Optical Receiver

10

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10.1. INTRODUCTION

In optical fiber communication systems, optical signals that reach fiber optic receivers are generally attenuated and distorted, as shown in Fig. 10.1.

The optical fiber receiver must convert the input signal and amplify the resulting electrical signal without distorting it to a point. An optical receiver consist of :

- Photodetector
- An amplifier
- Signal Processing Circuitry

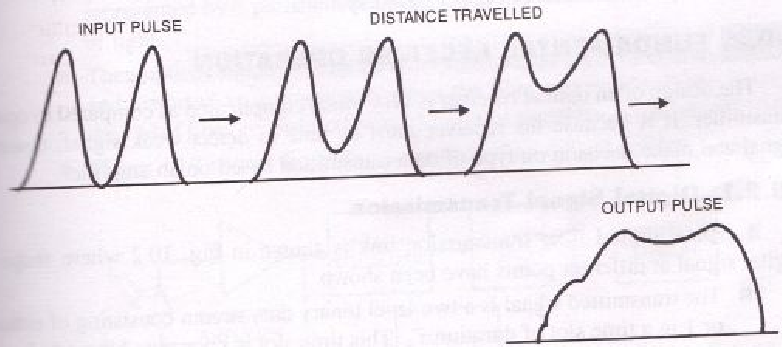


Fig. 10.1. Attenuated and Distorted Optical Signals.

In most fiber optic system, the **optical detector** is a PIN photodiode or APD. Receiver performance varies depending on the type of detector used.

The **amplifier** is generally described as having two stages :

- Pre-amplifier
- Post-amplifier.

The **Preamplifier** is defined as the first stage of amplification following the optical detector. The **postamplifier** is defined as the remaining stages of amplification required to raise the detector's electrical signal to a level suitable for further signal processing. The preamplifier is the domain contributor of the electrical noise in the receiver. Because of this, its design has a significant influence in determining the sensitivity of the receiver.

The **signal processing circuitry** processes the amplified signal into a form suitable for interfacing circuitry. For digital receivers, this circuitry may include low pass filters and comparators. For analog receivers this circuitry may also include low pass filters.

The **key operational parameters used** to define the receiver performance are :

- Receiver Sensitivity
- Bandwidth
- Dynamic Range

One goal in designing the fiber optic receivers is to optimize receiver sensitivity. To increase sensitivity, receiver noise resulting from signal dependent shot noise and thermal noise must be kept at a minimum. A more detailed discussion of receiver shot and thermal noise is provided later in this chapter.

In addition to optimizing sensitivity, optical receiver design goals also include optimizing the bandwidth and the dynamic range. A receiver that has the ability to operate over a wide range of optical power levels can operate efficiently in both short and long distance applications. Because conflicts arise when attempting to meet each goal, trade offs in receiver design are made to optimize overall performance.

While designing receiver it is useful to regard the limit on the performance of the system set by the signal to noise ratio (SNR) at the receiver. Therefore it is necessary to outline noise sources within optical fiber systems. In these systems the noise has different origins from that of copper based systems.

10.2. FUNDAMENTAL RECEIVER OPERATION

The design of an optical receiver is very much complicated as compared to optical transmitter. It is because the receiver must be able to detect weak signal, distorted signal and make decision on type of data transmitted based on an amplifier.

10.2.1. Digital Signal Transmission

A typical digital fiber transmission link is shown in Fig. 10.2 where shape of digital signal at different points have been shown.

- The transmitted signal is a two level binary data stream consisting of either 0 or 1 in a time slot of duration T_b. This time slot is known as **bit period**.
- One of the simplest technique of sending binary data is amplitude shift keying or on-off keying (OOK), where a voltage level is switched between two values, usually on or off. The resultant signal wave therefore consist of voltage pulse

of amplitude v relative to zero voltage when binary 1 occurs and zero voltage level space when binary 0 occurs. For simplicity it is assumed that when 1 is transmitted, voltage pulse of duration T_b occurs while for 0 the voltage remains at its zero level.

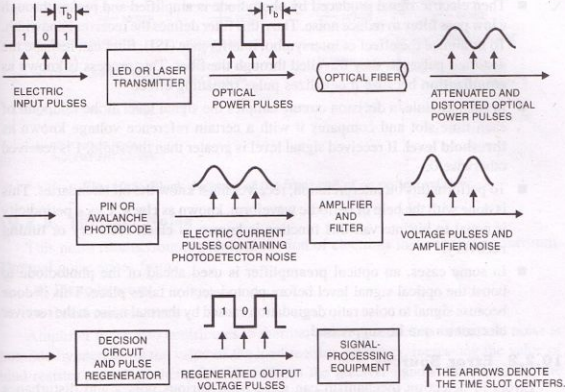


Fig. 10.2.

- The function of the optical transmitter is to convert the electrical signal in optical signal. One approach is by directly modulating the light source drive current with the information stream to produce a varying optical output power $P(t)$. Therefore in optical signal coming from LED or laser transmitter, 1 is represented by a pulse of optical power of duration T_b while 0 is the absence of light.
- Then optical signal is coupled from light source to fiber becomes attenuated and distorted as it propagates along the fiber waveguide. When it arrives at the end of fiber, receiver converts the optical signal back to an electrical format.
- Figure 10.3 shows the basic component of an optical receier.

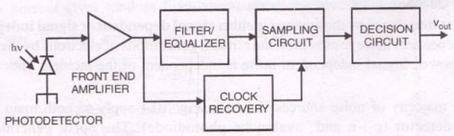


Fig. 10.3. Basic sections of an optical receiver

- First element is either pin or an avalanche photodiode which produces an electric current proportional to the received power level. Actually this electric current is weak, **front end amplifier** is used to boosts it to the level that can be used by the following electronics.
- Then electric signal produced by photodiode is amplified and passes through a **low pass filter** to reduce noise. Thus, this filter defines the receiver bandwidth. To minimize the effect of intersymbol interference (ISI), filter can reshape the distorted pulses as they travelled through the fiber. This process is known as **equalization** because it equalizes pulse spreading effect.
- In final module, a decision circuit samples the signal level at the midpoint of each time slot and compares it with a certain reference voltage known as **threshold level.** If received signal level is greater than threshold, 1 is received otherwise 0.
- To perform this bit interpretation, receiver must know the bit boundaries. This is done with the help of periodic waveform, known as **clock.** Clock periodicity is equal to bit interval. this function is known as **clock recovery** or **timing recovery.**
- In some cases, an optical preamplifier is used ahead of the photodiode to boost the optical signal level before photodetection takes place. This is done because signal to noise ratio degradation caused by thermal noise in the receiver electronics can be suppressed.

10.2.2. Error Sources

Errors in detection mechanism can arise from various noises and disturbance associated with the signal detection system.

Noise is defined as any spurious or undesired distrubance that mask the received signal in a communication system. There are many sources of noise in fiber optic systems, they are :

- Noise from the light source.
- Noise from the interaction of light with the optical fiber.
- Noise from the receiver itself.

There are three main types of noise due to spontaneous fluctuations in optical fiber communication systems :

- Thermal noise
- Dark current noise
- Quantum noise

Noise introduced by the receiver is either **signal dependent** or **signal independent.** Signal dependent noise results from the random generations of electrons by the incident optical power. Signal independent noise is independent of the incident optical power level.

The majority of noise sources shown in Fig. 10.4 apply to both main types of optical detector (p-i-n and avalanche photodiode). The noise generated from background radiation and beat noise generated from the various spectral components are negligible in the both type of optical fiber receiver.

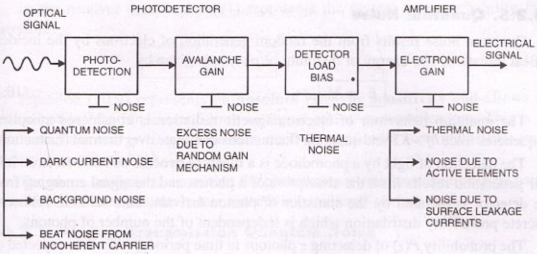


Fig. 10.4. Block diagram of the front end of an optical receiver.

10.2.3. Thermal Noise

This noise results from the random motion of electrons in a conducting medium. Thermal noise arises from :

- Photodetector
- Load resistor.

Amplifier noise also contributes to thermal noise. A reduction in thermal noise is possible by increasing the value of the load resistor. However, increasing the value of load resistor to reduce thermal noise reduces the receiver bandwidth. In APDs, the thermal noise is unaffected by the internal carrier multiplication.

The thermal noise current i_t in a resistor R may be expressed by its mean square value and is given by

$$\overline{i_t^2} = \frac{4KTB}{R} \quad ...(10.1)$$

where, K = Boltzmann's constant

T = Absolute temperature

B = Post detection bandwidth of the system.

10.2.4. Dark Current Noise

Dark current noise results from dark current that continues to flow in the photodiode when there is no incident light. Dark current noise is independent of the optical signal. The dark current gives random fluctuations about the average particle flow of the photocurrent. The dark current noise $\overline{i_d^2}$ is given by

$$\overline{i_d^2} = 2eBI_d \quad ...(10.2)$$

where, e = Charge on an electron

I_d = Dark current

B = Bandwidth

10.2.5. Quantum Noise

Quantum noise results from the random generation of electrons by the incident optical radiation. The energy of quantum or photon is given by

$$E = hf \quad ...(10.3)$$

The quantum behaviour of electromagnetic radiation is considered at optical frequencies since $hf > KT$ and quantum fluctuations dominate over thermal fluctuations.

The detection of light by a photodiode is a **discrete process** because electron hole pair generation results from the absorption of a photon and the signal emerging from the detector is dictated by the statistics of photon arrivals. This statistic follows a discrete probability distribution which is independent of the number of photons.

The probability $P(z)$ of detecting z photons in time period τ when it is expected on average to detect z_m photons follows the **Poisson distribution.**

$$P(z) = \frac{z_m^z \exp(-z_m)}{z!} \quad ...(10.4)$$

where, z_m = Variance of the probability distribution.

The electron rate r_e generated by incident photons is

$$r_e = \frac{\eta P_0}{hf} \quad ...(10.5)$$

The number of electrons generated in time τ is equal to the average number of photons detected over this time period z_m. Therefore,

$$z_m = \frac{\eta P_0 \tau}{hf} \quad ...(10.6)$$

Independent atoms emit incoherent light and therefore there is no phase relationship between the emitted photons. This property gives exponential intensity distribution for incoherent light which if averaged over the Poisson distribution gives

$$P(z) = \frac{z_m^z}{(1+z_m)^{z+1}} \quad ...(10.7)$$

Equation (10.7) is identical to **Bose-Einstein distribution,** which is used to describe the random statistics of light emitted in black body radiation.

10.2.6. Digital Signalling Quantum Noise

It is possible to calculate a fundamental lower limit to the energy in digital optical fiber systems. The concept of this analysis is that the ideal receiver has a sufficiently low amplifier noise to detect the displacement current of single electron-hole pair generated within the detector. Therefore in the absence of light and neglecting dark current, no current will flow. Thus an error can occur if light pulse is present and no electron hole pairs are generated. When a light pulse is present, the probability of no pairs being generated is given by

$$P\left(\frac{0}{1}\right) = \exp(-z_m) \quad ...(10.8)$$

In the receiver described $P(0/1)$ represents the system error probability $P(e)$, therefore

$$P(e) = \exp(-z_m) \quad ...(10.9)$$

Equation (10.9) represents an **absolute receiver sensitivity** and allows the determination of a fundamental limit in digital optical communications. This is the *minimum pulse energy* (E_{min}) required to maintain a given bit error rate (BER) and is known as **quantum limit**.

The above analysis assumes that the photodetector emits no electron-hole pair in the absence of illumination. In this sense it is considered perfect.

10.2.7. Analog Transmission Quantum Noise

In analog optical fiber systems quantum noise means shot noise and it has Poisson statics. The shot noise current i_s on the photocurrent I_P is given by

$$\overline{i_s^2} = 2eBI_p \quad ...(10.10)$$

By neglecting other noise sources, the SNR at the receiver may be written as

$$\frac{S}{N} = \frac{I_P^2}{\overline{i_s^2}} \quad ...(10.11)$$

Subtituting the value of $\overline{i_s^2}$, we get

$$\frac{S}{N} = \frac{I_P}{2eB} \quad ...(10.12)$$

Substituting the value of I_P from Eq. (9.9),

The SNR in terms of incident optical power P_0 given by

$$\frac{S}{N} = \frac{\eta P_0 e}{hf\, 2eB}$$

$$\frac{S}{N} = \frac{\eta P_0}{2hfB} \quad ...(10.13)$$

The incident optical power required at the receiver can be calculated by Eq. (10.13).

Another Error Source

Another error source is known as Intersymbol interference (ISI). It results from pulse spreading in the optical fiber. When a pulse is transmitted in a given list slot, most of the pulse energy will arrive in the corresponding time slot at the receiver as shown in Fig. 10.5.

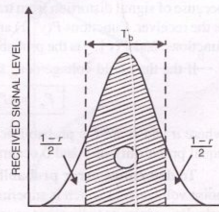


Fig. 10.5.

10.3. DIGITAL RECEIVER PERFORMANCE

Ideally in digital receiver, the decision circuit output signal voltage $v_{out}(t)$ would always exceed the threshold voltage when 1 is present and would be less than threshold when 0 (no pulse) was sent. But in actual system, deviations from the average value of $v_{out}(t)$ are caused by various noises, interferences from adjacent pulse and conditions wherein the light source is not completely extinguished during zero pulse.

10.3.1. Probability of Error

Practically number of ways are there to measure the rate of error occurrence in digital data stream. A simple approach is to divide the number N_e of errors occurring over a certain time interval t by the number N_t of pulses transmitted during this interval. This is known as **error rate** or **bit error rate**, abbreviated as BER.

$$BER = \frac{N_e}{N_t} = \frac{N_e}{BT} \quad ...(10.14)$$

where, $B = \frac{1}{T_b}$ = bit rate

The error rate is expressed by a number as example : 10^{-9} *i.e.*, one error occurs for every billion pulses sent. Typical error rate for optical fiber telecommunication system range from 10^{-9} to 10^{-12}. This error rate depends on signal to noise ratio at the receiver.

To calculate the bit error rate at the receiver, knowledge of probability distribution of the signal at the equalizer output should be known. Figure 10.6 shows the shapes of two signal probability distributions.

These are : $$P_1(v) = \int_{-\infty}^{v} P(y|1)\, dy \quad ...(10.15)$$

which is the probability that the equalizer output voltage is less than v when a logical 1 pulse is sent, and

$$P_0(v) = \int_{v}^{\infty} P(y|0)\, dy \quad ...(10.16)$$

which is the probability that output voltage exceeds v when logic 0 in transmitted. As shown in Fig. 10.6, different shapes of two probability distribution indicates that the noise power for logic 0 need not be same as logic 1. In optical system it occurs because of signal distortion from transmission impairments, noise and ISI contributions at the receiver. Functions $P(y \mid 1)$ and $P(y \mid 0)$ are the conditional probabilty distribution functions, and $P(y \mid x)$ is the probability that output voltage is y when x was transmitted.

If the threshold voltage is v_{th} then **the error probability (P_e) is given as**

$$P_e = aP_1(v_{th}) + bP_0(v_{th}) \quad ...(10.17)$$

where a and b are the probabilities that either 1 or 0 occurs. For unbiased data with equal probability of 1 and 0 occurances, $a = b = 0.5$.

To determine error probability, we should have the knowledge of mean square noise voltage ($\overline{v_N^2}$) which is superimposed on the signal voltage at decision time. Exact calculation is quite tedious but approximation can be done by making a tradeoff between computaional simplicity and accuracy of the results.

A simple approach is based on gaussian approximation. It is assumed that when sequence of optical input pulses is known, the equalizer output voltage $v_{out}(t)$ is a guassian random variable. Therefore to calculate error probability we need to know the mean and standard deviation of $v_{out}(t)$.

Now let us assume that a signal '*s*' has a guassian probability distribution function with a mean value *m*. If we sample the signal voltage level $s(t)$ at any orbitrary time t_1, the probability that the measured sample $s(t_1)$ falls in the range *s* to $s + ds$ is given by

$$f(s)\,ds = \frac{1}{\sqrt{2\pi\sigma^2}} e^{-(s-m)^2/2\sigma^2}\,ds \qquad ...(10.18)$$

where, $f(s)$ = probability density function

σ^2 = noise variance

σ = standard deviation, measure of width of probability distribution.

Quantity $2\sqrt{2}\sigma$ = measures full width of the probability distribution at the point where amplitude is $\frac{1}{e}$ of the maximum.

Now we can use the probability density function to determine the probability of error of a data stream in which the 1 pulses are all of amplitude *v*. As shown in Fig. 10.6 the mean and variance of the guassian output for 1 pulse are b_{on} and σ^2_{on} respectively while for 0 pulse they are b_{off} and σ^2_{off}.

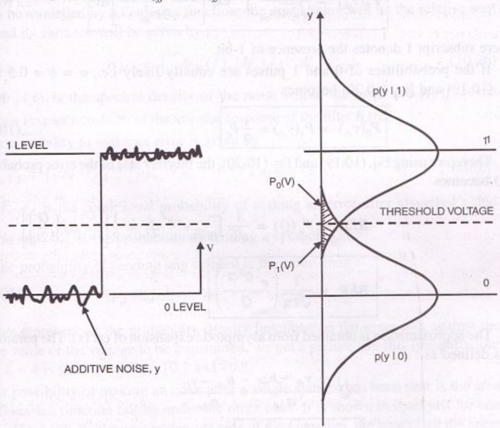


Fig. 10.6.

First let us consider the case of 0 pulse transmitted so that no pulse is present at decoding time. In this case the probability of error is the probability that the noise will exceed the threshold voltage v_{th} and be mistaken for 1 pulse. This probability of error $P_0(v)$ is the chance that the equalizer output voltage $v(t)$ will fall somewhere between v_{th} and ∞. From Eqs. (10.16) and (10.18), we get

$$P_0(v_{th}) = \int_{v_{th}}^{\infty} P(y|0)\,dy = \int_{v_{th}}^{\infty} f_0(y)\,dy$$

$$P_0(v_{th}) = \frac{1}{\sqrt{2\pi}\,\sigma_{off}} \int_{v_{th}}^{\infty} \exp\left[-\frac{(v-b_{off})^2}{2\sigma^2_{off}}\right] dv \qquad ...(10.19)$$

where the subscript 0 denotes the presence of 0-bit.

Similarly, we can find the probability of error that transmitted 1 is misinterpreted as 0 by decoder electronics following the equalizer. For this sampled signal plus noise pulse falls below v_{th}. From Eq. (10.17) and Eq. (10.18), we get

$$P_1(v_{th}) = \int_{-\infty}^{v_{th}} P(y|1)\,dy$$

$$= \int_{-\infty}^{v_{th}} f_1(v)\,dv$$

$$P_1(v_{th}) = \frac{1}{\sqrt{2\pi}\sigma_{on}} \int_{-\infty}^{v_{th}} \exp\left[-\frac{(b_{on}-v)^2}{2\sigma^2_{on}}\right] dv \qquad ...(10.20)$$

where subscript 1 denotes the presence of 1-bit.

If the probabilities of 0 and 1 pulses are equally likely *i.e.*, $a = b = 0.5$ then Eq. (10.19) and Eq. (10.20) becomes

$$P_0(v_{th}) = P_1(v_{th}) = \frac{1}{2} P_e \qquad ...(10.21)$$

Therefore using Eq. (10.19) and Eq. (10.20), the bit error rate of the error probability (P_e) becomes

$$BER = P_e(Q) = \frac{1}{\sqrt{\pi}} \int_{Q/\sqrt{2}}^{\infty} e^{-x^2}\,dx = \frac{1}{2}\left[1 - erf\left(\frac{Q}{2}\right)\right]$$

$$BER \approx \frac{1}{\sqrt{2\pi}} \left(\frac{e^{-Q^2/2}}{Q}\right) \qquad ...(10.22)$$

The approximation is obtained from asymptotic expansion of erf (x). The parameter Q is defined as :

$$Q = \frac{v_{th} - b_{off}}{\sigma_{off}} = \frac{b_{on} - v_{th}}{\sigma_{on}}$$

$$Q = \frac{b_{on} - b_{off}}{\sigma_{on} + \sigma_{off}} \qquad ...(10.23)$$

and $$erf(x) = \frac{2}{\sqrt{\pi}}\int_0^x e^{-y^2}\, dy \qquad ...(10.24)$$

The factor Q is widely used to specify receiver performance because it is related to signal to noise ratio. It is required to achieve a specific bit error rate.

Special Case: Let us consider, when $\sigma_{off} = \sigma_{on} = \sigma$ and $b_{off} = 0$

so that $$b_{on} = v$$

Then from Eq. (10.23), we have that threshold voltage $v_{th} = \frac{v}{2}$ so that $Q = \frac{v}{2\sigma}$.

Since σ is usually known as rms noise, the ratio $\frac{v}{\sigma}$ is the peak signal to rms noise ratio. In this case Eq. (10.22) becomes

$$P_e(\sigma_{on} = \sigma_{off}) = \frac{1}{2}\left[1 - erf\left(\frac{v}{2\sqrt{2}\sigma}\right)\right] \qquad ...(10.25)$$

10.3.2. Receiver Sensitivities

Receiver sensitivity is useful to transfer from consideration of SNR to bit error rate (BER).

1. ASK Detection

The probability density function to make an error after a certain symbol has been sent can be modelled by a Gaussian function; the mean value will be the relative sent value, and its variance will be given by :

$$\sigma_N = \int_{-\infty}^{+\infty} \Phi_N(f) \cdot |H_r(f)|^2\, df \qquad ...(10.26)$$

where $\Phi_N(f)$ is the spectral density of the noise within the band and $H_r(f)$ is the continuous Fourier transform of the impulse response of the filter $h_r(f)$.

The possibility to make an error is given by :

$$P_E = P_{e/H_0} \cdot P_{H_0} + P_{e/H_1} \cdot P_{H_1} + ... + P_{e/H_{L-1}} \cdot P_{H_{L-1}} \qquad ...(10.27)$$

where P_{e/H_0} is the conditional probability of making an error after a symbol v_i has been sent and P_{H_0} is the probability of sending a symbol v_0.

If the probability of sending any symbol is the same, then :

$$P_{H_i} = \frac{1}{L}$$

If we represent all the probability density functions on the same plot against the possible value of the voltage to be transmitted, we get a picture like this (the particular case of $L = 4$ is shown) in Figs. 10.7 and 10.8.

The possibility of making an error after a single symbol has been sent is the area of the Gaussian function falling under the other ones. It is shown in cyan just for one of them. If we call P^+ the area under one side of the Gaussian, the sum of all the areas will be : $2LP^+ - 2P^+$. The total **probability of making an error** can be expressed in the form :

$$P_e = 2\left(1 - \frac{1}{L}\right)P^+ \qquad ...(10.42)$$

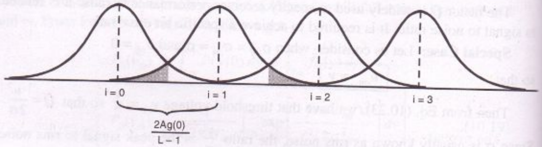


Fig. 10.7.

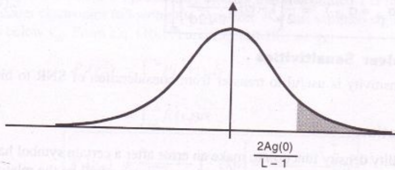


Fig. 10.8.

We have now to calculate the value of P^+. In order to do that, we can move the origin of the reference wherever we want: the area below the function will not change. We are in a situation like the one shown in the following picture :

It does not matter which Gaussian function we are considering, the area we want to calculate will be the same. The value we are looking for will be given by the following integral:

$$P^+ = \int_{\frac{A_g(0)}{L-1}}^{\infty} \frac{1}{\sqrt{2\pi}\sigma_N} e^{-\frac{x^2}{2\sigma_N^2}}\, dx$$

$$P^+ = \frac{1}{2}\, erfc\left(\frac{A_g(0)}{\sqrt{2}(L-1)\sigma_N}\right) \qquad ...(10.28)$$

where $erfc(\,)$ is the complementary error function. Putting all these results together, the probability to make an error is :

$$P_e = \left(1 - \frac{1}{L}\right) \text{erfc}\left(\frac{A_g(0)}{\sqrt{2}(L-1)\sigma_N}\right) \qquad ...(10.29)$$

from this formula it is clear that the probability to make an error decreases if the maximum amplitude of the transmitted signal or the amplification of the system becomes greater; on the other hand, it increases if the number of levels or the power of noise becomes greater.

2. FSK Detection

Let the two angular frequencies for the transmitted 1 and 0 bits are ω_1 and ω_2, so that

$$I_s(t) = \begin{cases} I_{SH}\cos(\omega_1 + \phi), & \text{for 1-bit} \\ I_{SH}\cos(\omega_2 + \phi), & \text{for 0-bit} \end{cases} \quad ...(10.30)$$

where,
$$I_{SH} = \frac{2\eta e}{hf}\sqrt{P_S P_L} \quad ...(10.31)$$

The **probability of error** in the shot noise for FSK heterodyne nonsynchronous or envelop detection is given by

$$P(e) = \frac{1}{2}\exp -\left(\frac{I_{SH}^2}{4\left(\overline{i_{SL}^2}\right)}\right) \quad ...(10.32)$$

Subtituting the value of I_{SH}^2 and $\overline{i_{SL}^2}$ in Eq. (10.31), we get

$$P(e) \simeq \frac{1}{2}\exp -\left(\frac{\eta P_S}{2hf\,B_T}\right) \quad ...(10.33)$$

3. PSK Detection

In this type of modulation format the information is transmitted by a carrier of one phase for a binary 1 and different phase for a binary 0.

Normally the employed phase shift is σ radians so that

$$I_S(t) = \begin{cases} I_{SH}\cos(\omega_{IF}t + \phi), & \text{for 1-bit} \\ I_{SH}\cos(\omega_{IF}t + \pi + \phi) & \\ \text{or} & \\ -I_{SH}\cos(\omega_{IF}t + \phi), & \text{for 0-bit} \end{cases} \quad ...(10.34)$$

Therefore in this modulation format, the **probability of error in detection** at the shot noise limit is given by

$$P(e) = \frac{1}{2}\exp -\left(\frac{I_{SH}^2}{2\left(\overline{i_{SL}^2}\right)}\right) \quad ...(10.35)$$

Subtituting the value of I_{SH} and $\overline{i_{SL}^2}$ in Eq. (10.35), we get

$$P(e) = \frac{1}{2}\exp -\left(\frac{\eta P_S}{hf\,B_T}\right) \quad ...(10.36)$$

10.3.3. The Quantum Limit

To design an optical system, it is helpful to know about fundamental physical bounds on system performance. Let us see what is this bound for the photodetection process.

Suppose we have an ideal photodetector which has unity quantum efficiency and produces no dark current (electron-hole pairs are generated in the presence of an optical pulse). Within this condition in a digital system it is possible to find the minimum received optical power required for specific bit-error rate performance. This minimum received power level is known as **quantum limit** because all system parameters are assumed ideal and the performance is limited only by photodetection statistics.

Let us assume that an optical pulse of energy E falls on the photodetector in a time interval τ. This can only be interpreted by the receiver as a 0 pulse if no electron hole pairs are generated with pulse present.

The probability that $n = 0$ electrons are emitted in a time interval t is

$$P_r(n) = \overline{N}^n \frac{e^{-\overline{N}}}{n!}$$

now putting $n = 0$ in above equation, we get

$$\boxed{P_r(0) = e^{-\overline{N}}} \quad ...(10.37)$$

where $\overline{N}$ = average number of electron-hole pairs

$\overline{N}$ is given as

$$\boxed{\overline{N} = \frac{\eta}{hv}\int_0^{\tau} P(t)\,dt = \frac{\eta E}{hv}} \quad ...(10.38)$$

Therefore, for a given error probability $P_r(0)$, we can find the minimum energy E required at a specific wavelength λ.

10.4. RECEIVER STRUCTURES

An optical receiver is treated as optical to electrical convertor. The receiver is designed with the aim to achieve maximum sensitivity for a given bit error rate (BER). Fig. 10.9 shows the block diagram of receiver.

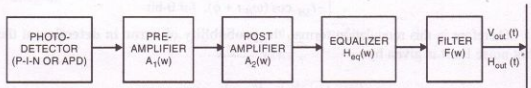


Fig. 10.9. Block diagram of a digital fiber optic receiver.

The signal voltage at the output of the linear channel is

$$V_s(w) = Z_T(w)\cdot I_s(w) \quad ...(10.39)$$

where, $Z_T(w)$ = Total system transfer function including eqalizer.

$I_s(w)$ = Fourier transform of $i_s(t)$.

$i_s(t)$ = Current produced due to electrons which are generated within the photodiode or detector.

Figure 10.10 shows the full equivalent circuit for the digital optical fiber receiver, where optical detector is represented as a current source i_{det}.

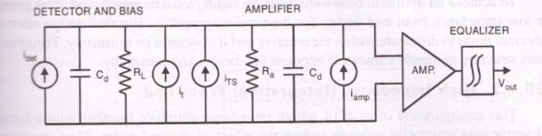


Fig. 10.10. A full equivalent circuit for a digital optical fiber receiver including the various noise sources.

The noise sources (i_t, i_{TS} and i_{amp}), the immediately following amplifier and equalizer are also shown in Fig. 10.5. Equalization compensates for distortion of the signal due to the combined transmitter, medium and receiver characteristics.

The equalizer is a **frequency shaping filter** which has a frequency response *i.e.*, the inverse of the overall system frequency response. The equalizer may also apply in wideband systems to boost the high frequency components to correct the overall amplitude of the frequency response. **Basically there are three receiver structures :**

1. Low Impedance front end.
2. High impedance front end.
3. The trans-impedance front end.

10.4.1. Low Impedance Front End

Figure 10.11 shows the simplest and most common receiver structure with voltage amplifier.

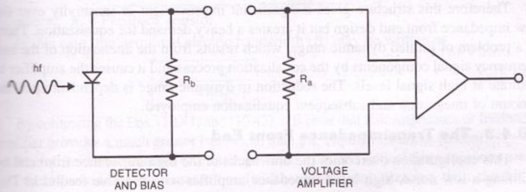


Fig. 10.11. Low impedance front end optical fiber receiver with voltage amplifier.

In order to make suitable design choices, it is required to considered both bandwidth and noise.

In most practical receivers the detector is loaded with a bias resistor R_b and an amplifier. The bandwidth is determined by the passive impedence which appears across the detector terminals which is taken as R_L. The resistance R_L may be modified to incorporate the parallel resistance of the detector bias R_b and the amplifier input resistance R_a. The total modified load resistance R_{TL} is given

$$R_{TL} = \frac{R_b\, R_a}{R_b + R_a} \qquad ...(10.40)$$

To achieve an optimum bandwidth both R_b and R_a must be minimized. This gives a low impedance front end design for the receiver amplifier. But this design allows thermal noise to dominate within the receiver and it may limit its sensitivity. Therefore this structure demands a trade off between bandwidth and sensitivity.

10.4.2. High Impedance (Integrating) Front End

This configuration consists of a high impedance amplifier together with a large detector bias resistor in order to reduce the effect of thermal noise. Thus structure gives a degraded frequency response. The detector output is effectively integrated over a large time constant and must be restored by differentiation. At a later stage, this may be performed by the correct equalization and it is shown in Fig. 10.12.

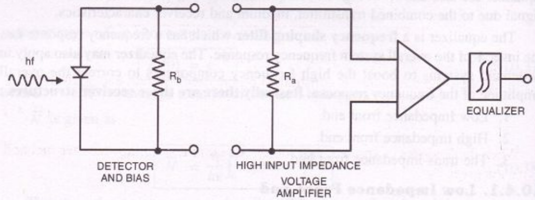


Fig. 10.12. High impedance integrating front end optical fiber receiver with equalized voltage amplifier.

Therefore this structure gives a significant improvement in sensitivity over the low impedance front end design but it creates a heavy demand for equalization. There is a problem of limited dynamic range, which results from the attenuation of the low frequency signal components by the equalization process and it causes the amplifier to saturate at high signal levels. The reduction in dynamic range is dependent upon the amount of integration and subsequent equalization employed.

10.4.3. The Transimpedance Front End

This configuration overcomes the drawbacks of the high impedance front end by utilizing a low noise, high input impedance amplifier with negative feedback. This device operates as a current mode amplifier where the high input impedance is reduced by negative feedback. Figure 10.13 shows its equivalent circuit.

In this equivalent circuit the parallel resistances and capacitances are combined into R_{TL} and C_T respectively. The open loop current to voltage transfer function $H_{OL}(w)$ for this configuration may be written

$$H_{OL}(w) = -G\frac{V_{in}}{i_{det}}$$

$$H_{OL}(w) = -G\frac{R_{TL}\cdot\dfrac{1}{jwC_T}}{R_{TL} + \dfrac{1}{jwC_T}} = \frac{-G.R_{TL}}{1 + jwR_{TL}C_T}\ (\text{VA}^{-1})$$

where, G = Open loop voltage gain of amplifier

w = Angular frequency of the input signal

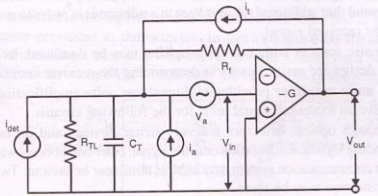


Fig. 10.13. An equivalent circuit for transimpedance front end optical receiver.

In this case the bandwidth is constrained by the time constant and is given by

$$\frac{1}{2\pi R_L C_T} \geq B \qquad \text{...(10.41)}$$

When the feedback is applied, the closed loop current to voltage transfer function $H_{CL}(w)$ is given by

$$H_{CL}(w) \cong \frac{R_f}{1+\left(\frac{jwR_fC_T}{G}\right)} \; (VA^{-1}) \qquad \text{...(10.42)}$$

where, R_f = Feedback resistor.

In this case, the permitted electric bandwidth (B) may be written as

$$B \leq \frac{G}{2\pi R_f C_T} \qquad \text{...(10.43)}$$

By comparing the Eqs. (10.41) and (10.43), it is clear that transimpedance or feedback amplifier provides a much greater bandwidth than the amplifiers without feedback.

Therefore the advantages of transimpedance front end optical fiber receivers are :

- It provides a far greater bandwidth without equalization than the high impedance front end.
- The transimpedance configuration has great dynamic range over the front end.

10.5. THE OPTICAL RECEIVER CIRCUIT

A block diagram of an optical fiber receiver is shown in Fig. 10.14.

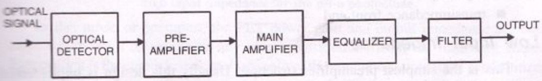


Fig. 10.14. Block diagram showing the major elements of an optical fiber receiver.

- A linear conversion of the received optical signal into an electrical current is performed at **detector.** Where it is amplified to obtain a suitable signal level.
- Initial amplification is performed in the **preamplifier** circuit where it is essential that additional noise is kept to a minimum in order to avoid corruption of the received signal.
- As noise sources within the preamplifier may be dominant, its configuration and design are major factors in determining the receiver sensitivity.
- The **main amplifier** provides additional low noise amplification of the signal to give an increased signal level for the following circuits.
- Although optical detectors are very lienar devices and do not themselves introduce significant distortion onto the signal, other components within the optical fiber communication system may exhibit nonlinear behaviour. Two major types of distortion may be there.
  (1) Due to the dispersive mechanism within the optical fiber.
  (2) Due to the transfer function of the preamplifier—main amplifier combination.
- Therefore to compensate for this distortion and to provide a suitable signal shape for the filter, an **equalizer** is sometimes included in the receiver.
- The **filter** is used to maximize the received signal to noise ratio while preserving the essential features of the signal. The filter is also designed to reduce the noise bandwidth as well as inband noise levels.
  (1) In digital systems, the filter is used to reduce the intersymbol interference.
  (2) In analog systems, generally the filter is required to hold the amplitude and the phase response of the received signal within certain limit.
- Sometimes a final element known as **linear channel** is used because all operations on the received optical signal may be considered to be mathematically linear.

10.5.1. The Preamplifier

For the preamplifier, the choice of circuit configuration is mainly dependent upon the system applicaton. Bipolar or field effect transistors (FETs) can be operated in three useful connections. These are :

- Common emitter or source
- Common base or gate
- The emitter or source follower

Each connection has characteristics which will contribute to a particular preamplifier configuration. Therefore it is useful to discuss the three basic preamplifier structures and indicate possible choices of transistor connection. The preamplifier structures are :

- low impedance
- high impedance
- transimpedance front end

Low Input Impedance

This is the simplest preamplfier structure. Usually this design is implemented using a BJT because of the high input impedance of FET. Figure 10.15 shows the common emitter and the grounded emitter amplifier configuration. This is favoured connection because it may be designed with reasonably low input impedance and gives operation over a moderate bandwidth without the need for equalization.

This is achieved at the expense of increased theremal noise due to the low effective load resistance presented to the detector. In this configuration the thermal noise is reduced by a choosing a transistor with characteristics which give a high current gain at a low emitter current in order to maintain the bandwidth of the stage.

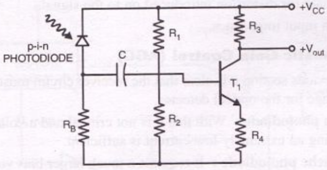


Fig. 10.15. A p-i-n photodiode with a grounded emitter low input impedance voltage preamplifier.

An inductance may be connected at the collector to provide partial equalization for any integration performed by the stage.

An alternative connection, wich gives very low input impedance is the common base circuit. Unfortunately its input impedance gives insufficient power gain when connected to the high impedance of the optical detector.

The same configuration with a common source connection, provides a similar high input impedance, is shown in Fig. 10.16.

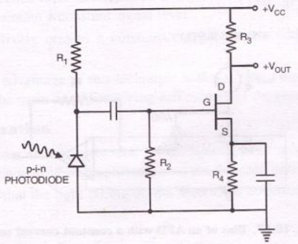


Fig. 10.16. An FET common source preamplifier configuration which provides high input impedance for the p-i-n photodiode.

In this mode of operation, the FET power gain and output impedance are both high, which tends to maximize any noise contributions from the following stages. Especially it is the case when the voltage gain of the common source stage is minimized in order to reduce the miller capacitance associated with the gate to drain capacitance of the FET.

This may be achieved by following the common source stage with a stage having a low input impedance.

High Input Impedance

The high impedance front end configuration provides a very low noise preamplifier design but suffers from two major drawbacks :

- Equalization : generally it must be tailored to the amplifier in order to compensate for distortion introduced on to the signal.
- The high input impedance.

10.5.2. Automatic Gain Control (AGC)

From the previous section it is clear that the receiver circuit must provide a steady reverse bias voltage for the optical detector.

With a p-i-n photodiode : With this it is not critical and a voltage of between 5 and 80 V supplying an extremely low current is sufficient.

With avalanche photodiode : It requires a much larger bias voltage of between 100 and 400 V, which define the multiplication factor for the device.

Usually an optimum multiplication factor is chosen so that the receiver signal to to noise ratio is maximized. The multiplication factor for the APD varies with the device temperature but it can be held constant by some form of automatic gain control (AGC).

Another advantages in the use of AGC is that it reduces the dynamic range of the signals applied to the preamplifier giving increased optical dynamic range at the receiver input.

There are two methods of providing AGC :

(*a*) **Method 1 :** It is simply to bias the APD with a constant d.c. current source I_{bias}, as shown in Fig. 10.17.

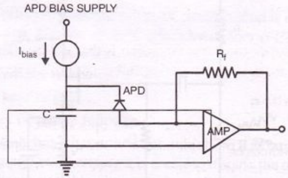


Fig. 10.17. Bias of an APD with a constant current source.

- The constant current source is decoupled by a capacitor C at all signal frequencies to prevent gain modulation.
- If mean optical input power is known, the mean current to the APD is defined by the bias which gives a constant multiplication factor (gain) at all temperatures.
- Any variation in the multiplication factor will produce a variation in the change on C, thus adjusting the biasing of the APD back to the required multiplication factor.

- Therefore the output current from the photodetector is only defined by the input current from the constant current source, giving full automatic gain control.
- This simple AGC technique is dependent on a constant, mean optical input power level, and takes no account of dark current generated within the detector.

(*b*) **Method 2 :** It is widely used method which allows the effect of variations in the detector dark current while providing critical AGC, as shown in Fig. 10.18.

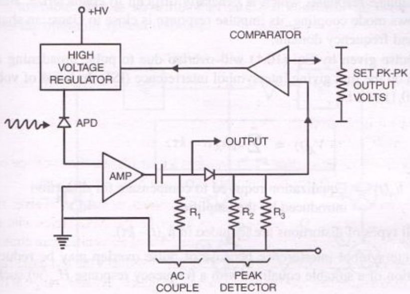


Fig. 10.18. Bias of an APD by peak detection and feedback to provide AGC.

- Here a signal from the final stage of the main amplifier is compared with a preset reference level and fedback to adjust the high voltage bias supply in order to maintain a constant signal level.
- This effectively creates a constant current source with the dark current subtracted.
- One more advantage of this technique is that it may also be used to provide AGC for the main amplifier giving full control of the receiver gain.

10.5.3. Equalization

Sometimes linear channel provided by the optical fiber receiver is required to perform equalization as well as amplification of the detected optical signal.

Let us assume that the light falling on the detector is consist of a series of pulses given by

$$P_0(t) = \sum_{k=-\infty}^{\infty} a_k h_P(t - k\tau) \quad ...(10.44)$$

where, $h_P(t)$ = Received pulse shape.

a_k = 0 or 1 corresponding to the binary information transmitted.,

τ = Pulse repetition time or pulse spacing.

(In digital transmission it is bit period)

All these parameters are important for point to point data links, local area networks (LANs), terrestrial communication and many more applications.

The **sensitivity** is an important parameter of receiver because the repeater spacing is directly dependent upon it. The repeater spacing *i.e.*, the spacing between the two repeaters may be increased if the receiver sensitivity and dynamic range are high. The receiver sensitivity itself depend on the amount of noise introduced by photodetector/ photodiode. A receiver should be capable of accepting power levels over a wide range of values of the input. The difference in decibel between maximum to minimum power range which a receiver can accept or detect is known as **dynamic range**. For a good receiver the dynamic range should be high.

10.6.1. p-n and p-i-n Photodiode Receiver

Basically in photodiodes there are two main sources of noise

- Dark Current Noise
- Quantum Noise

Both of which may be regarded as shot noise on the photocurrent. The total shot noise $\overline{i_{Ts}^2}$ is given by

$$\overline{i_{Ts}^2} = 2eB\left(I_p + I_d\right) \quad ...(10.48)$$

But the noise due to background radiation should also be considered, thus the above equation may be written as

$$\overline{i_{Ts}^2} = 2eB\left(I_p + I_d + I_b\right) \quad ...(10.49)$$

Since I_b is usually negligible, the expression given in Eq. (10.48) will be used in the further analysis.

In case of photodiode without internal avalanche gain, thermal noise from the detector load resistor and from active elements in the amplifier tends to dominate. For wideband systems operating in the 0.8 to 0.9 μm wavelength band, the dark currents can be made very small. The thermal noise $\overline{i_t^2}$ due to load resistor R_L may be given by

$$\overline{i_t^2} = \frac{4KTB}{R_L} \quad ...(10.50)$$

The **noise source** within the amplifier, associated with both the active and passive elements of the amplifier can be represented by a series voltage noise source $\overline{v_a^2}$ and a shunt current noise source $\overline{i_a^2}$.

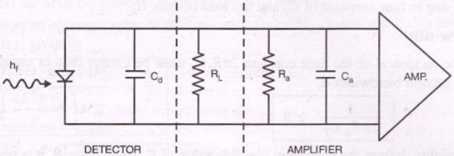
Thus the total noise associated with the amplifier $\overline{i_{amp}^2}$ is given by

$$\overline{i_{amp}^2} = \int_0^B \left(\overline{i_a^2} + \overline{v_a^2}\,|\gamma|^2\right) df \quad ...(10.51)$$

where, γ = Shunt admittance (combines the shunt capacitances and resistances)

f = Frequency

An equivalent circuit for the front end of the receiver is shown in Fig. 10.21.



C_d = Capacitance of detector

C_a = Effective input capacitance

R_a = Resistance of the amplifier

Fig. 10.21. The equivalent circuit for the front end of an optical fiber receiver.

The SNR for the p-n and p-i-n photodiode receiver may be obtained by summing all the noise contributions and is given by

$$\frac{S}{N} = \frac{I_p^2}{2eB(I_p + I_d) + \frac{4KTB}{R_L} + \overline{i_{amp}^2}} \quad ...(10.52)$$

- The thermal noise contribution may be reduced by inreasing the value of load resistor R_L and this reduction may be limited by bandwidth consideration.
- The amplifier noise $\overline{i_{amp}^2}$ may be reduced with low detector and amplifier capacitance.

The **noise figure** F_n for the amplifier may be obtained when amplifier noise $\overline{i_{amp}^2}$ is referred to the load resistor R_L. This allows $\overline{i_{amp}^2}$ to combined with the thermal noise from the load resistor $\overline{i_t^2}$ to give

$$\overline{i_t^2} + \overline{i_{amp}^2} = \frac{4KTBF_n}{R_L} \quad ...(10.53)$$

Therefore, from the Eq. (10.52) the expression for SNR may be rewritten as

$$\frac{S}{N} = \frac{I_p^2}{2eB(I_p + I_d) + \frac{4KTBF_n}{R_L}}$$

Thus if the noise figure F_n for the amplifier is known, SNR can be determined.

Receiver Capacitance

From the equivalent circuit shown in Fig 10.21, the total capacitance for the front end of an optical receiver C_T is given by

$$C_T = C_d + C_a \quad ...(10.54)$$

where, C_d = Detector capacitance

C_a = Amplifier input capacitance.

The total capacitance can also be minimized from the bandwidth penality which occurs due to time constant of C_T and the load resistor R_L.

Bandwidth

The reciprocal of the time constant $2\pi R_L C_T$ must be greater than or equal to the post detection bandwidth B,

$$\frac{1}{2\pi R_L C_T} \geq B \quad ...(10.55)$$

Equality defines the maximum possible value of B. To increase B, it is necessary to reduce R_L. However, this introduces a thermal noise penality.

10.6.2. Avalanche Photodiode (APD) Receiver

The signal current into the amplifier is increased by the internal gain mechanism in an APD receiver. This improves the SNR because the load resistance and amplifier noise remain unaffected.

The dark current and quantum noise are increased by the multiplication process. It become a limiting factor because the random gain mechanism introduces excess noise into the receiver. If the photocurrent is increased by a factor M (**mean avalanche multiplication factor**), then the shot noise is also increased by an excess noise factor M^x such that the total shot noise $\overline{i_{SA}^2}$ is given by

$$\overline{i_{SA}^2} = \left(2eB\, I_p + I_d\right) M^{2+x} \quad ...(10.56)$$

where, x = 0.3 to 0.5 (for silicon APDs)

= 0.7 to 1.0 (for germanium or III-V allow APDs)

The **SNR for the avalanche photodiode** may be obtained by summing the combined noise contribution from the load resistor and the amplifier, the SNR for the APD is given by

$$\frac{S}{N} = \frac{M^2 I_p^2}{2eB(I_p + I_d)M^{x+2} + \frac{4KTBF_n}{R_L}} \quad ...(10.57)$$

The relative significance of the combined thermal and amplifier noise term is reduced due to the avalanche multiplication of the shot noise term. Thus the Eq. (10.57) may be rewritten as

$$\frac{S}{N} = \frac{I_p^2}{2eB(I_p + I_d)M^x + \frac{4KTB\,F_n}{R_L} M^{-2}} \quad ...(10.58)$$

In the denominator the first term increases with increasing M whereas the second term decreases.

- For **low value of** M, when the signal level is increased the combined thermal and amplifier noise term dominates, the total noise power is virtually unaffected and it improves SNR.

- When M **is large,** the thermal and amplifier noise term becomes insignificant and the SNR decreases with increasing M at the rate of M^x.

Therefore an optimum value of multiplication factor M_{OP} exists which maximizes the SNR. It is given by

$$\frac{2eB(I_p + I_d)\, M_{op}^x}{\left(\frac{4KTB\, F_n}{R_L}\right) M_{op}^{-2}} = \frac{2}{x} \qquad ...(10.59)$$

and therefore

$$M_{op}^{x+2} = \frac{4KT\, F_n}{x\, e\, R_L\, (I_p + I_d)} \qquad ...(10.60)$$

10.6.3. Excess Avalanche Noise Factor

The value of excess noise factor is dependent upon :

- Detector material
- The shape of the electric field profile within the device.
- Whether the avalanche is initiated by holes or electrons.

Sometimes it is represented as $F(M)$. Let us consider one approximation for excess noise factor, where,

$$F(M) = M^x \qquad ...(10.61)$$

The resulting noise is assumed to be white with a Gaussian distribution. However, a second or more exact relationship is given by

$$F(M) = M\left[1 - (1-k)\left(\frac{M-1}{M}\right)^2\right] \qquad ...(10.62)$$

where injected electrons are the only carriers and k is the ratio of the ionization coefficients of holes and electrons.

If the only carries are injected holes, then

$$F(M) = M\left[1 + \left(\frac{1-k}{k}\right)\left[\frac{M-1}{M}\right]^2\right] \qquad ...(10.63)$$

when k is small, the best performance is achieved. The value of k for silicon APDs 0.02 and 0.10 where as for germanium and III-V alloy APDs k is between 0.3 and 1.0.

Solved Examples

Example 10.1. An InGaAs pin photodiode has the following parameters at a wavelength of 1100 nm :

I^*_D = 4 nA, η = 0.90, R_L = 1000 Ω and the surface leakage current is negligible. The incident optical power is 300 nW and the receiver bandwidth is 20 MHz. Find the value of quantum noise current, dark current and thermal noise current.

Solution. Given that,

$$I_D = 4 \text{ nA}$$