   Broad skirts of the PIN photodiode make it useful from the ultra-violet, through the visible, and well into the infrared region. This exceeds the range of any other device of comparable sensitivity.
5. Sensitivity:
   Noise equivalent power of the PIN photodiode is lower than that of any other type of photodetector. The signal levels are extremely low, however, and to achieve low level performance they require a high gain, high input resistance amplifier. Multiplier phototubes have built-in gain and do not require additional low-noise amplification. Moreover, the high input resistance needed for sensitive performance precludes fast response, whereas the response time of multiplier phototubes may be in the nanosecond region even in the sensitive mode.
6. Stability:
   The characteristics of noise, responsivity, and spectral response of the PIN photodiode are not dependent on time, temperature, or other environmental considerations. The same conditions may be hazardous to multiplier phototubes.
7. Overloading:
   In the presence of excessive signal, multiplier phototubes of comparable sensitivity are capable of destroying themselves as a result of excessive output current. The PIN photodiode is unaffected by exposure to room light or even direct sunlight.
8. Ruggedness:
   PIN photodiodes can tolerate exposure to extreme levels of shock and vibration. Typical shock capability is 1500 G's for 0.5 millisecond.
9. Magnetic Fields:
   Multiplier phototube gain is affected by fields as small as one gauss. If the interfering field is fluctuating, the output will be modulated by it. The PIN photodiode is insensitive to magnetic fields.
10. Precision:
    The responsivity of the PIN photodiode is inherently precise and repeatable. Within a given type, the characteristics agree (from unit to unit) within plus or minus 0.1 decade. Responsivity of multiplier phototubes may vary over more than a decade from one unit to another.
11. Sensitive Area:
    The small sensitive area of the PIN photodiode makes it unnecessary to establish an aperture which may be required for some applications. However, in some applications good optical alignment is imposed by the small area.

PIN PHOTODIODE DETECTORS

At the present time a variety of different types of solid-state photodetectors are available. Of these, the Silicon PIN Photodiode has the broadest applicability and is the subject of this note. The PIN photodiode's main advantages are: broad spectral response, a wide dynamic range, high speed, and extremely low noise. With appropriate terminal circuits it is well suited for many applications that require converting an optical signal to an electrical signal. The present discussion, however, will be limited to the description of the PIN photodiode's threshold detection sensitivity and the design of suitable terminal circuits that will realize this capability.

PHOTODIODE DESCRIPTION

Construction

A brief description of the PIN photodiode will be helpful in understanding its performance and the principles for designing appropriate circuits to be used with it. Figure 1 shows a typical construction of the PIN photodiode. This figure is for the purpose of explanation only and is not to scale. The relative proportions have been deliberately distorted for the sake of clarity.

The PIN structure is produced by diffusion through an oxide (S_iO_2) mask which also serves to protect the surface. Since most metals are very opaque to optical radiation, especially at infrared wavelengths, the gold contact is deposited only around the perimeter of the P-layer, and the gold contact pattern provides for lead attachment a short distance away from the junction region, so the lead is not in the light path.

Mode of Operation

When a photon is absorbed by the silicon it produces a hole and an electron. If the absorption of the photon occurs in the I-layer, as shown in Figure 1, the hole and the electron are separated by the electric field in the I-layer. For the highest quantum conversion efficiency (electrons per photon) it is desirable to have the P-layer as thin as

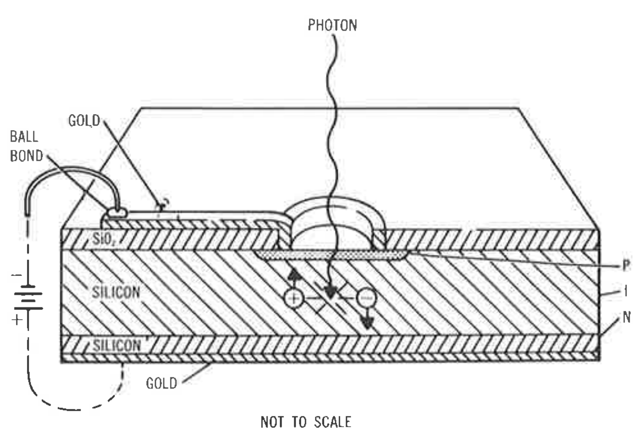


Figure 1. PIN Photodiode Cutaway View

possible and the I-layer as thick as possible. The thickness of the P-layer also determines the value of the parasitic series resistance (R_S in Figure 2). The thinner the P-layer the higher the R_S. Since R_S affects high frequency performance there is therefore a design trade-off between quantum efficiency and bandwidth. Once the trade-off is settled, the desired thickness is then controlled during the diffusion process. The effective thickness of the I-layer is controlled partly by the manufacturing diffusion process and partly by the magnitude of the electric field applied to the diode—the higher the field, the thicker will be the effective I-layer. It is therefore desirable to operate the diode with an external reverse bias, as shown in Figure 2. As the reverse bias voltage is increased from zero, there are three beneficial effects: hole and electron transit time decreases; conversion efficiency increases slightly; and most importantly, the capacitance decreases sharply with bias up to about ten volts and continues to decrease slightly up to about twenty volts reverse bias.

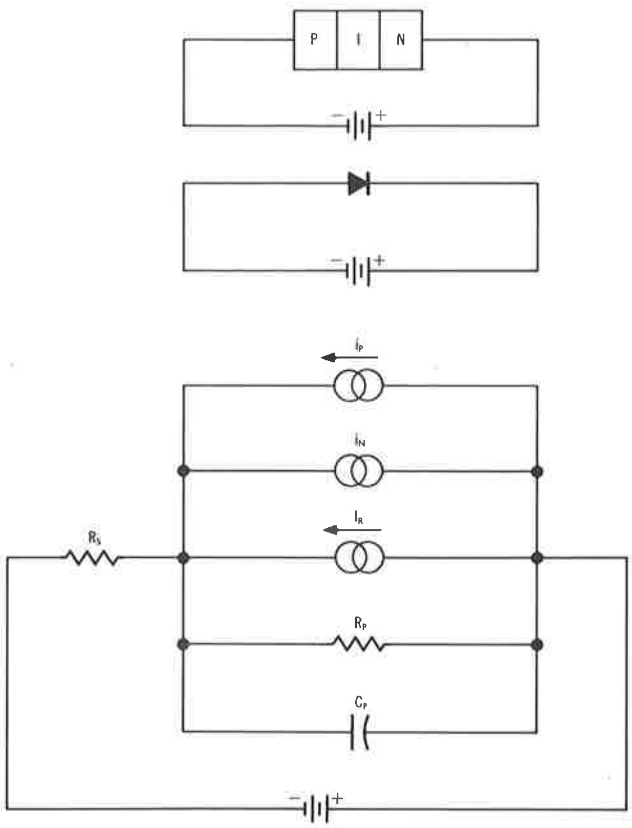


Figure 2. PIN Photodiode Schematic Symbol, and Equivalent Circuit

In the presence of optical signals there is a slight modulation of the shunt conductance as the presence of photon-produced holes and electrons in the I-layer modulate its conductivity. This effect can be quite significant at very high levels of illumination since the I-layer may become saturated, resulting in a decrease in quantum efficiency and an increase in rise time. Saturation can be prevented by applying a very high reverse bias voltage (up to 200 volts). However, such a high voltage, applied over a long period of time, may cause a degradation of the diode's leakage properties. Since our present concern is with threshold performance, reverse bias voltages greater than twenty volts need not be considered.

Equivalent Circuit

When properly biased, the PIN photodiode can be accurately represented by the equivalent circuit shown in Figure 2. Here i_p is the external current resulting when the diode is illuminated. It has a time constant of 10 picoseconds and a value of approximately 0.5 amp per watt of input at a wavelength of 8000 angstroms (800 nanometers). This corresponds to a quantum efficiency of 75%, that is, 0.75 electrons per photon. The quantum efficiency is constant from 500 nanometers to 800 nanometers (5,000 Å to 8,000 Å).

i_N is the noise current of the PIN photodiode. Since the diode is reverse biased, the shot noise formula is applicable, so that the noise current can be computed from:

$$\frac{i_N^2}{B} = 2qI_{dc} \qquad (1)$$

where B = system output bandwidth, Hz
q = electron charge, 1.6×10^{-19} coulombs
I_{dc} = dc current, Amp.

In the case of the photodiode, I_{dc} is simply the dark current, I_R, which has a value determined by the construction and dimensions of the particular diode type. Maximum values are: 100 picoamps for HPA 4204, 150 picoamps for HPA 4205 and 2 nanoamps for HPA 4203.

Shunt resistance, R_p, is very large, being usually greater than 10 gigaohms (10,000 megohms), and its noise current may therefore be neglected. Shunt capacitance, C_p, has a value from two to five picofarads, depending upon the diode type and reverse bias. For high frequency operation it is important to minimize C_p because the cutoff frequency is determined by:

$$f_c = \frac{1}{2\pi R_s C_p} \qquad (2)$$

Although our present concern is with low frequency threshold operation, there is another reason for minimizing C_p. This will be discussed later, when circuit design principles are presented.

Performance

Threshold performance can and has been specified in a number of different ways. The most commonly understood and usable expression takes the form of a noise equivalent input signal. This is the input signal which produces an output signal level that is equal in value to the noise level that is present when no input signal is applied. The noise equivalent input in watts is called Noise Equivalent Power (NEP) and is defined by:

$$\text{NEP} = \frac{\text{NOISE CURRENT (amps per root hertz)}}{\text{CURRENT RESPONSIVITY (amps per watt)}} \qquad (3)$$

which has the units of watts per root hertz. Devices for photo-detection could then be compared on the basis of NEP. The lower the NEP the more sensitive is the device.

Another method of defining threshold sensitivity is on the basis of signal-to-noise ratio for given input signal power levels. Taking a power level of one picowatt, for example, the signal-to-noise ratio at the output can be obtained from:

$$\text{SNR} = \frac{\text{RESPONSIVITY} \left(\frac{\text{amps}}{\text{watts}}\right) \times \text{INPUT (watts)}}{\text{NOISE CURRENT (amps)}} \qquad (4)$$

This is a ratio of currents. To express it in dB we would take twenty times its log to base ten, even though the expression converts linearly to a power ratio. This is because the devices respond *linearly* to input *power*.

Figure 3 shows spectral sensitivity characteristics of several PIN photodiodes and multiplier phototubes. Sensitivity is given in terms of SNR and NEP. The latter is in terms of dBm. Several interesting features are evident in Figure 3. Although the quantum efficiency for PIN photodiodes is constant from 500 to 800 nanometers, the sensitivity curve is not. This is due to the fact that the energy per quantum (photon) of radiant energy varies with wavelength.

The curves for the three different PIN photodiodes also show the dependence of sensitivity on leakage current. Here the highest sensitivity is obtained with the HPA 4204 which has a maximum leakage current of 100 picoamps. Next is the HPA 4205 with 150 picoamps and finally the HPA 4203 with maximum leakage of 2 nanoamps. The three curves are in effect displaced by the magnitude of the noise current difference because quantum efficiency is equal for all. These curves also show the inherent broad response of PIN photodiodes with respect to multiplier phototubes. Therefore, the power responsivity of the PIN photodiode has a corresponding slope. Notice how the inherently broad response of silicon, enhanced by the thick I-layer construction, extends the range of useful performance over the response ranges of two types of photocathodes.

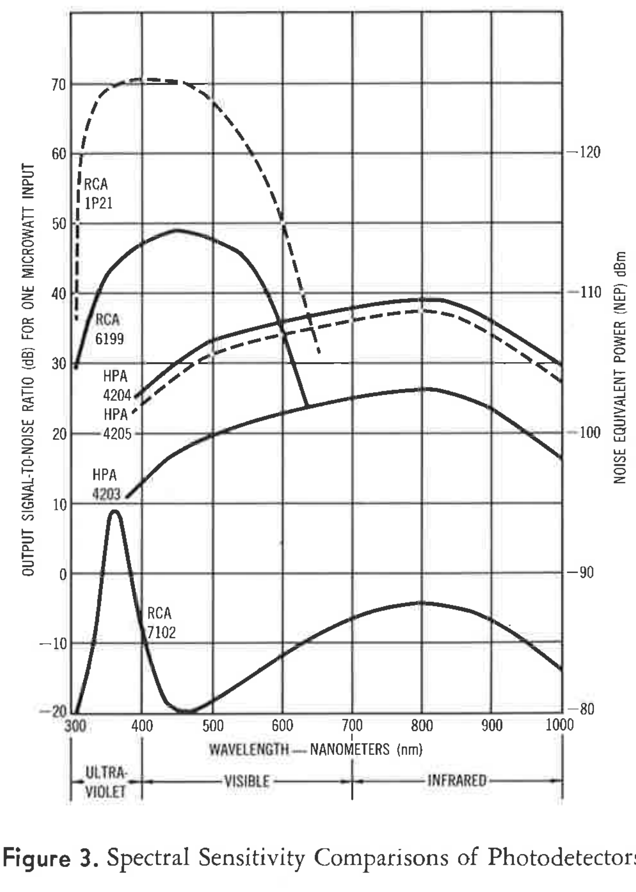


Figure 3. Spectral Sensitivity Comparisons of Photodetectors

Although the threshold sensitivity of multiplier phototubes is superior in the visible region, nevertheless for many applications the advantage is not significant enough to outweigh the disadvantages of generally unstable and temperature-sensitive gain, large size and weight, and the need of very high and stable power supply voltages. On the other hand, the superior red and infrared threshold performance of the PIN photodiode does not necessarily mean it is better in any application, because one must take into account its small sensitive area and low signal levels. Realization of the performance capability described in Figure 3 also requires fairly careful attention to the design of the terminal circuits into which the PIN photodiode operates.

TERMINAL CIRCUIT DESIGN PRINCIPLES

The design of the terminal amplifier must consider the usual design objectives of low noise, broad band, wide dynamic range, etc. In addition, there are two fundamental considerations which are dictated by the PIN photodiode:

1. High Reverse Voltage:
   The diode must be operated at ten to twenty volts of reverse bias to reduce shunt capacitance.
2. High Input Resistance:
   This is a fundamental consideration in the sensitivity/rise time trade-off.

The effects of reverse voltage on capacitance have been discussed earlier. However, the effect is sufficiently important to deserve a re-emphasis here.

A high input resistance is necessary in order to maintain a high signal-to-noise ratio. Since the output signal from the photodiode is a current, and its own internal noise is represented by a current, it is appropriate to represent the noise of the terminal amplifier as an equivalent noise current at the input. The smallest value of resistor which may be connected to the input is then limited by its noise current according to the formula for thermal noise:

$$\frac{i_N^2\ \text{(thermal)}}{B} = \frac{4kT}{R} \qquad (5)$$

By comparing eq(1), relating diode noise current to leakage current, with eq(5), relating resistor noise current to its resistance value, it is clear that there is some value of resistance below which the NEP of the system, i.e., threshold sensitivity, would be degraded at the rate of 5 dB per decade of decreasing resistance. For example, in the case of the HPA 4203, assuming a maximum leakage current of 2 nanoamps, the value of resistance should be greater than 25 megohms, to avoid degrading the threshold sensitivity.

TRANSISTOR AMPLIFIER

In addition to keeping the input noise current low by using large values of input resistance, it is also important to keep other sources of noise in the amplifier at a minimum. Using ordinary transistors (PNP or NPN) it is not possible to approach the ultimate sensitivity of which the PIN photodiode alone is capable, even when low-noise transistors, such as the 2N2484, are used. However, in those applications where it is possible to sacrifice sensitivity for simplicity, transistors may be used. A typical transistor circuit is shown in Figure 4. With this circuit, a sensitivity corresponding to an NEP of — 95 dBm was obtained. In this case, Q1 was operated at the lowest possible collector current which would still give adequate gain. A high loop gain was desired in order to compensate, with negative feedback, for the long open-loop rise time produced by the high input resistance. A resistance higher than 10 megohms was not necessary here, since the transistor itself sets the fundamental noise limitation. A PNP transistor was selected for Q2 in order to balance out most of the base-to-emitter voltage of Q1, so that the output would tend to be near zero without any zero adjustment. A slight zero adjustment, provided by R2 and R3,

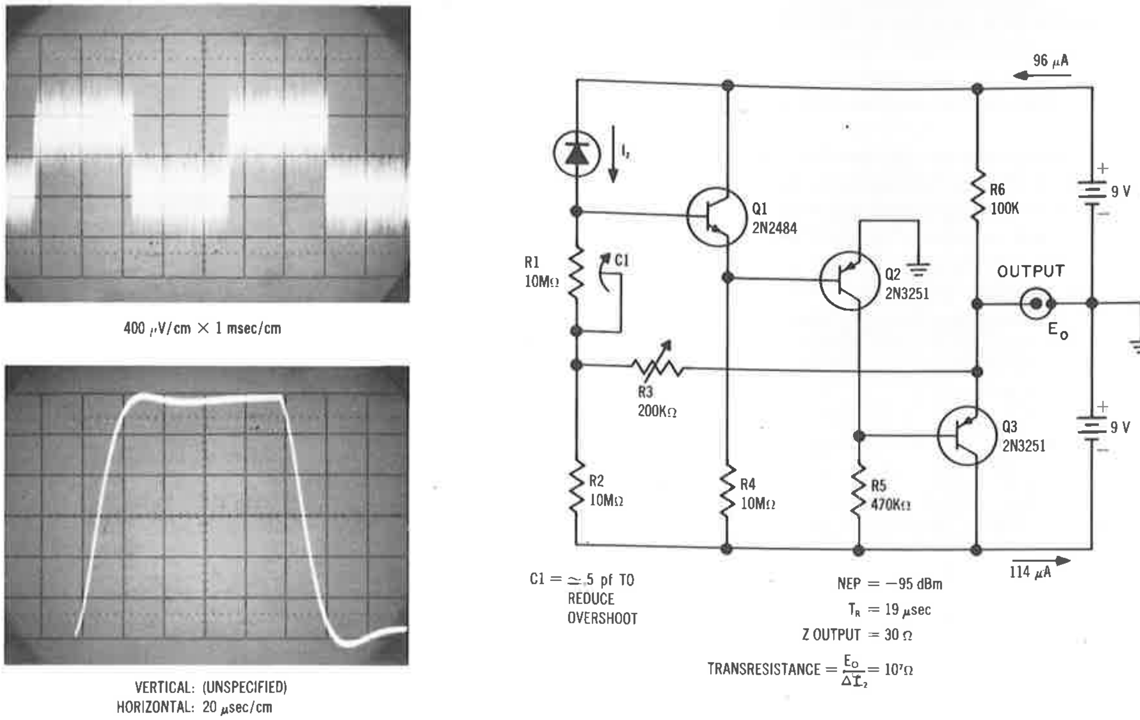


Figure 4. Transistor Photodiode Amplifier Schematic

gives the necessary range without appreciably attenuating the feedback current. As the photocurrent, I_2, increases, the amplifier causes the voltage at the emitter of Q3 to decrease, which causes a current in R1 to flow out of the node (base of Q1) into which I_2 flows.

LEAD NETWORK COMPENSATION

Negative feedback is not the only way to compensate for the low cutoff frequency imposed by high input resistance. The important thing is to preserve the signal-to-noise ratio. An ordinary lead network at the output can be used to compensate for the gain slope of the photodiode/amplifier system having low cutoff frequency. An example of such a network is shown in Figure 5. A word of caution here: There may be considerable attenuation in the lead network, but the signal level must not be allowed to fall so low that the signal-to-noise ratio is affected. This scheme therefore requires a higher amplifier gain, A, than there is loss in the lead network. Since the use of negative feedback will tend to stabilize the gain of the system, it is ordinarily preferred over lead-network compensation.

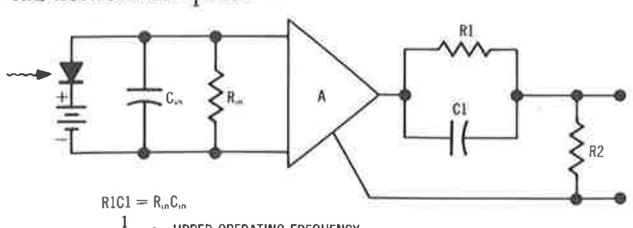
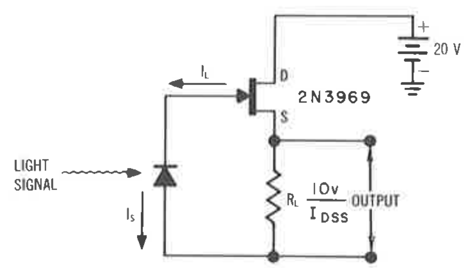


Figure 5. Lead-Network Compensation for Low Cutoff Frequency

When ultimate sensitivity is required, it is necessary to use an electrometer type of amplifier, but even with such an amplifier a careful design technique must be used. The principle involved in this technique is to simply represent all sources of noise in the amplifier as equivalent currents at the input. Noise sources which produce a constant output voltage with frequency, such as field-effect transistor channel noise and thermal noise, acquire an $(f)^{+1}$ spectral shape when they are referred to the input as equivalent currents because they are multiplied by the input susceptance. By plotting the asymptotes of noise current per root hertz from various sources in the amplifier, a profile of the variation with frequency of the signal-to-noise ratio can be obtained. Such a plot of asymptotes is shown in Figure 6. The limits within which the photodiode noise dominates are abundantly clear.



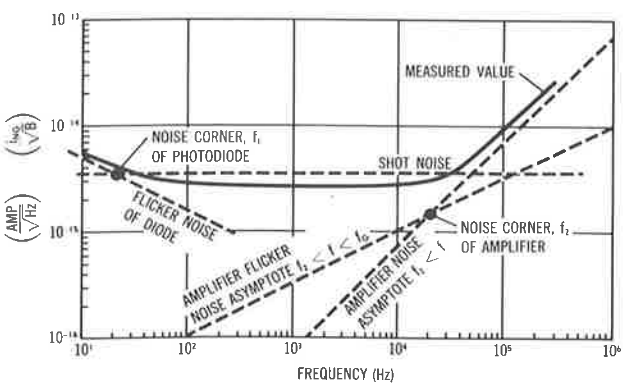
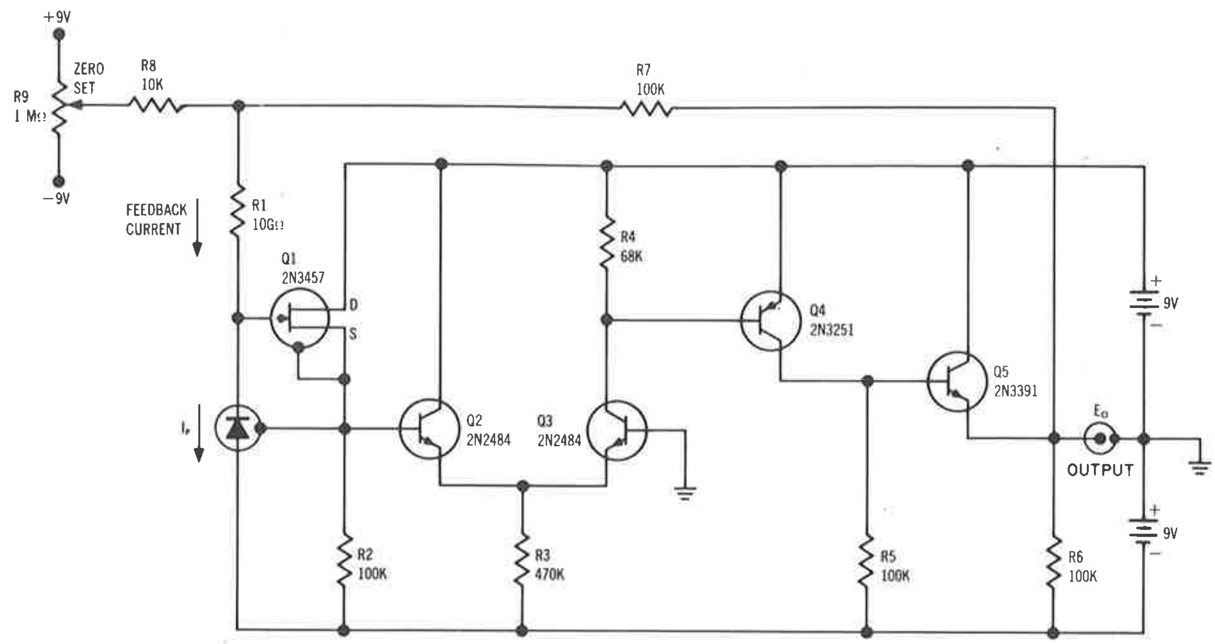


Figure 6. Calculated and Measured Noise Current Referred to Gate of Field Effect Transistor

A diffused-channel FET is selected for this application because its noise properties are much better than the MOS type. Suggested FET types are the 2N3457 and 2N3969 from Amelco Semiconductors, or similar low-noise field effect transistors.

In providing compensation for the low cutoff frequency, the circuit given in Figure 7 illustrates the feedback technique. A resistor of 10 gigaohms (10,000 megohms) was used for the feedback resistance, although the noise calculation indicated that 1.0 gigaohm (1000 megohms) would have been large enough. This is because the actual noise current is often actually less than the shot noise computed from leakage current, and leakage current is typically less than the specified maximum.

The FET is operated as a source follower, rather than a voltage amplifier, to avoid multiplication of the gate-to-drain capacitance. The source load resistance was selected so that, with I_{DSS} flowing in the channel, the source voltage would be near ground. A differential amplifier (Q2 and Q3) referred to ground was chosen rather than a single stage in order to keep the impedance at the source of Q1 as high as possible, and thus keep the signal-to-noise ratio high. Q4 provides the phase reversal needed for negative feedback and Q5 is added to keep the output impedance low.



AMPL. TRANSRESISTANCE $= \frac{\Delta E_o}{\Delta I_r} = 1.5 \times 10^{10} \frac{\text{Volt}}{\text{Amp.}} = \frac{\text{15 Volts}}{\text{per nA}}$

BANDWIDTH = 100 Hz

NOISE EQUIV. INPUT $= 42 \times 10^{-16} \frac{\text{watts}}{\sqrt{\text{Hz}}}$ = −104 dBm AT 800 NANOMETERS; = 2.1×10^{-5} FOOT-CANDLES AT 555 NANOMETERS

Figure 7. FET Amplifier with Current Feedback to Improve Bandwidth

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Printed in U.S.A. 30M-11/68