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DETECTORS FOR FIBEROPTIC COMMUNICATIONS

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DETECTORS

FOR

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In this chapter we discuss the optical detectors that are extensively used in fiber communications. We show their structure, the relevant characteristics, ther noise performance and their implementation in some typical receiver circuits.

THE PIN PHOTODIODE

The pin diode is the most significant detector for fiber communications. Its structure is given schematically in Fig. 1. A low conducting intrinsic (i) semiconductor zone is sandwiched between two thin heavily doped, i.e. highly conducting p and n zones. Photons reach the i zone through the antireflection coating and the p zone and they are absorbed by the semiconductor creating electron-hole pairs. In a diode biased in the reverse direction, as is normally

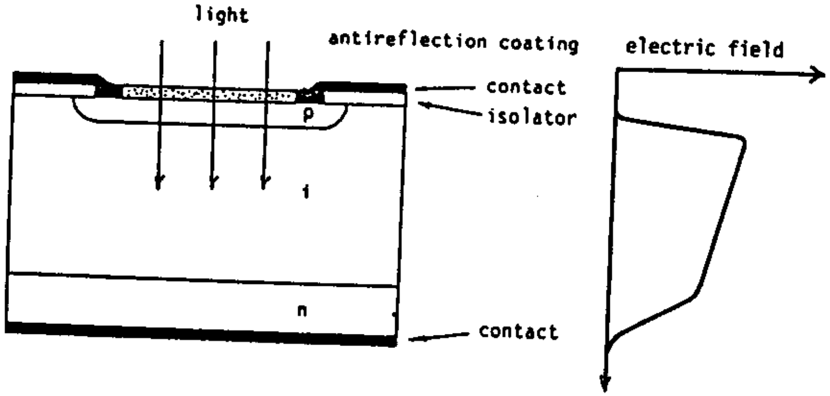


Fig. 1. Pin photodiode structure.

the case in a receiver circuit, but also in an unbiased diode, an electric field exists in the i zone that separates the electrons from the holes pulling them in opposite directions. Consequently, a photovoltage appears across the diode terminals and a photocurrent will flow in an external circuit.

In operation we may represent the diode as a current generator with an ideal diode in parallel, the latter being there to simulate the effect of the pn junction, Fig. 2. Furthermore, the shunt resistance R_{sh}, the capacitance C_d and the series resistance R_s are used to simulate the internal characteristics of the device. The photocurrent I_{ph} is given by the relation

$$I_{ph} = \frac{\eta e \lambda}{hc} \cdot P = R \cdot P$$

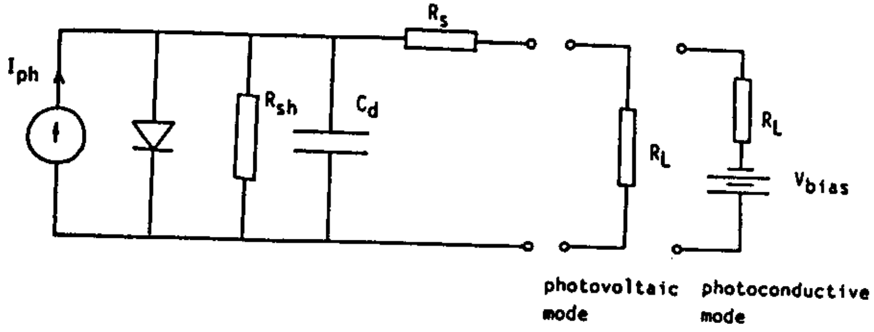


Fig. 2. Photodiode equivalent circuit.

e is the electronic charge, η the quantum efficiency, i.e. the ratio of the number of the generated electron-hole pairs to the number of incident photons, λ the wavelength, h the Planck's constant, c the velocity of light and P the optical power incident on the detector. R is the responsivity of the photodiode in A/W. If the device is left unbiased and a load is just connected across its terminals, it is said to operate in the photovoltaic mode. If, on the other hand, an external bias voltage is used in reverse direction, then the diode operates in the photoconductive mode.

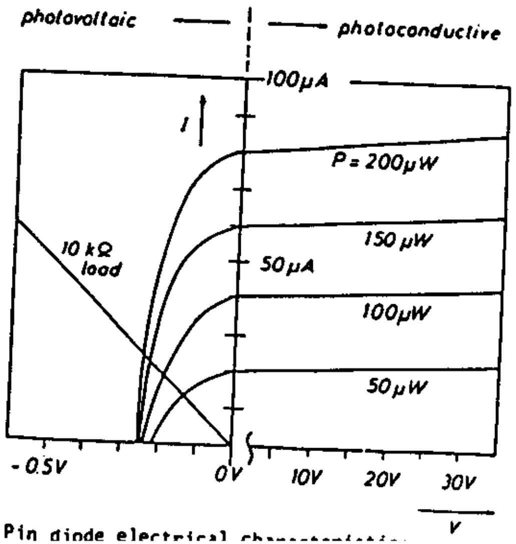


Fig. 3. Pin diode electrical ch...

Fig. 3 illustrates the electrical characteristics of a typical silicon pin detector for the two modes of operation.

In the photovoltaic mode the open-circuit voltage is given by:

$$V_o = \frac{kT}{e} \ln(1+\frac{I_{ph}}{I_s})$$

where k is Boltzsmann's constant $1.38 \cdot 10^{-23}$ JK^{-1}, T is the absolute temperature and I_s the diode saturation current. Since $I_{ph} >> I_s$ and $I_{ph} \sim P$, the open circuit voltage is a logarithmic function of the optical power. In this mode there is no current in the "dark" condition (P=0) which makes the photovoltaic operation attractive for the detection of low optical powers. If the diode is short-circuited, R_s creates a voltage drop which starts to limit the photocurrent at typically 1 mW optical power for silicon diodes depending on the active area. The disadvantage of the photovoltaic mode is the lower bandwidth caused by the large capacitance C_d.

In the photoconductive mode a relatively large negative bias (10-20V) is usually applied to the diode. The current flowing into the external circuit is linearly dependent on the optical power. In addition to this linearity, this mode offers the advantage of faster response, better stability and greater dynamic range. The main disadvantage is however the "dark" current which limits the ultimate sensitivity of the device. This current is in principle equal to I_s as is evident from the I-V-relationship of the diode

$$I = I_s \, [\exp(\frac{eV}{kT}) - 1]$$

if V is taken negative and much larger than kT/e. In practical values, for a 1 mm diameter silicon diode the linearity is very good up to about 1 mW optical power and the capacitance decreases to about 1 pF as the bias voltage becomes larger. The dark current is about 5 to 10 nA for silicon diodes at 25°C but it is significantly larger for germanium diodes (about 12 µA) and for Indium-Gallium-Arsenide-Phosphide (InGaAsP) diodes (about 6 µA). Since the dark current is haevily temperature dependent - it doubles every 7°C in Si, every 8°C in Ge and every 10°C in InGaAsP - and it generates extra noise, Ge and InGaAsP diodes must be made with small active areas (0.1-0.2 mm diameter) in order to reduce it. The responsitivity of the three diodes as a function of wavelength is given in Fig. 4.

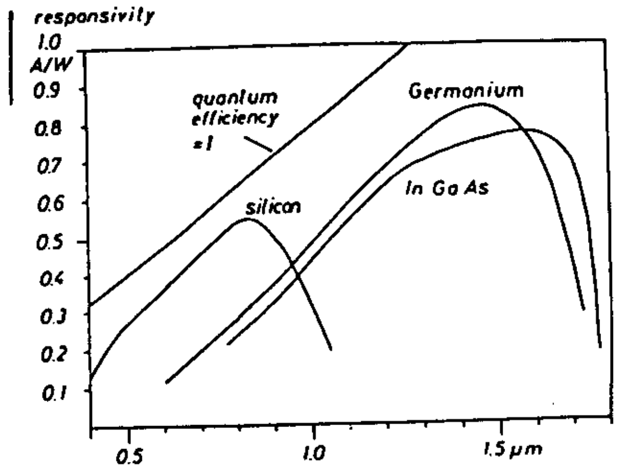


Fig. 4. Responsivity vs. wavelength for typical pin photodiodes.

Frequency response

Three main factors govern the frequency response of photodiodes

1) Diode capacitance.

   This is in reverse bias practically equal to the junction capacitance resulting from the charge stored in the depletion layer. It decreases with increasing negative bias. A cut-off frequency f_c can be defined as

$$f_c = \frac{1}{2\pi R_L C_d}$$

   if it is assumed that $R_{sh} >> R_s + R_L$ and $R_L >> R_s$, see Fig. 2.

   In order to increase f_c, diodes are fabricated with very small active areas. They should be operated at large reverse bias.

2) Diffusion time of carriers to the depletion region.

   Some carriers are generated outside the depletion region, which encompasses all the intrinsic layer, and diffuse relatively slowly to it where the are accelerated by the high field. The main result of the diffusion is a "slow tail" in the response of the diode to a rapid optical pulse.

3) Drift time of carriers through the depletion region.

In the high fields of the depletion region the drift velocities of the carriers tend to saturate. The longest transit time is then

$$\tau_{drift} = \frac{W}{v_{sat}}$$

where W is the width of the depletion region. In silicon we have $v_{sat} = 10^5$ m/s above 2000 V/m. So, for W = 5 µm, τ_{drift} becomes only 50 ps.

THE AVALANCHE PHOTODIODE (APD)

The APD has a higher responsitivity than a pin-photodiode due to its inherent gain process. It's construction is given in Fig. 5. A high electric field exists across the reverse

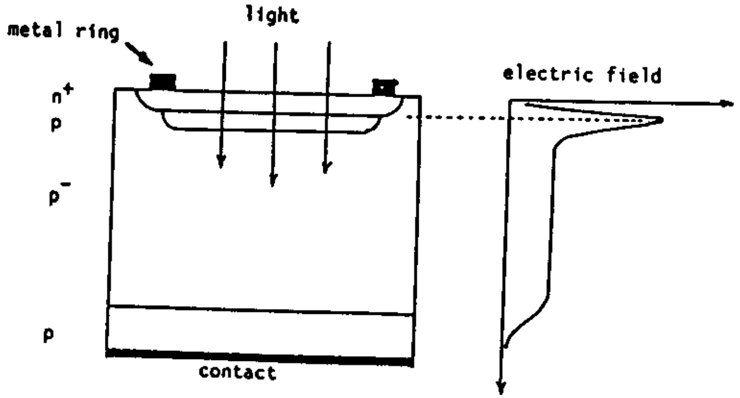


Fig. 5. Typical APD structure.

biased n+p junction. When carriers are generated in the lightly doped - practically intrinsic - p^-region, electrons travel to the high electric field where they gain sufficient kinetic energy to generate new electrons via carrier multiplication (avalanche process). If each of the electrons is assumed to create statistically M new electrons, the total photocurrent will be

$$I_{ph} = R \cdot M \cdot P$$

where M is the multiplication factor. The product R·M can be regarded to as a responsitivity [A/W] and much higher photocurrents are possible compared to the normal pin photodiode.

However, M depends strongly both on voltage and temperature, see Fig. 6. This makes some precautions necessary in practice. The power supply must

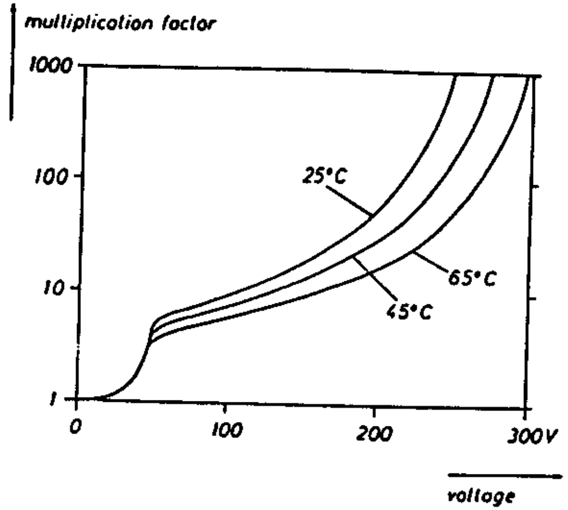


Fig. 6. Voltage and temperature dependence of the multiplication factor for a silicon APD.

be very stable in order to maintain a stable gain. Typical M-values are 50-150 for Si APD's and 20-40 for Ge APD's. Temperature stabilization will also be necessary. Another problem is the voltage drop across the load resistor which gives rise to non-linearities because of the steep M-V-curve at high voltages.

Dark current is also present in APD's. It depends on voltage and temperature. Typical values for a 0.1 mm diameter Si APD at 25°C lie between 1 and 10 nA for M = 100. In Ge APD's this current is larger, about 1 µA at M = 20. The dark current doubtes in Si APD's every 8°C temperature increase and every 9°C in Ge APD's.

The wavelength dependence of the responsitivity of APD's is quite similar to pin diodes, Fig. 4. It should be noted, however, that InGaAs-APD's are not yet technically as mature as Si- and Ge-APD's.

APD's are very fast photodetectors. Since they are made with very small active area diameter and they operate at large reverse voltage, their capacitance is low which results in large bandwidths. Si-APD's with 0.1 mm diameter have typical risetimes about 1 ns at M = 100 and 3 dB bandwidths about 300-500 MHz.

APD's with some GHz bandwidth have also been constructed. The frequency performance of Ge APD's is comparable to that Si-APD's.

NOISE OF THE PIN DIODE

The knowledge of the noise that is generated in a photodetector is of great importance because it directly influences the quality of signal at the output of the receiver circuit. The signal quality is usually given by the signal-to-noise ratio (SNR) for analog systems and by the bit-error rate (BER) for digital systems.

In a pin-diode the dominant noise source is shot noise. Physically, shot noise is due to the discrete nature of the carriers. It is generated by the statistical process of single carriers passing through the pin junction and it results to fluctuations of the current flow. The mean square noise current spectral density $\overline{i_n^2}$ is directly proportional to the diode current I

$$\overline{i_n^2} = 2eI \quad \text{in } [A^2/Hz]$$

The current I is the sum of the signal related photocurrent I_{signal} and of the dark current I_{dark}. Thus, the smallest detectable optical power P_{min} in a frequency bandwidth B will be determined by the dark current as follows:

$$P_{min} = \frac{1}{R}\sqrt{\overline{i_n^2} \cdot B} \text{ at } I_{signal} = 0$$

$$P_{min} = \frac{1}{R}\sqrt{2eI_{dark} \cdot B} \quad \text{in } [nW]$$

The noise equivalent power (NEP), a parameter given in many photodiode data-sheets, is defined as the (fictitious) optical power per square root of Hertz causing the dark current

$$NEP = \frac{1}{R}\sqrt{2eI_{dark}} \quad [W/\sqrt{Hz}]$$

Some typical values of NEP are given below.

Silicon diode, 1 mm diameter, I_{dark} = 5 nA, NEP = $8 \cdot 10^{-14}$ $W/\sqrt{Hz}$.
Ge diode, 0.2 mm diameter, I_{dark} = 0.5 µA, NEP = 10^{-12} $W/\sqrt{Hz}$.
InGaAs diode, 0.1 mm diameter, I_{dark} = 60 nA, NEP = $2.7 \cdot 10^{-13}$ $W/\sqrt{Hz}$.
It should be noted, however, that the measurements for the Ge and the InGaAs diode were made at 1.3 µm i.e. near their highest responsivity whereas the Si diode was measured at 0.8 µm.

NOISE OF THE AVALANCHE PHOTODIODE

In an APD the mean square noise current is multiplied by the square of the multiplication factor. If the original current before multiplication is I', the actual current will be

$$I = M \cdot I' \quad \text{and so we have}$$

$$\overline{i_n^2} = 2eI' \, M^2 \, B \cdot F(M) \quad \text{and}$$

$$\overline{i_n^2} = 2eI \, M \, B \cdot F(M)$$

where F(M) is an additional factor, called excess noise factor, that is introduced by the statistical nature of the multiplication process. Since I is the sum of signal current and dark current we may write:

$$\overline{i_n^2} = 2e\,(I_{signal} + I_{dark})\, MF(M).B$$

The factor F(M) increases with increasing M and may by given by the approximate relation

$$F(M) = M^x$$

where x = 0.2 - 0.5 for Si APD's and

x = 0.9 - 1.0 for Ge APD's

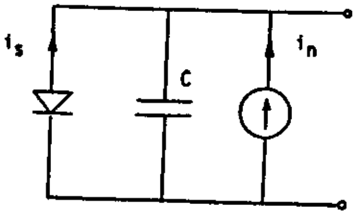
The noise equivalent power NEP for APD's is given by:

$$NEP = \frac{1}{RM}\sqrt{2eI_{dark} \cdot M \cdot F(M)} = \frac{1}{R}\sqrt{2eI_{dark}\, F(M)/M}$$

Comparing this value with the NEP of pin photodiodes we should bear in mind that I_{dark}/M is the same as the dark current of the pin diode. Thus, the NEP of the APD is higher than the NEP of the pin diode by $\sqrt{F(M)}$. Typical NEP's are $5 \cdot 10^{-14}$ $W/\sqrt{Hz}$ at 0.9 µm for Si APD's and $5 \cdot 10^{-13}$ $W/\sqrt{Hz}$ at 1.3 µm for Ge-APD's.

PRACTICAL RECEIVER DESIGN

We now turn our attention to the realization of practical receivers with special emphasis to the signal quality at the output. We assume that the shunt resistor in the photodiode model Fig. 2 is much larger than the sum of load and series resistor and that the load resistor itself is much larger than the series resistor. So, the resistors R_{sh} and R_s can be omitted from the equivalent circuit which now comprises a current noise generator to account for the statistical mature of the detection process, Fig. 7. This simplified noise equivalent circuit can be used for both pin and APD's. A word of caution is, however, necessary. If LED light is detected, the noise current in the detector



signal current $i_s = R \cdot M \cdot P_{optical}$

noise current

$$\overline{i_n^2} = 2eB\,(R \cdot P_{opt} + I_{dark})\; M^2 \cdot F(M)$$

Fig. 7. Simplified noise equivalent circuit for pin and APD's.

diode can be calculated from the equations given in Fig. 7. In the case of laser light, however, the noise current can be significantly above the level predicted by the equations. This noise originates in the laser radiation and may stem from different causes such as the laser process itself or the power competition between the modes. Also, fluctuations of the laser power due to optical reflections from fiber, connectors or splices back into the laser are possible. Finally, intensity noise caused by speckle fluctuations at fiber connectors or splices ("modal noise") may increase the noise current in the photodetector. The calculations carried below assume LED radiation carrying the signal because laser light fluctuations are more complex and depend heavily on the experimental situation.

Before we proceed with the calculations of typical receiver circuits, we should have a look at the resistor noise. Any resistor R at an absolute temperature T produces thermal noise which may be described by the mean square noise voltage spectral density apearing at the resistor terminals

$$\overline{u_n^2} = 4\,kTR$$

An alternative may to describe this noise is to assume a current source in parallel with the resistor generating the mean square noise current spectral density $\overline{i_n^2}$

$$\overline{i_n^2} = 4\,kT \quad /R.$$

We now turn our attention to two practical receiver circuits, widely used in fiber communications: the low-impedance type and the transimpedance type. In a simplified form, these circuits are illustrated in Figs 8 and 9 respectively. In the low-impedance type, the photodiode is in series with the resistor R_1 and the photocurrent produces a voltage across R_1 which is amplified in the following amplifier. The resistance R_1 may be thought in the real situation

to be the parallel combination of the load (or bias) resistor R_L, the diode shunt resistor R_{sh} (see Fig. 2) and the amplifier input resistor R_A neglecting the diode series resistor R_s.

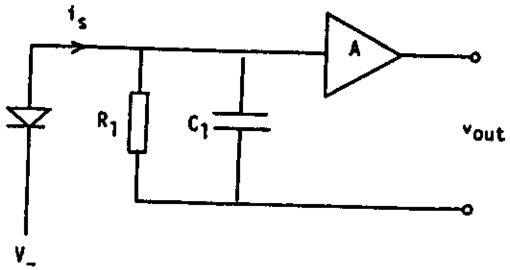


Fig. 8. Schematic diagram of the low-impedance type receiver.

$$\frac{1}{R_1} = \frac{1}{R_{sh}} + \frac{1}{R_L} + \frac{1}{R_A}$$

The capacitor C_1 incorporates the diode C_d, amplifier C_A and the stray capacitance C_s

$$C_1 = C_d + C_A + C_s$$

If a sinusoidally varying optical power impinges on the detector, then the photogenerated current i_s will also be sinusoidally modulated and the voltage v_{in} at the amplifier input is:

$$v_{in}(f) = \frac{R_1 \cdot i_s(f)}{1 + j2\pi f C_1 R_1}$$

So, for a good figh-frequency response, C_1 must be kept low by applying a large negative diode bias voltage, reducing the stray capacitances and using amplifiers with very low input capacitance. The bandwidth can also be maximized by reducing R_1 and by providing some sort of high-frequency equalization.

An important prerequisite for the calculation of the total noise appearing at the receiver output is the knowledge of the noise generated by the active elements that compose the amplifier. The usual way to treat the subject is to take the amplifier as a whole and to consider the noise equivalent circuit given in Fig. 9.

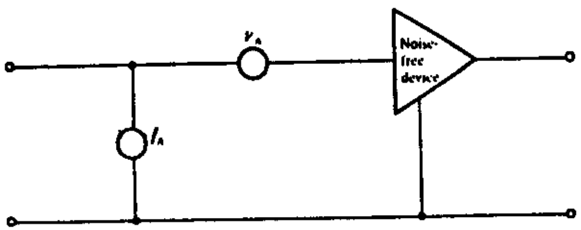


Fig. 9. Noise equivalent circuit of an amplifier.

In this circuit, all the noise sources internal to the amplifier are represented by a noise voltage v_A and a noise current source i_A at the input terminals, the amplifier itself beeing now a noise free circuit. It is also usual in the practical application to consider these noise sources as uncorrelated to each other and also independent of the other noise sources of the receiver, such as the detector noise.

If the first active device of the amplifier is a field effect transistor (FET) then a major source of noise is the thermal noise associated with its channel resistance. As a result, the main component of v_A will be represented by the term $(\delta 4kT/g_m)^{1/2}$ where g_m is the forward transconductance of the FET. The multiplication factor δ is about 0.7 for silicon and 1.1 for GaAs devices.

If the first active device is a bipolar transistor, then a major source of noise is the shot noise of the base and collector currents, I_B and I_C respectively. The base current gives rise to a term $\sqrt{2eI_B}$ to the noise current i_A and the collector current to a term $\sqrt{2(kT)^2/eI_C}$ to the noise voltage v_A. The dc currents I_B and I_C are of course linked together by the relation $I_C = \beta I_B$, where β is the current gain of the transistor.

We can now draw a circuit which considers all the main sources of noise in our low-impedance type receiver, see Fig. 10:

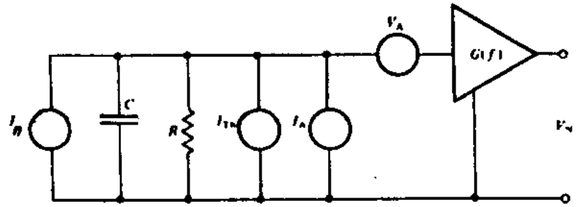


Fig. 10. Noise equivalent circuit of a low-impedance receiver

In order to simplify things let us suppose that we use an amplifier with constant gain A in the frequency band of interest. Further we assume a digital signal with optical detected power P_1 at the "1"-level, 0 at the "0"-level and an even distribution of "ones" and "zeros". Then, the signal bandwidth will be

$$B = \frac{1}{2\pi R_1 C_1}$$

The output voltage v_{out} for the "1"-level is:

$$v_{out} = R_1 \cdot A\, i_s = R_1 \cdot A \cdot P_1 \cdot R, \quad R \text{ is the responsivity.}$$

The noise voltage at the output is then sum of the amplified noise voltages appearing across the resistor R_1 and generated by the signal current, the dark current and the thermal noise current. To these currents we have to add the noise generated in the amplifier itself. It can be shown that this noise gives rise to two terms:

$\frac{4}{3}\frac{v_A^2}{R_1^2}$, where v_A is the equivalent voltage noise of the amplifier and i_A^2, where i_A is the equivalent current noise of the amplifier. The average mean square value of the output noise voltage is, assuming an even distribution of "ones" and "zero":

$$\overline{v_n^2} = A^2 R_1^2 B \left(2eR\frac{P_1}{2} + 2eI_{dark} + \frac{4kT}{R_1} + \frac{4}{3}\frac{v_A^2}{R_1^2} + i_A^2\right)$$

We can now give the signal to noise ratio at the output:

$$SNR = 10 \log \left(v_{out}^2 / \overline{v_n^2}\right) =$$

$$= 10 \log \frac{P_1^2 R^2}{B\left(eP_1R + 2eI_{dark} + \frac{4kT}{R_1} + \frac{4}{3}\frac{v_A^2}{R_1^2} + i_A^2\right)}$$

Numerical example: 1) <u>Low impedance receiver with pin diode</u> $R_{sh} >> R_B$

$P_1 = 1\ \mu W = -30$ dBm; $I_{dark} = 10$ nA; $R = 0.4$ A/W, $k = 1.38 \cdot 10^{-23}$ J/K

$A = 40$; $C_1 = 5$ pF; $B = 70$ MHz suitable for a 140 Mbit/s digital transmission system. The amplifier has $v_A = 2\ nV/\sqrt{Hz}$ and $i_A = 2\ pA/\sqrt{Hz}$.

The resistor R_1 becomes : $R_1 = \frac{1}{2\pi B \cdot C_1} = 455\ \Omega$

By taking $R_1 = 470\ \Omega$, B will be 68 MHz

For the "1"-level, the output voltage is:

$$v_{out} = 7.5\ mV$$

The noise output is

$$\overline{v_n^2} = 2.5 \cdot 10^{16}(6.4 \cdot 10^{-26} + 3.2 \cdot 10^{-27} + 3.5 \cdot 10^{-23} + 2.4 \cdot 10^{-23} + 4 \cdot 10^{-24}) =$$

$$= 1.58 \cdot 10^{-6} V^2$$

It is evident, that in our case the resistor is the most important noise source but the contribution of the amplifier we have chosen comes close to it. Unfortunately, we cannot minimize the resistor noise because if we encrease the resistor, thus reducting the resistor noise current, the signal bandwidth will decrease.

The SNR becomes

SNR = 15.5 dB (electrical).

Since we have assumed a digital transmission, we must somehow relate this SNR-value to the quality of the received signal. This is done by determining the bit error rate (BER) which is the number of false bits relative to the total number of bits in the data stream. In the field, many systems require a BER of 10^{-9} and some systems even 10^{-12}. Under some simplified assumptions, the BER can be calculated as a function of the SNR. It turns out that for a BER of 10^{-9}, a SNR of 21.6 dB is required. This value is clearly above the SNR of our system.

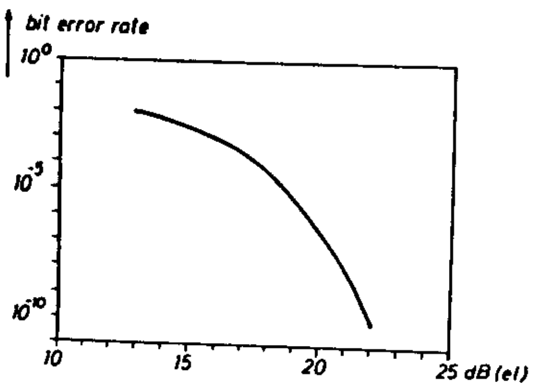


Fig. 11. BER vs. SNR

In order to obtain the BER of 10^{-9} we would have to increase the optical power at the photodiode. This power increase should result to an increase of the SNR from 15.5 to 21.6 (about 6 dB electrical), which means that the optical power would have to be larger by 3 dB (optical). Instead of 0.5 µW average power (-33 dBm) we should have 1 µW (-30 dBm), or 2 µW for the "1"-level.

The transimpedance receiver circuit of Fig. 12 is very popular in optical communications. If the amplifier has a large input impedance, the resistance R which is the parallel combination of diode shunt resistance, diode bias resistance and amplifier input resistance can be made larger and the photodiode current (signal current plus dark current) will flow through the resistor R_2. The calculations yield to the following approximate results:

Signal Bandwidth $\quad B = \frac{A}{2\pi R_2 C_2}$

"1"-level output voltage $\quad v_{out} = R_2 i_s = R_2 P_1 R$

Output noise voltage $\quad \overline{v_n^2} = R_2^2\ B\ (2e\frac{P_1}{2}R + 2eI_{dark} + \frac{4kT}{R_2} + \frac{4}{3}\frac{v_A^2}{R_2^2} + i_A^2)$

Signal-to-noise-ratio $\quad SNR = 10 \log (v_{out}^2 / \overline{v_n^2})$

$$= 10 \log \frac{P_1^2 R^2}{B(eP_1 R + 2eI_{dark} + \frac{4kT}{R_2} + \frac{4}{3}\frac{v_A^2}{R_2^2} + i_A^2)}$$

2) Transimpedance receiver with pin diode, numerical example

For the benefice of comparison with the low-impedance receiver we assume the same values for the numerical example. The calculations yield:

Resistor $\quad R_2 = \frac{A}{2\pi B\ C_2} = 18.7\ k\Omega$.

By taking R_2 = 18 kΩ B will be 70.7 MHz

v_{out} = 7.2 mV

$\overline{v_n^2}$ = $2.3\cdot10^{16}(6.4\cdot10^{-26}+3.2\cdot10^{-27}+9.2\cdot10^{-25}+1.6\cdot10^{-26}+4\cdot10^{-24}$
= $1.15\cdot10^{-7}V^2$

We see that the resistor noise and the noise due to the voltage noise of the amplifier are significantly lower than in the previous example. The dominant noise term is the amplifier current noise but now the average noise voltage is lower than in the previous case.

SNR = 26.5 dB (electrical)

The SNR is now much higher than before and also larger thant the value required for BER of 10^{-9}. This is due to the lower noise generated by the feedback resistor R_2 in the transimpedance circuit and because of the reduced influence of the equivalent voltage noise of the amplifier.

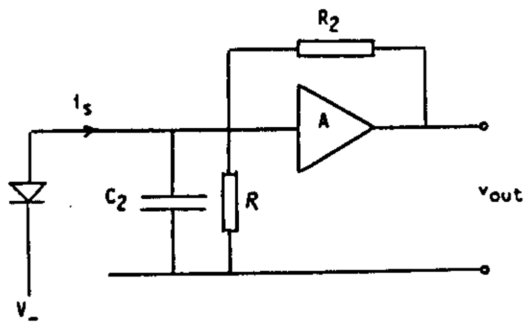


Fig. 12. Transimpedance receiver circuit.

FOR FURTHER READING

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