Chapter 5

Lightwave Systems

The preceding three chapters focused on the three main components of a fiber-optic communication system—optical fibers, optical transmitters, and optical receivers. In this chapter we consider the issues related to system design and performance when the three components are put together to form a practical lightwave system. Section 5.1 provides an overview of various system architectures. The design guidelines for fiber-optic communication systems are discussed in Section 5.2 by considering the effects of fiber losses and group-velocity dispersion. The power and the rise-time budgets are also described in this section. Section 5.3 focuses on long-haul systems for which the nonlinear effects become quite important. This section also covers various terrestrial and undersea lightwave systems that have been developed since 1977 when the first field trial was completed in Chicago. Issues related to system performance are treated in Section 5.4 with emphasis on performance degradation occurring as a result of signal transmission through the optical fiber. The physical mechanisms that can lead to power penalty in actual lightwave systems include modal noise, mode-partition noise, source spectral width, frequency chirp, and reflection feedback; each of them is discussed in separate subsections. In Section 5.5 we emphasize the importance of computer-aided design for lightwave systems.

5.1 System Architectures

From an architectural standpoint, fiber-optic communication systems can be classified into three broad categories—point-to-point links, distribution networks, and local-area networks [1]–[7]. This section focuses on the main characteristics of these three system architectures.

5.1.1 Point-to-Point Links

Point-to-point links constitute the simplest kind of lightwave systems. Their role is to transport information, available in the form of a digital bit stream, from one place to another as accurately as possible. The link length can vary from less than a kilometer
Figure 5.1: Point-to-point fiber links with periodic loss compensation through (a) regenerators and (b) optical amplifiers. A regenerator consists of a receiver followed by a transmitter.

(short haul) to thousands of kilometers (long haul), depending on the specific application. For example, optical data links are used to connect computers and terminals within the same building or between two buildings with a relatively short transmission distance (<10 km). The low loss and the wide bandwidth of optical fibers are not of primary importance for such data links; fibers are used mainly because of their other advantages, such as immunity to electromagnetic interference. In contrast, undersea lightwave systems are used for high-speed transmission across continents with a link length of several thousands of kilometers. Low losses and a large bandwidth of optical fibers are important factors in the design of transoceanic systems from the standpoint of reducing the overall operating cost.

When the link length exceeds a certain value, in the range 20–100 km depending on the operating wavelength, it becomes necessary to compensate for fiber losses, as the signal would otherwise become too weak to be detected reliably. Figure 5.1 shows two schemes used commonly for loss compensation. Until 1990, optoelectronic repeaters, called regenerators because they regenerate the optical signal, were used exclusively. As seen in Fig. 5.1(a), a regenerator is nothing but a receiver–transmitter pair that detects the incoming optical signal, recovers the electrical bit stream, and then converts it back into optical form by modulating an optical source. Fiber losses can also be compensated by using optical amplifiers, which amplify the optical bit stream directly without requiring conversion of the signal to the electric domain. The advent of optical amplifiers around 1990 revolutionized the development of fiber-optic communication systems [8]–[10]. Amplifiers are especially valuable for wavelength-division multiplexed (WDM) lightwave systems as they can amplify many channels simultaneously; Chapter 6 is devoted to them.

Optical amplifiers solve the loss problem but they add noise (see Chapter 6) and worsen the impact of fiber dispersion and nonlinearity since signal degradation keeps on accumulating over multiple amplification stages. Indeed, periodically amplified lightwave systems are often limited by fiber dispersion unless dispersion-compensation techniques (discussed in Chapter 7) are used. Optoelectronic repeaters do not suffer from this problem as they regenerate the original bit stream and thus effectively compensate for all sources of signal degradation automatically. An optical regenerator should perform the same three functions—reamplification, reshaping, and retiming.
5.1. SYSTEM ARCHITECTURES

5.1.1 Regenerators

The 3Rs—to replace an optoelectronic repeater. Although considerable research effort is being directed toward developing such all-optical regenerators [11], most terrestrial systems use a combination of the two techniques shown in Fig. 5.1 and place an optoelectronic regenerator after a certain number of optical amplifiers. Until 2000, the regenerator spacing was in the range of 600–800 km. Since then, ultralong-haul systems have been developed that are capable of transmitting optical signals over 3000 km or more without using a regenerator [12].

The spacing \( L \) between regenerators or optical amplifiers (see Fig. 5.1), often called the **repeater spacing**, is a major design parameter simply because the system cost reduces as \( L \) increases. However, as discussed in Section 2.4, the distance \( L \) depends on the bit rate \( B \) because of fiber dispersion. The bit rate–distance product, \( BL \), is generally used as a measure of the system performance for point-to-point links. The \( BL \) product depends on the operating wavelength, since both fiber losses and fiber dispersion are wavelength dependent. The first three generations of lightwave systems correspond to three different operating wavelengths near 0.85, 1.3, and 1.55 \( \mu \)m. Whereas the \( BL \) product was \( \sim 1 \) (Gb/s)-km for the first-generation systems operating near 0.85 \( \mu \)m, it becomes \( \sim 1 \) (Tb/s)-km for the third-generation systems operating near 1.55 \( \mu \)m and can exceed 100 (Tb/s)-km for the fourth-generation systems.

5.1.2 Distribution Networks

Many applications of optical communication systems require that information is not only transmitted but is also distributed to a group of subscribers. Examples include local-loop distribution of telephone services and broadcast of multiple video channels over cable television (CATV, short for common-antenna television). Considerable effort is directed toward the integration of audio and video services through a broadband integrated-services digital network (ISDN). Such a network has the ability to distribute a wide range of services, including telephone, facsimile, computer data, and video broadcasts. Transmission distances are relatively short (\( L < 50 \) km), but the bit rate can be as high as 10 Gb/s for a broadband ISDN.

Figure 5.2 shows two topologies for distribution networks. In the case of **hub topology**, channel distribution takes place at central locations (or hubs), where an automated cross-connect facility switches channels in the electrical domain. Such networks are called metropolitan-area networks (MANs) as hubs are typically located in major cities [13]. The role of fiber is similar to the case of point-to-point links. Since the fiber bandwidth is generally much larger than that required by a single hub office, several offices can share a single fiber headed for the main hub. Telephone networks employ hub topology for distribution of audio channels within a city. A concern for the hub topology is related to its reliability—outage of a single fiber cable can affect the service to a large portion of the network. Additional point-to-point links can be used to guard against such a possibility by connecting important hub locations directly.

In the case of **bus topology**, a single fiber cable carries the multichannel optical signal throughout the area of service. Distribution is done by using optical taps, which divert a small fraction of the optical power to each subscriber. A simple CATV application of bus topology consists of distributing multiple video channels within a city. The use of optical fiber permits distribution of a large number of channels (100 or more)
because of its large bandwidth compared with coaxial cables. The advent of high-definition television (HDTV) also requires lightwave transmission because of a large bandwidth (about 100 Mb/s) of each video channel unless a compression technique (such as MPEG-2, or 2nd recommendation of the motion-picture entertainment group) is used.

A problem with the bus topology is that the signal loss increases exponentially with the number of taps and limits the number of subscribers served by a single optical bus. Even when fiber losses are neglected, the power available at the Nth tap is given by

\[ P_N = P_T C [(1 - \delta) (1 - C)]^{N-1}, \]

where \( P_T \) is the transmitted power, \( C \) is the fraction of power coupled out at each tap, and \( \delta \) accounts for insertion losses, assumed to be the same at each tap. The derivation of Eq. (5.1.1) is left as an exercise for the reader. If we use \( \delta = 0.05, C = 0.05, P_T = 1 \text{ mW}, \text{and} P_N = 0.1 \mu \text{W} \) as illustrative values, \( N \) should not exceed 60. A solution to this problem is offered by optical amplifiers which can boost the optical power of the bus periodically and thus permit distribution to a large number of subscribers as long as the effects of fiber dispersion remain negligible.

### 5.1.3 Local-Area Networks

Many applications of fiber-optic communication technology require networks in which a large number of users within a local area (e.g., a university campus) are intercon-
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Figure 5.3: (a) Ring topology and (b) star topology for local-area networks.

connected in such a way that any user can access the network randomly to transmit data to any other user [14]–[16]. Such networks are called local-area networks (LANs). Optical-access networks used in a local subscriber loop also fall in this category [17]. Since the transmission distances are relatively short (<10 km), fiber losses are not of much concern for LAN applications. The major motivation behind the use of optical fibers is the large bandwidth offered by fiber-optic communication systems.

The main difference between MANs and LANs is related to the random access offered to multiple users of a LAN. The system architecture plays an important role for LANs, since the establishment of predefined protocol rules is a necessity in such an environment. Three commonly used topologies are known as bus, ring, and star configurations. The bus topology is similar to that shown in Fig. 5.2(b). A well-known example of bus topology is provided by the Ethernet, a network protocol used to connect multiple computers and used by the Internet. The Ethernet operates at speeds up to 1 Gb/s by using a protocol based on carrier-sense multiple access (CSMA) with collision detection. Although the Ethernet LAN architecture has proven to be quite successful when coaxial cables are used for the bus, a number of difficulties arise when optical fibers are used. A major limitation is related to the losses occurring at each tap, which limits the number of users [see Eq. (5.1.1)].

Figure 5.3 shows the ring and star topologies for LAN applications. In the ring
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topology \[18\], consecutive nodes are connected by point-to-point links to form a closed ring. Each node can transmit and receive the data by using a transmitter–receiver pair, which also acts as a repeater. A token (a predefined bit sequence) is passed around the ring. Each node monitors the bit stream to listen for its own address and to receive the data. It can also transmit by appending the data to an empty token. The use of ring topology for fiber-optic LANs has been commercialized with the standardized interface known as the fiber distributed data interface, FDDI for short \[18\]. The FDDI operates at 100 Mb/s by using multimode fibers and 1.3-\(\mu\)m transmitters based on light-emitting diodes (LEDs). It is designed to provide backbone services such as the interconnection of lower-speed LANs or mainframe computers.

In the star topology, all nodes are connected through point-to-point links to a central node called a hub, or simply a star. Such LANs are further subclassified as active-star or passive-star networks, depending on whether the central node is an active or passive device. In the active-star configuration, all incoming optical signals are converted to the electrical domain through optical receivers. The electrical signal is then distributed to drive individual node transmitters. Switching operations can also be performed at the central node since distribution takes place in the electrical domain. In the passive-star configuration, distribution takes place in the optical domain through devices such as directional couplers. Since the input from one node is distributed to many output nodes, the power transmitted to each node depends on the number of users. Similar to the case of bus topology, the number of users supported by passive-star LANs is limited by the distribution losses. For an ideal \(N \times N\) star coupler, the power reaching each node is simply \(P_T/N\) (if we neglect transmission losses) since the transmitted power \(P_T\) is divided equally among \(N\) users. For a passive star composed of directional couplers (see Section 8.2.4), the power is further reduced because of insertion losses and can be written as \[1\]

\[ P_N = (P_T/N)(1 - \delta)^{\log_2 N}, \]

where \(\delta\) is the insertion loss of each directional coupler. If we use \(\delta = 0.05\), \(P_T = 1\) mW, and \(P_N = 0.1\) \(\mu\)W as illustrative values, \(N\) can be as large as 500. This value of \(N\) should be compared with \(N = 60\) obtained for the case of bus topology by using Eq. (5.1.1). A relatively large value of \(N\) makes star topology attractive for LAN applications. The remainder of this chapter focuses on the design and performance of point-to-point links, which constitute a basic element of all communication systems, including LANs, MANS, and other distribution networks.

5.2 Design Guidelines

The design of fiber-optic communication systems requires a clear understanding of the limitations imposed by the loss, dispersion, and nonlinearity of the fiber. Since fiber properties are wavelength dependent, the choice of operating wavelength is a major design issue. In this section we discuss how the bit rate and the transmission distance of a single-channel system are limited by fiber loss and dispersion; Chapter 8 is devoted to multichannel systems. We also consider the power and rise-time budgets and illustrate them through specific examples \[5\]. The power budget is also called the link budget, and the rise-time budget is sometimes referred to as the bandwidth budget.
5.2. DESIGN GUIDELINES

5.2.1 Loss-Limited Lightwave Systems

Except for some short-haul fiber links, fiber losses play an important role in the system design. Consider an optical transmitter that is capable of launching an average power $\overline{P}_{\text{tr}}$. If the signal is detected by a receiver that requires a minimum average power $\overline{P}_{\text{rec}}$ at the bit rate $B$, the maximum transmission distance is limited by

$$L = \frac{10}{\alpha_f} \log_{10} \left( \frac{\overline{P}_{\text{tr}}}{\overline{P}_{\text{rec}}} \right),$$

where $\alpha_f$ is the net loss (in dB/km) of the fiber cable, including splice and connector losses. The bit-rate dependence of $L$ arises from the linear dependence of $\overline{P}_{\text{rec}}$ on the bit rate $B$. Noting that $\overline{P}_{\text{rec}} = \overline{N}_p h \nu B$, where $h \nu$ is the photon energy and $\overline{N}_p$ is the average number of photons/bit required by the receiver [see Eq. (4.5.24)], the distance $L$ decreases logarithmically as $B$ increases at a given operating wavelength.

The solid lines in Fig. 5.4 show the dependence of $L$ on $B$ for three common operating wavelengths of 0.85, 1.3, and 1.55 $\mu$m by using $\alpha_f = 2.5, 0.4,$ and 0.25 dB/km, respectively. The transmitted power is taken to be $\overline{P}_{\text{tr}} = 1$ mW at the three wavelengths, whereas $\overline{N}_p = 300$ at $\lambda = 0.85$ $\mu$m and $\overline{N}_p = 500$ at 1.3 and 1.55 $\mu$m. The smallest value of $L$ occurs for first-generation systems operating at 0.85 $\mu$m because of relatively large fiber losses near that wavelength. The repeater spacing of such systems is limited to 10–25 km, depending on the bit rate and the exact value of the loss parameter. In contrast, a repeater spacing of more than 100 km is possible for lightwave systems operating near 1.55 $\mu$m.

It is interesting to compare the loss limit of 0.85-$\mu$m lightwave systems with that of electrical communication systems based on coaxial cables. The dotted line in Fig. \[\text{Figure 5.4: Loss (solid lines) and dispersion (dashed lines) limits on transmission distance $L$ as a function of bit rate $B$ for the three wavelength windows. The dotted line corresponds to coaxial cables. Circles denote commercial lightwave systems; triangles show laboratory experiments. (After Ref. [1]; ©1988 Academic Press; reprinted with permission.)}\]
5.4 shows the bit-rate dependence of $L$ for coaxial cables by assuming that the loss increases as $\sqrt{B}$. The transmission distance is larger for coaxial cables at small bit rates ($B < 5$ Mb/s), but fiber-optic systems take over at bit rates in excess of 5 Mb/s. Since a longer transmission distance translates into a smaller number of repeaters in a long-haul point-to-point link, fiber-optic communication systems offer an economic advantage when the operating bit rate exceeds 10 Mb/s.

The system requirements typically specified in advance are the bit rate $B$ and the transmission distance $L$. The performance criterion is specified through the bit-error rate (BER), a typical requirement being $BER < 10^{-9}$. The first decision of the system designer concerns the choice of the operating wavelength. As a practical matter, the cost of components is lowest near 0.85 $\mu$m and increases as wavelength shifts toward 1.3 and 1.55 $\mu$m. Figure 5.4 can be quite helpful in determining the appropriate operating wavelength. Generally speaking, a fiber-optic link can operate near 0.85 $\mu$m if $B < 200$ Mb/s and $L < 20$ km. This is the case for many LAN applications. On the other hand, the operating wavelength is by necessity in the 1.55-$\mu$m region for long-haul lightwave systems operating at bit rates in excess of 2 Gb/s. The curves shown in Fig. 5.4 provide only a guide to the system design. Many other issues need to be addressed while designing a realistic fiber-optic communication system. Among them are the choice of the operating wavelength, selection of appropriate transmitters, receivers, and fibers, compatibility of various components, issue of cost versus performance, and system reliability and upgradability concerns.

### 5.2.2 Dispersion-Limited Lightwave Systems

In Section 2.4 we discussed how fiber dispersion limits the bit rate–distance product $BL$ because of pulse broadening. When the dispersion-limited transmission distance is shorter than the loss-limited distance of Eq. (5.2.1), the system is said to be dispersion-limited. The dashed lines in Fig. 5.4 show the dispersion-limited transmission distance as a function of the bit rate. Since the physical mechanisms leading to dispersion limitation can be different for different operating wavelengths, let us examine each case separately.

Consider first the case of 0.85-$\mu$m lightwave systems, which often use multimode fibers to minimize the system cost. As discussed in Section 2.1, the most limiting factor for multimode fibers is intermodal dispersion. In the case of step-index multimode fibers, Eq. (2.1.6) provides an approximate upper bound on the $BL$ product. A slightly more restrictive condition $BL = c/(2n_1 \Delta)$ is plotted in Fig. 5.4 by using typical values $n_1 = 1.46$ and $\Delta = 0.01$. Even at a low bit rate of 1 Mb/s, such multimode systems are dispersion-limited, and their transmission distance is limited to below 10 km. For this reason, multimode step-index fibers are rarely used in the design of fiber-optic communication systems. Considerable improvement can be realized by using graded-index fibers for which intermodal dispersion limits the $BL$ product to values given by Eq. (2.1.11). The condition $BL = 2c/(n_1 \Delta^2)$ is plotted in Fig. 5.4 and shows that 0.85-$\mu$m lightwave systems are loss-limited, rather than dispersion-limited, for bit rates up to 100 Mb/s when graded-index fibers are used. The first generation of terrestrial telecommunication systems took advantage of such an improvement and used graded-
The first commercial system became available in 1980 and operated at a bit rate of 45 Mb/s with a repeater spacing of less than 10 km.

The second generation of lightwave systems used primarily single-mode fibers near the minimum-dispersion wavelength occurring at about 1.31 \( \mu m \). The most limiting factor for such systems is dispersion-induced pulse broadening dominated by a relatively large source spectral width. As discussed in Section 2.4.3, the \( BL \) product is then limited by [see Eq. (2.4.26)]

\[ BL \leq \left( \frac{4|D|}{\sigma_\lambda} \right)^{-1}, \tag{5.2.2} \]

where \( \sigma_\lambda \) is the root-mean-square (RMS) width of the source spectrum. The actual value of \( |D| \) depends on how close the operating wavelength is to the zero-dispersion wavelength of the fiber and is typically \( \sim 1 \) ps/(km-nm). Figure 5.4 shows the dispersion limit for 1.3-\( \mu m \) lightwave systems by choosing \( |D|\sigma_\lambda = 2 \) ps/km so that \( BL \leq 125 \) (Gb/s)-km. As seen there, such systems are generally loss-limited for bit rates up to 1 Gb/s but become dispersion-limited at higher bit rates.

Third- and fourth-generation lightwave systems operate near 1.55 \( \mu m \) to take advantage of the smallest fiber losses occurring in this wavelength region. However, fiber dispersion becomes a major problem for such systems since \( D \approx 16 \) ps/(km-nm) near 1.55 \( \mu m \) for standard silica fibers. Semiconductor lasers operating in a single longitudinal mode provide a solution to this problem. The ultimate limit is then given by [see Eq. (2.4.30)]

\[ B^2L < \left( \frac{16|\beta_2|}{\beta_2} \right)^{-1}, \tag{5.2.3} \]

where \( \beta_2 \) is related to \( D \) as in Eq. (2.3.5). Figure 5.4 shows this limit by choosing \( B^2L = 4000 \) (Gb/s)^2-km. As seen there, such 1.55-\( \mu m \) systems become dispersion-limited only for \( B > 5 \) Gb/s. In practice, the frequency chirp imposed on the optical pulse during direct modulation provides a much more severe limitation. The effect of frequency chirp on system performance is discussed in Section 5.4.4. Qualitatively speaking, the frequency chirp manifests through a broadening of the pulse spectrum. If we use Eq. (5.2.2) with \( D = 16 \) ps/(km-nm) and \( \sigma_\lambda = 0.1 \) nm, the \( BL \) product is limited to 150 (Gb/s)-km. As a result, the frequency chirp limits the transmission distance to 75 km at \( B = 2 \) Gb/s, even though loss-limited distance exceeds 150 km. The frequency-chirp problem is often solved by using an external modulator for systems operating at bit rates >5 Gb/s.

A solution to the dispersion problem is offered by dispersion-shifted fibers for which dispersion and loss both are minimum near 1.55 \( \mu m \). Figure 5.4 shows the improvement by using Eq. (5.2.3) with \( |\beta_2| = 2 \) ps^2/km. Such systems can be operated at 20 Gb/s with a repeater spacing of about 80 km. Further improvement is possible by operating the lightwave system very close to the zero-dispersion wavelength, a task that requires careful matching of the laser wavelength to the zero-dispersion wavelength and is not always feasible because of variations in the dispersive properties of the fiber along the transmission link. In practice, the frequency chirp makes it difficult to achieve even the limit indicated in Fig. 5.4. By 1989, two laboratory experiments had demonstrated transmission over 81 km at 11 Gb/s [19] and over 100 km at 10 Gb/s [20] by using low-chirp semiconductor lasers together with dispersion-shifted fibers. The triangles in Fig. 5.4 show that such systems operate quite close to the fundamental...
limits set by fiber dispersion. Transmission over longer distances requires the use of dispersion-management techniques discussed in Chapter 7.

5.2.3 Power Budget

The purpose of the power budget is to ensure that enough power will reach the receiver to maintain reliable performance during the entire system lifetime. The minimum average power required by the receiver is the receiver sensitivity $P_{\text{rec}}$ (see Section 4.4). The average launch power $P_{\text{l}}$ is generally known for any transmitter. The power budget takes an especially simple form in decibel units with optical powers expressed in dBm units (see Appendix A). More specifically,

$$P_{\text{l}} = P_{\text{rec}} + C_L + M_s, \quad (5.2.4)$$

where $C_L$ is the total channel loss and $M_s$ is the system margin. The purpose of system margin is to allocate a certain amount of power to additional sources of power penalty that may develop during system lifetime because of component degradation or other unforeseen events. A system margin of 4–6 dB is typically allocated during the design process.

The channel loss $C_L$ should take into account all possible sources of power loss, including connector and splice losses. If $\alpha_f$ is the fiber loss in decibels per kilometer, $C_L$ can be written as

$$C_L = \alpha_f L + \alpha_{\text{con}} + \alpha_{\text{splice}}, \quad (5.2.5)$$

where $\alpha_{\text{con}}$ and $\alpha_{\text{splice}}$ account for the connector and splice losses throughout the fiber link. Sometimes splice loss is included within the specified loss of the fiber cable. The connector loss $\alpha_{\text{con}}$ includes connectors at the transmitter and receiver ends but must include other connectors if used within the fiber link.

Equations (5.2.4) and (5.2.5) can be used to estimate the maximum transmission distance for a given choice of the components. As an illustration, consider the design of a fiber link operating at 100 Mb/s and requiring a maximum transmission distance of 8 km. As seen in Fig. 5.4, such a system can be designed to operate near 0.85 µm provided that a graded-index multimode fiber is used for the optical cable. The operation near 0.85 µm is desirable from the economic standpoint. Once the operating wavelength is selected, a decision must be made about the appropriate transmitters and receivers. The GaAs transmitter can use a semiconductor laser or an LED as an optical source. Similarly, the receiver can be designed to use either a $p-i-n$ or an avalanche photodiode. Keeping the low cost in mind, let us choose a $p-i-n$ receiver and assume that it requires 2500 photons/bit on average to operate reliably with a BER below $10^{-9}$. Using the relation $P_{\text{rec}} = \bar{N}_p h \nu B$ with $\bar{N}_p = 2500$ and $B = 100$ Mb/s, the receiver sensitivity is given by $P_{\text{rec}} = -42$ dBm. The average launch power for LED and laser-based transmitters is typically 50 µW and 1 mW, respectively.

Table 5.1 shows the power budget for the two transmitters by assuming that the splice loss is included within the cable loss. The transmission distance $L$ is limited to 6 km in the case of LED-based transmitters. If the system specification is 8 km, a more expensive laser-based transmitter must be used. The alternative is to use an avalanche photodiode (APD) receiver. If the receiver sensitivity improves by more than 7 dB
Table 5.1 Power budget of a 0.85-μm lightwave system

<table>
<thead>
<tr>
<th>Quantity</th>
<th>Symbol</th>
<th>Laser</th>
<th>LED</th>
</tr>
</thead>
<tbody>
<tr>
<td>Transmitter power</td>
<td>$P_{tr}$</td>
<td>0 dBm</td>
<td>−13 dBm</td>
</tr>
<tr>
<td>Receiver sensitivity</td>
<td>$P_{rec}$</td>
<td>−42 dBm</td>
<td>−42 dBm</td>
</tr>
<tr>
<td>System margin</td>
<td>$M_s$</td>
<td>6 dB</td>
<td>6 dB</td>
</tr>
<tr>
<td>Available channel loss</td>
<td>$C_L$</td>
<td>36 dB</td>
<td>23 dB</td>
</tr>
<tr>
<td>Connector loss</td>
<td>$\alpha_{con}$</td>
<td>2 dB</td>
<td>2 dB</td>
</tr>
<tr>
<td>Fiber cable loss</td>
<td>$\alpha_f$</td>
<td>3.5 dB/km</td>
<td>3.5 dB/km</td>
</tr>
<tr>
<td>Maximum fiber length</td>
<td>$L$</td>
<td>9.7 km</td>
<td>6 km</td>
</tr>
</tbody>
</table>

when an APD is used in place of a p–i–n photodiode, the transmission distance can be increased to 8 km even for an LED-based transmitter. Economic considerations would then dictate the choice between the laser-based transmitters and APD receivers.

5.2.4 Rise-Time Budget

The purpose of the **rise-time budget** is to ensure that the system is able to operate properly at the intended bit rate. Even if the bandwidth of the individual system components exceeds the bit rate, it is still possible that the total system may not be able to operate at that bit rate. The concept of rise time is used to allocate the bandwidth among various components. The rise time $T_r$ of a linear system is defined as the time during which the response increases from 10 to 90% of its final output value when the input is changed abruptly. Figure 5.5 illustrates the concept graphically.

An inverse relationship exists between the bandwidth $\Delta f$ and the rise time $T_r$ associated with a linear system. This relationship can be understood by considering a simple $RC$ circuit as an example of the linear system. When the input voltage across an $RC$ circuit changes instantaneously from 0 to $V_0$, the output voltage changes as

$$V_{out}(t) = V_0[1 - \exp(-t/RC)], \quad (5.2.6)$$

where $R$ is the resistance and $C$ is the capacitance of the $RC$ circuit. The rise time is found to be given by

$$T_r = (\ln 9)RC \approx 2.2RC. \quad (5.2.7)$$

![Figure 5.5: Rise time $T_r$ associated with a bandwidth-limited linear system.](image)
The transfer function $H(f)$ of the $RC$ circuit is obtained by taking the Fourier transform of Eq. (5.2.6) and is of the form

$$H(f) = (1 + i2\pi fRC)^{-1}. \quad (5.2.8)$$

The bandwidth $\Delta f$ of the $RC$ circuit corresponds to the frequency at which $|H(f)|^2 = 1/2$ and is given by the well-known expression $\Delta f = (2\pi RC)^{-1}$. By using Eq. (5.2.7), $\Delta f$ and $T_r$ are related as

$$T_r = \frac{2.2}{2\pi \Delta f} = \frac{0.35}{\Delta f}. \quad (5.2.9)$$

The inverse relationship between the rise time and the bandwidth is expected to hold for any linear system. However, the product $T_r\Delta f$ would generally be different than 0.35. One can use $T_r\Delta f = 0.35$ in the design of optical communication systems as a conservative guideline. The relationship between the bandwidth $\Delta f$ and the bit rate $B$ depends on the digital format. In the case of return-to-zero (RZ) format (see Section 1.2), $\Delta f = B$ and $BT_r = 0.35$. By contrast, $\Delta f \approx B/2$ for the nonreturn-to-zero (NRZ) format and $BT_r = 0.7$. In both cases, the specified bit rate imposes an upper limit on the maximum rise time that can be tolerated. The communication system must be designed to ensure that $T_r$ is below this maximum value, i.e.,

$$T_r \leq \begin{cases} 0.35/B & \text{for RZ format}, \\ 0.70/B & \text{for NRZ format}. \end{cases} \quad (5.2.10)$$

The three components of fiber-optic communication systems have individual rise times. The total rise time of the whole system is related to the individual component rise times approximately as [21]

$$T_r^2 = T_{tr}^2 + T_{fiber}^2 + T_{rec}^2, \quad (5.2.11)$$

where $T_{tr}$, $T_{fiber}$, and $T_{rec}$ are the rise times associated with the transmitter, fiber, and receiver, respectively. The rise times of the transmitter and the receiver are generally known to the system designer. The transmitter rise time $T_{tr}$ is determined primarily by the electronic components of the driving circuit and the electrical parasitics associated with the optical source. Typically, $T_{tr}$ is a few nanoseconds for LED-based transmitters but can be shorter than 0.1 ns for laser-based transmitters. The receiver rise time $T_{rec}$ is determined primarily by the 3-dB electrical bandwidth of the receiver front end. Equation (5.2.9) can be used to estimate $T_{rec}$ if the front-end bandwidth is specified.

The fiber rise time $T_{fiber}$ should in general include the contributions of both the intermodal dispersion and group-velocity dispersion (GVD) through the relation

$$T_{fiber}^2 = T_{modal}^2 + T_{GVD}^2. \quad (5.2.12)$$

For single-mode fibers, $T_{modal} = 0$ and $T_{fiber} = T_{GVD}$. In principle, one can use the concept of fiber bandwidth discussed in Section 2.4.4 and relate $T_{fiber}$ to the 3-dB fiber bandwidth $f_{3dB}$ through a relation similar to Eq. (5.2.9). In practice it is not easy to calculate $f_{3dB}$, especially in the case of modal dispersion. The reason is that a fiber link consists of many concatenated fiber sections (typical length 5 km), which may have
5.3. LONG-HAUL SYSTEMS

different dispersion characteristics. Furthermore, mode mixing occurring at splices and connectors tends to average out the propagation delay associated with different modes of a multimode fiber. A statistical approach is often necessary to estimate the fiber bandwidth and the corresponding rise time [22]–[25].

In a phenomenological approach, $T_{\text{modal}}$ can be approximated by the time delay $\Delta T$ given by Eq. (2.1.5) in the absence of mode mixing, i.e.,

$$T_{\text{modal}} \approx (n_1 \Delta/c)L,$$  \hspace{1cm} (5.2.13)

where $n_1 \approx n_2$ was used. For graded-index fibers, Eq. (2.1.10) is used in place of Eq. (2.1.5), resulting in $T_{\text{modal}} \approx (n_1 \Delta^2/8c)L$. In both cases, the effect of mode mixing is included by changing the linear dependence on $L$ by a sublinear dependence $L^{q}$, where $q$ has a value in the range 0.5–1, depending on the extent of mode mixing. A reasonable estimate based on the experimental data is $q = 0.7$. The contribution $T_{\text{GVD}}$ can also be approximated by $\Delta T$ given by Eq. (2.3.4), so that

$$T_{\text{GVD}} \approx |D|L\Delta \lambda,$$  \hspace{1cm} (5.2.14)

where $\Delta \lambda$ is the spectral width of the optical source (taken as a full width at half maximum). The dispersion parameter $D$ may change along the fiber link if different sections have different dispersion characteristics; an average value should be used in Eq. (5.2.14) in that case.

As an illustration of the rise-time budget, consider a 1.3-µm lightwave system designed to operate at 1 Gb/s over a single-mode fiber with a repeater spacing of 50 km. The rise times for the transmitter and the receiver have been specified as $T_{\text{tr}} = 0.25$ ns and $T_{\text{rec}} = 0.35$ ns. The source spectral width is specified as $\Delta \lambda = 3$ nm, whereas the average value of $D$ is 2 ps/(km-nm) at the operating wavelength. From Eq. (5.2.14), $T_{\text{GVD}} = 0.3$ ns for a link length $L = 50$ km. Modal dispersion does not occur in single-mode fibers. Hence $T_{\text{modal}} = 0$ and $T_{\text{fiber}} = 0.3$ ns. The system rise time is estimated by using Eq. (5.2.11) and is found to be $T_r = 0.524$ ns. The use of Eq. (5.2.10) indicates that such a system cannot be operated at 1 Gb/s when the RZ format is employed for the optical bit stream. However, it would operate properly if digital format is changed to the NRZ format. If the use of RZ format is a prerequisite, the designer must choose different transmitters and receivers to meet the rise-time budget requirement. The NRZ format is often used as it permits a larger system rise time at the same bit rate.

5.3 Long-Haul Systems

With the advent of optical amplifiers, fiber losses can be compensated by inserting amplifiers periodically along a long-haul fiber link (see Fig. 5.1). At the same time, the effects of fiber dispersion (GVD) can be reduced by using dispersion management (see Chapter 7). Since neither the fiber loss nor the GVD is then a limiting factor, one may ask how many in-line amplifiers can be cascaded in series, and what limits the total link length. This topic is covered in Chapter 6 in the context of erbium-doped fiber amplifiers. Here we focus on the factors that limit the performance of amplified fiber links and provide a few design guidelines. The section also outlines the progress
realized in the development of terrestrial and undersea lightwave systems since 1977 when the first field trial was completed.

5.3.1 Performance-Limiting Factors

The most important consideration in designing a periodically amplified fiber link is related to the nonlinear effects occurring inside all optical fibers [26] (see Section 2.6). For single-channel lightwave systems, the dominant nonlinear phenomenon that limits the system performance is self-phase modulation (SPM). When optoelectronic regenerators are used, the SPM effects accumulate only over one repeater spacing (typically <100 km) and are of little concern if the launch power satisfies Eq. (2.6.15) or the condition $P_{in} \ll 22 \text{ mW}$ when $N_A = 1$. In contrast, the SPM effects accumulate over long lengths ($\sim$1000 km) when in-line amplifiers are used periodically for loss compensation. A rough estimate of the limitation imposed by the SPM is again obtained from Eq. (2.6.15). This equation predicts that the peak power should be below 2.2 mW for 10 cascaded amplifiers when the nonlinear parameter $\gamma = 2 \text{ W}^{-1}/\text{km}$. The condition on the average power depends on the modulation format and the shape of optical pulses. It is nonetheless clear that the average power should be reduced to below 1 mW for SPM effects to remain negligible for a lightwave system designed to operate over a distance of more than 1000 km. The limiting value of the average power also depends on the type of fiber in which light is propagating through the effective core area $A_{eff}$.

The SPM effects are most dominant inside dispersion-compensating fibers for which $A_{eff}$ is typically close to 20 $\mu\text{m}^2$.

The forgoing discussion of the SPM-induced limitations is too simplistic to be accurate since it completely ignores the role of fiber dispersion. In fact, as the dispersive and nonlinear effects act on the optical signal simultaneously, their mutual interplay becomes quite important [26]. The effect of SPM on pulses propagating inside an optical fiber can be included by using the nonlinear Schrödinger (NLS) equation of Section 2.6. This equation is of the form [see Eq. (2.6.18)]

$$\frac{\partial A}{\partial z} + i\frac{\beta_2}{2} \frac{\partial^2 A}{\partial t^2} = -\frac{\alpha}{2} A + i|A|^2 A,$$

(5.3.1)

where fiber losses are included through the $\alpha$ term. This term can also include periodic amplification of the signal by treating $\alpha$ as a function of $z$. The NLS equation is used routinely for designing modern lightwave systems.

Because of the nonlinear nature of Eq. (5.3.1), it should be solved numerically in general. A numerical approach has indeed been adopted (see Appendix E) since the early 1990s for quantifying the impact of SPM on the performance of long-haul lightwave systems [27]–[35]. The use of a large-effective-area fiber (LEAF) helps by reducing the nonlinear parameter $\gamma$ defined as $\gamma = 2\pi n_2/(\lambda A_{eff})$. Appropriate chirping of input pulses can also be beneficial for reducing the SPM effects. This feature has led to the adoption of a new modulation format known as the chirped RZ or CRZ format. Numerical simulations show that, in general, the launch power must be optimized to a value that depends on many design parameters such as the bit rate, total link length, and amplifier spacing. In one study, the optimum launch power was found to be about 1 mW for a 5-Gb/s signal transmitted over 9000 km with 40-km amplifier spacing [31].
The combined effects of GVD and SPM also depend on the sign of the dispersion parameter $\beta_2$. In the case of anomalous dispersion ($\beta_2 < 0$), the nonlinear phenomenon of modulation instability [26] can affect the system performance drastically [32]. This problem can be overcome by using a combination of fibers with normal and anomalous GVD such that the average dispersion over the entire fiber link is “normal.” However, a new kind of modulation instability, referred to as sideband instability [36], can occur in both the normal and anomalous GVD regions. It has its origin in the periodic variation of the signal power along the fiber link when equally spaced optical amplifiers are used to compensate for fiber losses. Since the quantity $\gamma |A|^2$ in Eq. (5.3.1) is then a periodic function of $z$, the resulting nonlinear-index grating can initiate a four-wave-mixing process that generates sidebands in the signal spectrum. It can be avoided by making the amplifier spacing nonuniform.

Another factor that plays a crucial role is the noise added by optical amplifiers. Similar to the case of electronic amplifiers (see Section 4.4), the noise of optical amplifiers is quantified through an amplifier noise figure $F_n$ (see Chapter 6). The nonlinear interaction between the amplified spontaneous emission and the signal can lead to a large spectral broadening through the nonlinear phenomena such as cross-phase modulation and four-wave mixing [37]. Because the noise has a much larger bandwidth than the signal, its impact can be reduced by using optical filters. Numerical simulations indeed show a considerable improvement when optical filters are used after every in-line amplifier [31].

The polarization effects that are totally negligible in the traditional “nonamplified” lightwave systems become of concern for long-haul systems with in-line amplifiers. The polarization-mode dispersion (PMD) issue has been discussed in Section 2.3.5. In addition to PMD, optical amplifiers can also induce polarization-dependent gain and loss [30]. Although the PMD effects must be considered during system design, their impact depends on the design parameters such as the bit rate and the transmission distance. For bit rates as high as 10-Gb/s, the PMD effects can be reduced to an acceptable level with a proper design. However, PMD becomes of major concern for 40-Gb/s systems for which the bit slot is only 25 ps wide. The use of a PMD-compensation technique appears to be necessary at such high bit rates.

The fourth generation of lightwave systems began in 1995 when lightwave systems employing amplifiers first became available commercially. Of course, the laboratory demonstrations began as early as 1989. Many experiments used a recirculating fiber loop to demonstrate system feasibility as it was not practical to use long lengths of fiber in a laboratory setting. Already in 1991, an experiment showed the possibility of data transmission over 21,000 km at 2.5 Gb/s, and over 14,300 km at 5 Gb/s, by using the recirculating-loop configuration [38]. In a system trial carried out in 1995 by using actual submarine cables and repeaters [39], a 5.3-Gb/s signal was transmitted over 11,300 km with 60 km of amplifier spacing. This system trial led to the deployment of a commercial transpacific cable (TPC–5) that began operating in 1996.

The bit rate of fourth-generation systems was extended to 10 Gb/s beginning in 1992. As early as 1995, a 10-Gb/s signal was transmitted over 6480 km with 90-km amplifier spacing [40]. With a further increase in the distance, the SNR decreased below the value needed to maintain the BER below $10^{-9}$. One may think that the performance should improve by operating close to the zero-dispersion wavelength of the
Table 5.2 Terrestrial lightwave systems

<table>
<thead>
<tr>
<th>System</th>
<th>Year</th>
<th>$\lambda$ (µm)</th>
<th>B (Mb/s)</th>
<th>L (km)</th>
<th>Voice Channels</th>
</tr>
</thead>
<tbody>
<tr>
<td>FT–3</td>
<td>1980</td>
<td>0.85</td>
<td>45</td>
<td>&lt; 10</td>
<td>672</td>
</tr>
<tr>
<td>FT–3C</td>
<td>1983</td>
<td>0.85</td>
<td>90</td>
<td>&lt; 15</td>
<td>1,344</td>
</tr>
<tr>
<td>FT–3X</td>
<td>1984</td>
<td>1.30</td>
<td>180</td>
<td>&lt; 25</td>
<td>2,688</td>
</tr>
<tr>
<td>FT–G</td>
<td>1985</td>
<td>1.30</td>
<td>417</td>
<td>&lt; 40</td>
<td>6,048</td>
</tr>
<tr>
<td>FT–G-1.7</td>
<td>1987</td>
<td>1.30</td>
<td>1,668</td>
<td>&lt; 46</td>
<td>24,192</td>
</tr>
<tr>
<td>STM–16</td>
<td>1991</td>
<td>1.55</td>
<td>2,488</td>
<td>&lt; 85</td>
<td>32,256</td>
</tr>
<tr>
<td>STM–64</td>
<td>1996</td>
<td>1.55</td>
<td>9,953</td>
<td>&lt; 90</td>
<td>129,024</td>
</tr>
<tr>
<td>STM–256</td>
<td>2002</td>
<td>1.55</td>
<td>39,813</td>
<td>&lt; 90</td>
<td>516,096</td>
</tr>
</tbody>
</table>

fiber. However, an experiment, performed under such conditions, achieved a distance of only 6000 km at 10 Gb/s even with 40-km amplifier spacing [41], and the situation became worse when the RZ modulation format was used. Starting in 1999, the single-channel bit rate was pushed toward 40 Gb/s in several experiments [42]–[44]. The design of 40-Gb/s lightwave systems requires the use of several new ideas including the CRZ format, dispersion management with GVD-slope compensation, and distributed Raman amplification. Even then, the combined effects of the higher-order dispersion, PMD, and SPM degrade the system performance considerably at a bit rate of 40 Gb/s.

5.3.2 Terrestrial Lightwave Systems

An important application of fiber-optic communication links is for enhancing the capacity of telecommunication networks worldwide. Indeed, it is this application that started the field of optical fiber communications in 1977 and has propelled it since then by demanding systems with higher and higher capacities. Here we focus on the status of commercial systems by considering the terrestrial and undersea systems separately.

After a successful Chicago field trial in 1977, terrestrial lightwave systems became available commercially beginning in 1980 [45]–[47]. Table 5.2 lists the operating characteristics of several terrestrial systems developed since then. The first-generation systems operated near 0.85 µm and used multimode graded-index fibers as the transmission medium. As seen in Fig. 5.4, the $BL$ product of such systems is limited to 2 (Gb/s)-km. A commercial lightwave system (FT–3C) operating at 90 Mb/s with a repeater spacing of about 12 km realized a $BL$ product of nearly 1 (Gb/s)-km; it is shown by a filled circle in Fig. 5.4. The operating wavelength moved to 1.3 µm in second-generation lightwave systems to take advantage of low fiber losses and low dispersion near this wavelength. The $BL$ product of 1.3-µm lightwave systems is limited to about 100 (Gb/s)-km when a multimode semiconductor laser is used inside the transmitter. In 1987, a commercial 1.3-µm lightwave system provided data transmission at 1.7 Gb/s with a repeater spacing of about 45 km. A filled circle in Fig. 5.4 shows that this system operates quite close to the dispersion limit.
The third generation of lightwave systems became available commercially in 1991. They operate near 1.55 $\mu$m at bit rates in excess of 2 Gb/s, typically at 2.488 Gb/s, corresponding to the OC-48 level of the synchronized optical network (SONET) [or the STS–16 level of the synchronous digital hierarchy (SDH)] specifications. The switch to the 1.55-$\mu$m wavelength helps to increase the loss-limited transmission distance to more than 100 km because of fiber losses of less than 0.25 dB/km in this wavelength region. However, the repeater spacing was limited to below 100 km because of the high GVD of standard telecommunication fibers. In fact, the deployment of third-generation lightwave systems was possible only after the development of distributed feedback (DFB) semiconductor lasers, which reduce the impact of fiber dispersion by reducing the source spectral width to below 100 MHz (see Section 2.4).

The fourth generation of lightwave systems appeared around 1996. Such systems operate in the 1.55-$\mu$m region at a bit rate as high as 40 Gb/s by using dispersion-shifted fibers in combination with optical amplifiers. However, more than 50 million kilometers of the standard telecommunication fiber is already installed in the worldwide telephone network. Economic reasons dictate that the fourth generation of lightwave systems make use of this existing base. Two approaches are being used to solve the dispersion problem. First, several dispersion-management schemes (discussed in Chapter 7) make it possible to extend the bit rate to 10 Gb/s while maintaining an amplifier spacing of up to 100 km. Second, several 10-Gb/s signals can be transmitted simultaneously by using the WDM technique discussed in Chapter 8. Moreover, if the WDM technique is combined with dispersion management, the total transmission distance can approach several thousand kilometers provided that fiber losses are compensated periodically by using optical amplifiers. Such WDM lightwave systems were deployed commercially worldwide beginning in 1996 and allowed a system capacity of 1.6 Tb/s by 2000 for the 160-channel commercial WDM systems.

The fifth generation of lightwave systems was just beginning to emerge in 2001. The bit rate of each channel in this generation of WDM systems is 40 Gb/s (corresponding to the STM-256 or OC–768 level). Several new techniques developed in recent years make it possible to transmit a 40-Gb/s optical signal over long distances. New fibers known as reverse-dispersion fibers have been developed with a negative GVD slope. Their use in combination with tunable dispersion-compensating techniques can compensate the GVD for all channels simultaneously. The PMD compensation helps to reduce the PMD-induced degradation of the signal. The use of Raman amplification helps to reduce the noise and improves the signal-to-noise ratio (SNR) at the receiver. The use of a forward-error-correction technique helps to increase the transmission distance by reducing the required SNR. The number of WDM channels can be increased by using the L and S bands located on the long- and short-wavelength sides of the conventional C band occupying the 1530–1570-nm spectral region. In one 3-Tb/s experiment, 77 channels, each operating at 42.7-Gb/s, were transmitted over 1200 km by using the C and L bands simultaneously [48]. In another experiment, the system capacity was extended to 10.2 Tb/s by transmitting 256 channels over 100 km at 42.7 Gb/s per channel using only the C and L bands, resulting in a spectral efficiency of 1.28 (b/s)/Hz [49]. The bit rate was 42.7 Gb/s in both of these experiments because of the overhead associated with the forward-error-correction technique. The highest capacity achieved in 2001 was 11 Tb/s and was realized by transmitting 273 channels
Table 5.3 Commercial transatlantic lightwave systems

<table>
<thead>
<tr>
<th>System</th>
<th>Year</th>
<th>Capacity (Gb/s)</th>
<th>(L) (km)</th>
<th>Comments</th>
</tr>
</thead>
<tbody>
<tr>
<td>TAT–8</td>
<td>1988</td>
<td>0.28</td>
<td>70</td>
<td>1.3 (\mu)m, multimode lasers</td>
</tr>
<tr>
<td>TAT–9</td>
<td>1991</td>
<td>0.56</td>
<td>80</td>
<td>1.55 (\mu)m, DFB lasers</td>
</tr>
<tr>
<td>TAT–10/11</td>
<td>1993</td>
<td>0.56</td>
<td>80</td>
<td>1.55 (\mu)m, DFB lasers</td>
</tr>
<tr>
<td>TAT–12/13</td>
<td>1996</td>
<td>5.00</td>
<td>50</td>
<td>1.55 (\mu)m, optical amplifiers</td>
</tr>
<tr>
<td>AC–1</td>
<td>1998</td>
<td>80.0</td>
<td>50</td>
<td>1.55 (\mu)m, WDM with amplifiers</td>
</tr>
<tr>
<td>TAT–14</td>
<td>2001</td>
<td>1280</td>
<td>50</td>
<td>1.55 (\mu)m, dense WDM</td>
</tr>
<tr>
<td>AC–2</td>
<td>2001</td>
<td>1280</td>
<td>50</td>
<td>1.55 (\mu)m, dense WDM</td>
</tr>
<tr>
<td>360Atlantic-1</td>
<td>2001</td>
<td>1920</td>
<td>50</td>
<td>1.55 (\mu)m, dense WDM</td>
</tr>
<tr>
<td>Tycom</td>
<td>2001</td>
<td>2560</td>
<td>50</td>
<td>1.55 (\mu)m, dense WDM</td>
</tr>
<tr>
<td>FLAG Atlantic-1</td>
<td>2001</td>
<td>4800</td>
<td>50</td>
<td>1.55 (\mu)m, dense WDM</td>
</tr>
</tbody>
</table>

over 117 km at 40 Gb/s per channel while using all three bands simultaneously [50].

5.3.3 Undersea Lightwave Systems

Undersea or submarine transmission systems are used for intercontinental communications and are capable of providing a network spanning the whole earth [51]–[53]. Figure 1.5 shows several undersea systems deployed worldwide. Reliability is of major concern for such systems as repairs are expensive. Generally, undersea systems are designed for a 25-year service life, with at most three failures during operation. Table 5.3 lists the main characteristics of several transatlantic fiber-optic cable systems. The first undersea fiber-optic cable (TAT–8) was a second-generation system. It was installed in 1988 in the Atlantic Ocean for operation at a bit rate of 280 Mb/s with a repeater spacing of up to 70 km. The system design was on the conservative side, mainly to ensure reliability. The same technology was used for the first transpacific lightwave system (TPC–3), which became operational in 1989.

By 1990 the third-generation lightwave systems had been developed. The TAT–9 submarine system used this technology in 1991; it was designed to operate near 1.55 \(\mu\)m at a bit rate of 560 Mb/s with a repeater spacing of about 80 km. The increasing traffic across the Atlantic Ocean led to the deployment of the TAT–10 and TAT–11 lightwave systems by 1993 with the same technology. The advent of optical amplifiers prompted their use in the next generation of undersea systems, and the TAT–12 submarine fiber-optic cable became operational by 1996. This fourth-generation system employed optical amplifiers in place of optoelectronic regenerators and operated at a bit rate of 5.3 Gb/s with an amplifier spacing of about 50 km. The bit rate is slightly larger than the STM-32-level bit rate of 5 Gb/s because of the overhead associated with the forward-error-correction technique. As discussed earlier, the design of such lightwave systems is much more complex than that of previous undersea systems because of the cumulative effects of fiber dispersion and nonlinearity, which must be controlled over long distances. The transmitter power and the dispersion profile along the link must be
optimized to combat such effects. Even then, amplifier spacing is typically limited to 50 km, and the use of an error-correction scheme is essential to ensure a bit-error rate of $< 2 \times 10^{-11}$.

A second category of undersea lightwave systems requires repeaterless transmission over several hundred kilometers [52]. Such systems are used for interisland communication or for looping a shoreline such that the signal is regenerated on the shore periodically after a few hundred kilometers of undersea transmission. The dispersive and nonlinear effects are of less concern for such systems than for transoceanic lightwave systems, but fiber losses become a major issue. The reason is easily appreciated by noting that the cable loss exceeds 100 dB over a distance of 500 km even under the best operating conditions. In the 1990s several laboratory experiments demonstrated repeaterless transmission at 2.5 Gb/s over more than 500 km by using two in-line amplifiers that were pumped remotely from the transmitter and receiver ends with high-power pump lasers. Another amplifier at the transmitter boosted the launched power to close to 100 mW.

Such high input powers exceed the threshold level for stimulated Brillouin scattering (SBS), a nonlinear phenomenon discussed in Section 2.6. The suppression of SBS is often realized by modulating the phase of the optical carrier such that the carrier linewidth is broadened to 200 MHz or more from its initial value of $<10$ MHz [54]. Directly modulated DFB lasers can also be used for this purpose. In a 1996 experiment, a 2.5-Gb/s signal was transmitted over 465 km by direct modulation of a DFB laser [55]. Chirping of the modulated signal broadened the spectrum enough that an external phase modulator was not required provided that the launched power was kept below 100 mW. The bit rate of repeaterless undersea systems can be increased to 10 Gb/s by employing the same techniques used at 2.5 Gb/s. In a 1996 experiment [56], the 10-Gb/s signal was transmitted over 442 km by using two remotely pumped in-line amplifiers. Two external modulators were used, one for SBS suppression and another for signal generation. In a 1998 experiment, a 40-Gb/s signal was transmitted over 240 km using the RZ format and an alternating polarization format [57]. These results indicate that undersea lightwave systems looping a shoreline can operate at 10 Gb/s or more with only shore-based electronics [58].

The use of the WDM technique in combination with optical amplifiers, dispersion management, and error correction has revolutionized the design of submarine fiber-optic systems. In 1998, a submarine cable known as Atlantic-Crossing 1 (AC–1) with a capacity of 80 Gb/s was deployed using the WDM technology. An identically designed system (Pacific-Crossing 1 or PC–1) crossed the Pacific Ocean. The use of dense WDM, in combination with multiple fiber pairs per cable, resulted in systems with much larger capacities. By 2001, several systems with a capacity of $>1$ Tb/s became operational across the Atlantic Ocean (see Table 5.3). These systems employ a ring configuration and cross the Atlantic Ocean twice to ensure fault tolerance. The “360Atlantic” submarine system can operate at speeds up to 1.92 Tb/s and spans a total distance of 11,700 km. Another system, known as FLAG Atlantic-1, is capable of carrying traffic at speeds up to 4.8 Tb/s as it employs six fiber pairs. A global network, spanning 250,000 km and capable of operating at 3.2 Tb/s using 80 channels (at 10 Gb/s) over 4 fibers, was under development in 2001 [53]. Such a submarine network can transmit nearly 40 million voice channels simultaneously, a capacity that should be
contrast with the TAT–8 capacity of 8000 channels in 1988, which in turn should be compared to the 48-channel capacity of TAT–1 in 1959.

5.4 Sources of Power Penalty

The sensitivity of the optical receiver in a realistic lightwave system is affected by several physical phenomena which, in combination with fiber dispersion, degrade the SNR at the decision circuit. Among the phenomena that degrade the receiver sensitivity are modal noise, dispersion broadening and intersymbol interference, mode-partition noise, frequency chirp, and reflection feedback. In this section we discuss how the system performance is affected by fiber dispersion by considering the extent of power penalty resulting from these phenomena.

5.4.1 Modal Noise

Modal noise is associated with multimode fibers and was studied extensively during the 1980s [59]–[72]. Its origin can be understood as follows. Interference among various propagating modes in a multimode fiber creates a speckle pattern at the photodetector. The nonuniform intensity distribution associated with the speckle pattern is harmless in itself, as the receiver performance is governed by the total power integrated over the detector area. However, if the speckle pattern fluctuates with time, it will lead to fluctuations in the received power that would degrade the SNR. Such fluctuations are referred to as modal noise. They invariably occur in multimode fiber links because of mechanical disturbances such as vibrations and microbends. In addition, splices and connectors act as spatial filters. Any temporal changes in spatial filtering translate into speckle fluctuations and enhancement of the modal noise. Modal noise is strongly affected by the source spectral bandwidth $\Delta \nu$ since mode interference occurs only if the coherence time ($T_c \approx 1/\Delta \nu$) is longer than the intermodal delay time $\Delta T$ given by Eq. (2.1.5). For LED-based transmitters $\Delta \nu$ is large enough ($\Delta \nu \sim 5$ THz) that this condition is not satisfied. Most lightwave systems that use multimode fibers also use LEDs to avoid the modal-noise problem.

Modal noise becomes a serious problem when semiconductor lasers are used in combination with multimode fibers. Attempts have been made to estimate the extent of sensitivity degradation induced by modal noise [61]–[63] by calculating the BER after adding modal noise to the other sources of receiver noise. Figure 5.6 shows the power penalty at a BER of $10^{-12}$ calculated for a 1.3-$\mu$m lightwave system operating at 140 Mb/s. The graded-index fiber has a 50-$\mu$m core diameter and supports 146 modes. The power penalty depends on the mode-selective coupling loss occurring at splices and connectors. It also depends on the longitudinal-mode spectrum of the semiconductor laser. As expected, power penalty decreases as the number of longitudinal modes increases because of a reduction in the coherence time of the emitted light.

Modal noise can also occur in single-mode systems if short sections of fiber are installed between two connectors or splices during repair or normal maintenance [63]–[66]. A higher-order mode can be excited at the fiber discontinuity occurring at the first splice and then converted back to the fundamental mode at the second connector.
5.4. SOURCES OF POWER PENALTY

Figure 5.6: Modal-noise power penalty versus mode-selective loss. The parameter $M$ is defined as the total number of longitudinal modes whose power exceeds 10% of the peak power. (After Ref. [61]; ©1986 IEEE; reprinted with permission.)

or splice. Since a higher-order mode cannot propagate far from its excitation point, this problem can be avoided by ensuring that the spacing between two connectors or splices exceeds 2 m. Generally speaking, modal noise is not a problem for properly designed and maintained single-mode fiber-optic communication systems.

With the development of the vertical-cavity surface-emitting laser (VCSEL), the modal-noise issue has resurfaced in recent years [67]–[71]. The use of such lasers in short-haul optical data links, making use of multimode fibers (even those made of plastic), is of considerable interest because of the high bandwidth associated with VCSELs. Indeed, rates of several gigabits per second have been demonstrated in laboratory experiments with plastic-cladded multimode fibers [73]. However, VCSELs have a long coherence length as they oscillate in a single longitudinal mode. In a 1994 experiment the BER measurements showed an error floor at a level of $10^{-7}$ even when the mode-selective loss was only 1 dB [68]. This problem can be avoided to some extent by using larger-diameter VCSELs which oscillate in several transverse modes and thus have a shorter coherence length. Computer models are generally used to estimate the power penalty for optical data links under realistic operating conditions [70]. Analytic tools such as the saddle-point method can also provide a reasonable estimate of the BER [71].
5.4.2 Dispersive Pulse Broadening

The use of single-mode fibers for lightwave systems nearly avoids the problem of intermodal dispersion and the associated modal noise. The group-velocity dispersion still limits the bit rate–distance product $BL$ by broadening optical pulses beyond their allocated bit slot; Eq. (5.2.2) provides the limiting $BL$ product and shows how it depends on the source spectral width $\sigma_\lambda$. Dispersion-induced pulse broadening can also decrease the receiver sensitivity. In this subsection we discuss the power penalty associated with such a decrease in receiver sensitivity.

Dispersion-induced pulse broadening affects the receiver performance in two ways. First, a part of the pulse energy spreads beyond the allocated bit slot and leads to intersymbol interference (ISI). In practice, the system is designed to minimize the effect of ISI (see Section 4.3.2). Second, the pulse energy within the bit slot is reduced when the optical pulse broadens. Such a decrease in the pulse energy reduces the SNR at the decision circuit. Since the SNR should remain constant to maintain the system performance, the receiver requires more average power. This is the origin of dispersion-induced power penalty $\delta_d$. An exact calculation of $\delta_d$ is difficult, as it depends on many details, such as the extent of pulse shaping at the receiver. A rough estimate is obtained by following the analysis of Section 2.4.2, where broadening of Gaussian pulses is discussed. Equation (2.4.16) shows that the optical pulse remains Gaussian, but its peak power is reduced by a pulse-broadening factor given by Eq. (2.4.24). If we define the power penalty $\delta_d$ as the increase (in dB) in the received power that would compensate the peak-power reduction, $\delta_d$ is given by

$$\delta_d = 10 \log_{10} f_b,$$

where $f_b$ is the pulse broadening factor. When pulse broadening is due mainly to a wide source spectrum at the transmitter, the broadening factor $f_b$ is given by Eq. (2.4.24), i.e.,

$$f_b = \sigma / \sigma_0 = [1 + (DL_\lambda / \sigma_0)^2]^{1/2},$$

where $\sigma_0$ is the RMS width of the optical pulse at the fiber input and $\sigma_\lambda$ is the RMS width of the source spectrum assumed to be Gaussian.

Equations (5.4.1) and (5.4.2) can be used to estimate the dispersion penalty for lightwave systems that use single-mode fiber together with a multimode laser or an LED. The ISI is minimized when the bit rate $B$ is such that $4B\sigma \leq 1$, as little pulse energy spreads beyond the bit slot ($T_R = 1/B$). By using $\sigma = (4B)^{-1}$, Eq. (5.4.2) can be written as

$$f_b^2 = 1 + (4BLD\sigma_\lambda f_b)^2.$$

By solving this equation for $f_b$ and substituting it in Eq. (5.4.1), the power penalty is given by

$$\delta_d = -5 \log_{10} [1 - (4BLD\sigma_\lambda)^2].$$

Figure 5.7 shows the power penalty as a function of the dimensionless parameter combination $BLD\sigma_\lambda$. Although the power penalty is negligible ($\delta_d = 0.38$ dB) for $BLD\sigma_\lambda = 0.1$, it increases to 2.2 dB when $BLD\sigma_\lambda = 0.2$ and becomes infinite when $BLD\sigma_\lambda = 0.25$. The $BL$ product, shown in Fig. 5.4, is truly limiting, since receiver
sensitivity degrades severely when a system is designed to approach it. Most lightwave systems are designed such that $\text{BLD}\sigma_\lambda < 0.2$, so that the dispersion penalty is below 2 dB. It should be stressed that Eq. (5.4.4) provides a rough estimate only as its derivation is based on several simplifying assumptions, such as a Gaussian pulse shape and a Gaussian source spectrum. These assumptions are not always satisfied in practice. Moreover, it is based on the condition $4B\sigma = 1$, so that the pulse remains nearly confined within the bit slot. It is possible to design a system such that the pulse spreads outside the bit slot but ISI is reduced through pulse shaping at the receiver.

### 5.4.3 Mode-Partition Noise

As discussed in Section 3.5.4, multimode semiconductor lasers exhibit *mode-partition noise* (MPN), a phenomenon occurring because of an anticorrelation among pairs of longitudinal modes. In particular, various longitudinal modes fluctuate in such a way that individual modes exhibit large intensity fluctuations even though the total intensity remains relatively constant. MPN would be harmless in the absence of fiber dispersion, as all modes would remain synchronized during transmission and detection. In practice, different modes become unsynchronized, since they travel at slightly different speeds inside the fiber because of group-velocity dispersion. As a result of such desynchronization, the receiver current exhibits additional fluctuations, and the SNR at the decision circuit becomes worse than that expected in the absence of MPN. A power penalty must be paid to improve the SNR to the same value that is necessary to achieve the required BER (see Section 4.5). The effect of MPN on system performance has been studied extensively for both multimode semiconductor lasers [74]–[83] and nearly single-mode lasers [84]–[98].
In the case of multimode semiconductor lasers, the power penalty can be calculated by following an approach similar to that of Section 4.6.2 and is given by [74]

$$\delta_{mpn} = -5 \log_{10}(1 - Q^2 r_{mpn}^2), \quad (5.4.5)$$

where $r_{mpn}$ is the relative noise level of the received power in the presence of MPN. A simple model for estimating the parameter $r_{mpn}$ assumes that laser modes fluctuate in such a way that the total power remains constant under CW operation [75]. It also assumes that the average mode power is distributed according to a Gaussian distribution of RMS width $\sigma_\lambda$ and that the pulse shape at the decision circuit of the receiver is described by a cosine function [74]. Different laser modes are assumed to have the same cross-correlation coefficient $\gamma_{cc}$, i.e.,

$$\gamma_{cc} = \frac{\langle P_i P_j \rangle}{\langle P_i \rangle \langle P_j \rangle}, \quad (5.4.6)$$

for all $i$ and $j$ such that $i \neq j$. The angular brackets denote an average over power fluctuations associated with mode partitioning. A straightforward calculation shows that $r_{mpn}$ is given by [78]

$$r_{mpn} = \left( \frac{k}{\sqrt{2}} \right) \left\{ 1 - \exp \left[ -\left( \frac{\pi BLD \sigma_\lambda}{2} \right)^2 \right] \right\}, \quad (5.4.7)$$

where the mode-partition coefficient $k$ is related to $\gamma_{cc}$ as $k = \sqrt{1 - \gamma_{cc}}$. The model assumes that mode partition can be quantified in terms of a single parameter $k$ with values in the range 0–1. The numerical value of $k$ is difficult to estimate and is likely to vary from laser to laser. Experimental measurements suggest that the values of $k$ are in the range 0.6–0.8 and vary for different mode pairs [75], [80].

Equations (5.4.5) and (5.4.7) can be used to calculate the MPN-induced power penalty. Figure 5.8 shows the power penalty at a BER of $10^{-9}$ ($Q = 6$) as a function of the normalized dispersion parameter $BLD \sigma_\lambda$ for several values of the mode-partition coefficient $k$. For a given value of $k$, the variation of power penalty is similar to that shown in Fig. 5.7; $\delta_{mpn}$ increases rapidly with an increase in $BLD \sigma_\lambda$ and becomes infinite when $BLD \sigma_\lambda$ reaches a critical value. For $k > 0.5$, the MPN-induced power penalty is larger than the penalty occurring due to dispersion-induced pulse broadening (see Fig. 5.7). However, it can be reduced to a negligible level ($\delta_{mpn} < 0.5 \text{ dB}$) by designing the optical communication system such that $BLD \sigma_\lambda < 0.1$. As an example, consider a 1.3-µm lightwave system. If we assume that the operating wavelength is matched to the zero-dispersion wavelength to within 10 nm, $D \approx 1 \text{ ps/(km-nm)}$. A typical value of $\sigma_\lambda$ for multimode semiconductor lasers is 2 nm. The MPN-induced power penalty would be negligible if the $BL$ product were below 50 (Gb/s)-km. At $B = 2 \text{ Gb/s}$ the transmission distance is then limited to 25 km. The situation becomes worse for 1.55-µm lightwave systems for which $D \approx 16 \text{ ps/(km-nm)}$ unless dispersion-shifted fibers are used. In general, the MPN-induced power penalty is quite sensitive to the spectral bandwidth of the multimode laser and can be reduced by reducing the bandwidth. In one study [83], a reduction in the carrier lifetime from 340 to 130 ps, realized by $p$-doping of the active layer, reduced the bandwidth of 1.3-µm semiconductor lasers by only 40% (from 5.6 to 3.4 nm), but the power penalty decreased from an infinite value (BER floor above $10^{-9}$ level) to a mere 0.5 dB.
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Figure 5.8: MPN-induced Power penalty versus $BLD\sigma_\lambda$ for a multimode semiconductor laser of RMS spectral width $\sigma_\lambda$. Different curves correspond to different values of the mode-partition coefficient $k$.

One may think that MPN can be avoided completely by using DFB lasers designed to oscillate in a single longitudinal mode. Unfortunately, this is not necessarily the case [88]–[91]. The reason is that the main mode of any DFB laser is accompanied by several side modes of much smaller amplitudes. The single-mode nature of DFB lasers is quantified through the mode-suppression ratio (MSR), defined as the ratio of the main-mode power $P_m$ to the power $P_s$ of the most dominant side mode. Clearly, the effect of MPN on system performance would depend on the MSR. Attempts have therefore been made to estimate the dependence of the MPN-induced power penalty on the MSR [84]–[98].

A major difference between the multimode and nearly single-mode semiconductor lasers is related to the statistics associated with mode-partition fluctuations. In a multimode laser, both main and side modes are above threshold and their fluctuations are well described by a Gaussian probability density function. By contrast, side modes in a DFB semiconductor laser are typically below threshold, and the optical power associated with them follows an exponential distribution given by [84]

$$p(P_s) = \bar{P_s}^{-1} \exp\left[-(P_s/\bar{P}_s)\right], \quad (5.4.8)$$

where $\bar{P}_s$ is the average value of the random variable $P_s$.

The effect of side-mode fluctuations on system performance can be appreciated by considering an ideal receiver. Let us assume that the relative delay $\Delta T = DL\Delta\lambda$ between the main and side modes is large enough that the side mode appears outside the bit slot (i.e., $\Delta T > 1/B$ or $BLD\Delta\lambda L > 1$, where $\Delta\lambda_L$ is the mode spacing). The decision circuit of the receiver would make an error for 0 bits if the side-mode power $P_s$ were to exceed the decision threshold set at $\bar{P}_m/2$, where $\bar{P}_m$ is the average main-mode
power. Furthermore, the two modes are anticorrelated in such a way that the main-mode power drops below \( P_m / 2 \) whenever side-mode power exceeds \( P_m / 2 \), so that the total power remains nearly constant [85]. Thus, an error would occur even for “1” bits whenever \( P_s > P_m / 2 \). Since the two terms in Eq. (4.5.2) make equal contributions, the BER is given by [84]

\[
\text{BER} = \int_{P_m/2}^{\infty} p(P_s) dP_s = \exp \left( -\frac{P_m}{2P_s} \right) = \exp \left( -\frac{R_{ms}}{2} \right) .
\]

(5.4.9)

The BER depends on the MSR defined as \( R_{ms} = P_m / P_s \) and exceeds 10\(^{-9}\) when MSR < 42.

To calculate the MPN-induced power penalty in the presence of receiver noise, one should follow the analysis in Section 4.5.1 and add an additional noise term that accounts for side-mode fluctuations. For a \( p-i-n \) receiver the BER is found to be [85]

\[
\text{BER} = \frac{1}{2} \text{erfc} \left( \frac{Q}{\sqrt{2}} \right) + \exp \left( -\frac{R_{ms}}{2} + \frac{R_{ms}^2}{4Q^2} \right) \left[ 1 - \frac{1}{2} \text{erfc} \left( \frac{Q}{\sqrt{2}} - \frac{R_{ms}}{Q\sqrt{2}} \right) \right] ,
\]

(5.4.10)

where the parameter \( Q \) is defined by Eq. (4.5.10). In the limit of an infinite MSR, Eq. (5.4.10) reduces to Eq. (4.5.9). For a noise-free receiver \( (Q = \infty) \), Eq. (5.4.10) reduces to Eq. (5.4.9). Figure 5.9 shows the BER versus the power penalty at a BER of 10\(^{-9}\) as a function of MSR. As expected, the power penalty becomes infinite for MSR values below 42, since the 10\(^{-9}\) BER cannot be realized irrespective of the power received. The penalty can be reduced to a negligible level (<0.1 dB) for MSR values in excess of 100 (20 dB).

The experimental measurements of the BER in several transmission experiments show that a BER floor above the 10\(^{-9}\) level can occur even for DFB lasers which exhibit a MSR in excess of 30 dB under continuous-wave (CW) operation [88]–[91]. The reason behind the failure of apparently good lasers is related to the possibility of side-mode excitation under transient conditions occurring when the laser is repeatedly turned on and off to generate the bit stream. When the laser is biased below threshold and modulated at a high bit rate \( (B \geq 1 \text{ Gb/s}) \), the probability of side-mode excitation above \( P_m / 2 \) is much higher than that predicted by Eq. (5.4.8). Considerable attention has been paid to calculate, both analytically and numerically, the probability of the transient excitation of side modes and its dependence on various device parameters [87]–[98]. An important device parameter is found to be the gain margin between the main and side modes. The gain margin should exceed a critical value which depends on the bit rate. The critical value is about 5–6 cm\(^{-1}\) at 500 Mb/s [88] but can exceed 15 cm\(^{-1}\) at high bit rates, depending on the bias and modulation currents [93]. The bias current plays a critical role. Numerical simulations show that the best performance is achieved when the DFB laser is biased close to but slightly below threshold to avoid the bit-pattern effects [98]. Moreover, the effects of MPN are independent of the bit rate as long as the gain margin exceeds a certain value. The required value of gain margin exceeds 25 cm\(^{-1}\) for the 5-GHz modulation frequency. Phase-shifted DFB lasers have a large built-in gain margin and have been developed for this purpose.
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Figure 5.9: Effect of MPN on bit-error rate of DFB lasers for several values of MSR. Intersection of the dashed line with the solid curves provides MPN-induced power penalty. (After Ref. [85]; ©1985 IEEE; reprinted with permission.)

5.4.4 Frequency Chirping

Frequency chirping is an important phenomenon that is known to limit the performance of 1.55-μm lightwave systems even when a DFB laser with a large MSR is used to generate the digital bit stream [99]–[112]. As discussed in Section 3.5.3, intensity modulation in semiconductor lasers is invariably accompanied by phase modulation because of the carrier-induced change in the refractive index governed by the linewidth enhancement factor. Optical pulses with a time-dependent phase shift are called chirped. As a result of the frequency chirp imposed on an optical pulse, its spectrum is considerably broadened. Such spectral broadening affects the pulse shape at the fiber output because of fiber dispersion and degrades system performance.

An exact calculation of the chirp-induced power penalty $\delta_c$ is difficult because frequency chirp depends on both the shape and the width of the optical pulse [101]–[104]. For nearly rectangular pulses, experimental measurements of time-resolved pulse spectra show that frequency chirp occurs mainly near the leading and trailing edges such that the leading edge shifts toward the blue while the trailing edge shifts toward the red. Because of the spectral shift, the power contained in the chirped portion of the pulse moves out of the bit slot when the pulse propagates inside the optical fiber. Such
a power loss decreases the SNR at the receiver and results in power penalty. In a simple model the chirp-induced power penalty is given by \[ \delta_c = -10 \log_{10}(1 - 4B\Delta\lambda c), \] (5.4.11)

where \( \Delta\lambda_c \) is the spectral shift associated with frequency chirping. This equation applies as long as \( LDA\Delta\lambda_c < t_c \), where \( t_c \) is the chirp duration. Typically, \( t_c \) is 100–200 ps, depending on the relaxation-oscillation frequency, since chirping lasts for about one-half of the relaxation-oscillation period. By the time \( LDA\Delta\lambda_c \) equals \( t_c \), the power penalty stops increasing because all the chirped power has left the bit interval. For \( LDA\Delta\lambda_c > t_c \), the product \( LDA\Delta\lambda_c \) in Eq. (5.4.11) should be replaced by \( t_c \).

The model above is overly simplistic, as it does not take into account pulse shaping at the receiver. A more accurate calculation based on raised-cosine filtering (see Section 4.3.2) leads to the following expression \[ \delta = -20 \log_{10}\left\{1 - (4\pi^2/3 - 8)B^2LDA\Delta\lambda_c t_c [1 + (2B/3)(LDA\Delta\lambda_c - t_c)]\right\}. \] (5.4.12)

The receiver is assumed to contain a \( p-i-n \) photodiode. The penalty is larger for an APD, depending on the excess-noise factor of the APD. Figure 5.10 shows the power penalty \( \delta_c \) as a function of the parameter combination \( BLD\Delta\lambda_c \) for several values of the parameter \( Bt_c \), which is a measure of the fraction of the bit period over which chirping occurs. As expected, \( \delta_c \) increases with both the chirp \( \Delta\lambda_c \) and the chirp duration \( t_c \). The power penalty can be kept below 1 dB if the system is designed such that \( BLD\Delta\lambda_c < 0.1 \) and \( Bt_c < 0.2 \). A shortcoming of this model is that \( \Delta\lambda_c \) and \( t_c \) appear as free
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Figure 5.11: Chirp-induced power penalty as a function of $|\beta_2|B^2L$ for several values of the chirp parameter $C$. The Gaussian optical pulse is assumed to be linearly chirped over its entire width.

Parameters and must be determined for each laser through experimental measurements of the frequency chirp. In practice, $\Delta\lambda_c$ itself depends on the bit rate $B$ and increases with it.

For lightwave systems operating at high bit rates ($B > 2$ Gb/s), the bit duration is generally shorter than the total duration $2\tau_c$ over which chirping is assumed to occur in the foregoing model. The frequency chirp in that case increases almost linearly over the entire pulse width (or bit slot). A similar situation occurs even at low bit rates if the optical pulses do not contain sharp leading and trailing edges but have long rise and fall times (Gaussian-like shape rather than a rectangular shape). If we assume a Gaussian pulse shape and a linear chirp, the analysis of Section 2.4.2 can be used to estimate the chirp-induced power penalty. Equation (2.4.16) shows that the chirped Gaussian pulse remains Gaussian but its peak power decreases because of dispersion-induced pulse broadening. Defining the power penalty as the increase (in dB) in the received power that would compensate the peak-power reduction, $\delta_c$ is given by

$$\delta_c = 10\log_{10} f_b,$$

where $f_b$ is the broadening factor given by Eq. (2.4.22) with $\beta_3 = 0$. The RMS width $\sigma_0$ of the input pulse should be such that $4\sigma_0 \leq 1/B$. Choosing the worst-case condition $\sigma_0 = 1/4B$, the power penalty is given by

$$\delta_c = 5\log_{10}[(1 + 8C|\beta_2|B^2L)^2 + (8|\beta_2|B^2L)^2].$$

Figure 5.11 shows the chirp-induced power penalty as a function of $|\beta_2|B^2L$ for several values of the chirp parameter $C$. The parameter $\beta_2$ is taken to be negative,
as is the case for 1.55-µm lightwave systems. The $C = 0$ curve corresponds to the case of a chirp-free pulse. The power penalty is negligible ($<0.1$ dB) in this ideal case as long as $|\beta_2|B^2L < 0.05$. However, the penalty can exceed 5 dB if the pulses transmitted are chirped such that $C = -6$. To keep the penalty below 0.1 dB, the system should be designed with $|\beta_2|B^2L < 0.002$. For $|\beta_2| = 20$ ps$^2$/km, $B^2L$ is limited to 100 (Gb/s)$^2$-km. Interestingly, system performance is improved for positive values of $C$ since the optical pulse then goes through an initial compression phase (see Section 2.4). Unfortunately, $C$ is negative for semiconductor lasers; it can be approximated by $-\beta_c$, where $\beta_c$ is the linewidth enhancement factor with positive values of 2–6.

It is important to stress that the analytic results shown in Figs. 5.10 and 5.11 provide only a rough estimate of the power penalty. In practice, the chirp-induced power penalty depends on many system parameters. For instance, several system experiments have shown that the effect of chirp can be reduced by biasing the semiconductor laser above threshold [103]. However, above-threshold biasing increases that extinction ratio $r_{\text{ex}}$, defined in Eq. (4.6.1) as $r_{\text{ex}} = P_0/P_1$, where $P_0$ and $P_1$ are the powers received for bit 0 and bit 1, respectively. As discussed in Section 4.6.1, an increase in $r_{\text{ex}}$ decreases the receiver sensitivity and leads to its own power penalty. Clearly, $r_{\text{ex}}$ cannot be increased indefinitely in an attempt to reduce the chirp penalty. The total system performance can be optimized by designing the system so that it operates with an optimum value of $r_{\text{ex}}$ that takes into account the trade-off between the chirp and the extinction ratio. Numerical simulations are often used to understand such trade-offs in actual lightwave systems [110]–[113]. Figure 5.12 shows the power penalty as a function of the extinction ratio $r_{\text{ex}}$ by simulating numerically the performance of a 1.55-µm lightwave system transmitting at 4 Gb/s over a 100-km-long fiber. The total penalty can be reduced
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below 2 dB by operating the system with an extinction ratio of about 0.1. The optimum values of $e_{ex}$ and the total penalty are sensitive to many other laser parameters such as the active-region width. A semiconductor laser with a wider active region is found to have a larger chirp penalty [105]. The physical phenomenon behind this width dependence appears to be the nonlinear gain [see Eq. (3.3.40)] and the associated damping of relaxation oscillations. In general, rapid damping of relaxation oscillations decreases the effect of frequency chirp and improves system performance [113].

The origin of chirp in semiconductor lasers is related to carrier-induced index changes governed by the linewidth enhancement factor $\beta_c$. The frequency chirp would be absent for a laser with $\beta_c = 0$. Unfortunately, $\beta_c$ cannot be made zero for semiconductor lasers, although it can be reduced by adopting a multi-quantum-well (MQW) design [114]–[118]. The use of a MQW active region reduces $\beta_c$ by about a factor of 2. In one 1.55-μm experiment [120], the 10-Gb/s signal could be transmitted over 60–70 km, despite the high dispersion of standard telecommunication fiber, by biasing the laser above threshold. The MQW DFB laser used in the experiment had $\beta_c \approx 3$. A further reduction in $\beta_c$ occurs for strained quantum wells [118]. Indeed, $\beta_c \approx 1$ has been measured in modulation-doped strained MQW lasers [119]. Such lasers exhibit low chirp under direct modulation at bit rates as high as 10 Gb/s.

An alternative scheme eliminates the laser-chirp problem completely by operating the laser continuously and using an external modulator to generate the bit stream. This approach has become practical with the development of optical transmitters in which a modulator is integrated monolithically with a DFB laser (see Section 3.6.4). The chirp parameter $C$ is close to zero in such transmitters. As shown by the $C = 0$ curve in Fig. 5.11, the dispersion penalty is below 2 dB in that case even when $|\beta_2|B^2L$ is close to 0.2. Moreover, an external modulator can be used to modulate the phase of the optical carrier in such a way that $\beta_2C < 0$ in Eq. (5.4.14). As seen in Fig. 5.11, the chirp-induced power penalty becomes negative over a certain range of $|\beta_2|B^2L$, implying that such frequency chirping is beneficial to combat the effects of dispersion. In a 1996 experiment [121], the 10-Gb/s signal was transmitted penalty free over 100 km of standard telecommunication fiber by using a modulator-integrated transmitter such that $C$ was effectively positive. By using $\beta_2 \approx -20$ ps²/km, it is easy to verify that $|\beta_2|B^2L = 0.2$ for this experiment, a value that would have produced a power penalty of more than 8 dB if the DFB laser were modulated directly.

5.4.5 Reflection Feedback and Noise

In most fiber-optic communication systems, some light is invariably reflected back because of refractive-index discontinuities occurring at splices, connectors, and fiber ends. The effects of such unintentional feedback have been studied extensively [122]–[140] because they can degrade the performance of lightwave systems considerably. Even a relatively small amount of optical feedback affects the operation of semiconductor lasers [126] and can lead to excess noise in the transmitter output. Even when an isolator is used between the transmitter and the fiber, multiple reflections between splices and connectors can generate additional intensity noise and degrade receiver performance [128]. This subsection is devoted to the effect of reflection-induced noise on receiver sensitivity.
Most reflections in a fiber link originate at glass–air interfaces whose reflectivity can be estimated by using 

\[ R_f = \frac{(n_f - 1)^2}{(n_f + 1)^2}, \]

where \( n_f \) is the refractive index of the fiber material. For silica fibers \( R_f = 3.6\% \) (\(-14.4\) dB) if we use \( n_f = 1.47 \). This value increases to 5.3\% for polished fiber ends since polishing can create a thin surface layer with a refractive index of about 1.6. In the case of multiple reflections occurring between two splices or connectors, the reflection feedback can increase considerably because the two reflecting surfaces act as mirrors of a Fabry–Perot interferometer. When the resonance condition is satisfied, the reflectivity increases to 14\% for unpolished surfaces and to over 22\% for polished surfaces. Clearly, a considerable fraction of the signal transmitted can be reflected back unless precautions are taken to reduce the optical feedback. A common technique for reducing reflection feedback is to use index-matching oil or gel near glass–air interfaces. Sometimes the tip of the fiber is curved or cut at an angle so that the reflected light deviates from the fiber axis. Reflection feedback can be reduced to below 0.1\% by such techniques.

Semiconductor lasers are extremely sensitive to optical feedback [133]; their operating characteristics can be affected by feedback as small as \(-80\) dB [126]. The most dramatic effect of feedback is on the laser linewidth, which can narrow or broaden by several orders of magnitude, depending on the exact location of the surface where feedback originates [122]. The reason behind such a sensitivity is related to the fact that the phase of the reflected light can perturb the laser phase significantly even for relatively weak feedback levels. Such feedback-induced phase changes are detrimental mainly for coherent communication systems. The performance of direct-detection lightwave systems is affected by intensity noise rather than phase noise.

Optical feedback can increase the intensity noise significantly. Several experiments have shown a feedback-induced enhancement of the intensity noise occurring at frequencies corresponding to multiples of the external-cavity mode spacing [123]–[125]. In fact, there are several mechanisms through which the relative intensity noise (RIN) of a semiconductor laser can be enhanced by the external optical feedback. In a simple model [127], the feedback-induced enhancement of the intensity noise is attributed to the onset of multiple, closely spaced, external-cavity longitudinal modes whose spacing is determined by the distance between the laser output facet and the glass–air interface where feedback originates. The number and the amplitudes of the external-cavity modes depend on the amount of feedback. In this model, the RIN enhancement is due to intensity fluctuations of the feedback-generated side modes. Another source of RIN enhancement has its origin in the feedback-induced chaos in semiconductor lasers. Numerical simulations of the rate equations show that the RIN can be enhanced by 20 dB or more when the feedback level exceeds a certain value [134]. Even though the feedback-induced chaos is deterministic in nature, it manifests as an apparent RIN increase.

Experimental measurements of the RIN and the BER in the presence of optical feedback confirm that the feedback-induced RIN enhancement leads to a power penalty in lightwave systems [137]–[140]. Figure 5.13 shows the results of the BER measurements for a VCSEL operating at 958 nm. Such a laser operates in a single longitudinal mode because of an ultrashort cavity length (\( \sim 1\) \(\mu\)m) and exhibits a RIN near \(-130\) dB/Hz in the absence of reflection feedback. However, the RIN increases by as much as 20 dB when the feedback exceeds the \(-30\)-dB level. The BER measurements
at a bit rate of 500 Mb/s show a power penalty of 0.8 dB at a BER of $10^{-9}$ for $-30$-dB feedback, and the penalty increases rapidly at higher feedback levels [139].

The power penalty can be calculated by following the analysis of Section 4.6.2 and is given by

$$\delta_{ref} = -10 \log_{10}(1 - r_{eff}^2 Q^2),$$  \hspace{1cm} (5.4.15)

where $r_{eff}$ is the effective intensity noise over the receiver bandwidth $\Delta f$ and is obtained from

$$r_{eff}^2 = \frac{1}{2\pi} \int_{-\infty}^{\infty} RIN(\omega) d\omega = 2(RIN)\Delta f.$$  \hspace{1cm} (5.4.16)

In the case of feedback-induced external-cavity modes, $r_{eff}$ can be calculated by using a simple model and is found to be [127]

$$r_{eff}^2 \approx r_I^2 + N/(\text{MSR})^2,$$  \hspace{1cm} (5.4.17)

where $r_I$ is the relative noise level in the absence of reflection feedback, $N$ is the number of external-cavity modes, and MSR is the factor by which the external-cavity modes remain suppressed. Figure 5.14 shows the reflection-noise power penalty as a function of MSR for several values of $N$ by choosing $r_I = 0.01$. The penalty is negligible in the absence of feedback ($N = 0$). However, it increases with an increase in $N$ and a decrease in MSR. In fact, the penalty becomes infinite when MSR is reduced below a critical
value. Thus, reflection feedback can degrade system performance to the extent that the system cannot achieve the desired BER despite an indefinite increase in the power received. Such reflection-induced BER floors have been observed experimentally [125] and indicate the severe impact of reflection noise on the performance of lightwave systems. An example of the reflection-induced BER floor is seen in Fig. 5.13, where the BER remains above $10^{-9}$ for feedback levels in excess of $-25$ dB. Generally speaking, most lightwave systems operate satisfactorily when the reflection feedback is below $-30$ dB. In practice, the problem can be nearly eliminated by using an optical isolator within the transmitter module.

Even when an isolator is used, reflection noise can be a problem for lightwave systems. In long-haul fiber links making use of optical amplifiers, fiber dispersion can convert the phase noise to intensity noise, leading to performance degradation [130]. Similarly, two reflecting surfaces anywhere along the fiber link act as a Fabry–Perot interferometer which can convert phase noise into intensity noise [128]. Such a conversion can be understood by noting that multiple reflections inside a Fabry–Perot interferometer lead to a phase-dependent term in the transmitted intensity which fluctuates in response to phase fluctuations. As a result, the RIN of the signal incident on the receiver is higher than that occurring in the absence of reflection feedback. Most of the RIN enhancement occurs over a narrow frequency band whose spectral width is governed by the laser linewidth ($\sim 100$ MHz). Since the total noise is obtained by integrating over the receiver bandwidth, it can affect system performance considerably at bit rates larger than the laser linewidth. The power penalty can still be calculated by using Eq. (5.4.15). A simple model that includes only two reflections between the reflecting interfaces shows that $r_{\text{eff}}$ is proportional to $(R_1 R_2)^{1/2}$, where $R_1$ and $R_2$ are the
reflectivities of the two interfaces [128]. Figure 4.19 can be used to estimate the power penalty. It shows that power penalty can become infinite and lead to BER floors when $r_{eff}$ exceeds 0.2. Such BER floors have been observed experimentally [128]. They can be avoided only by eliminating or reducing parasitic reflections along the entire fiber link. It is therefore necessary to employ connectors and splices that reduce reflections through the use of index matching or other techniques.

## 5.5 Computer-Aided Design

The design of a fiber-optic communication system involves optimization of a large number of parameters associated with transmitters, optical fibers, in-line amplifiers, and receivers. The design aspects discussed in Section 5.2 are too simple to provide the optimized values for all system parameters. The power and the rise-time budgets are only useful for obtaining a conservative estimate of the transmission distance (or repeater spacing) and the bit rate. The system margin in Eq. (5.2.4) is used as a vehicle to include various sources of power penalties discussed in Section 5.4. Such a simple approach fails for modern high-capacity systems designed to operate over long distances using optical amplifiers.

An alternative approach uses computer simulations and provides a much more realistic modeling of fiber-optic communication systems [141]–[156]. The computer-aided design techniques are capable of optimizing the whole system and can provide the optimum values of various system parameters such that the design objectives are met at a minimum cost. Figure 5.15 illustrates the various steps involved in the simulation process. The approach consists of generating an optical bit pattern at the transmitter, transmitting it through the fiber link, detecting it at the receiver, and then analyzing it through the tools such as the eye diagram and the $Q$ factor.

![Figure 5.15: Steps involved in computer modeling of fiber-optic communication systems.](image)
Each step in the block diagram shown in Fig. 5.15 can be carried out numerically by using the material given in Chapters 2–4. The input to the optical transmitter is a pseudorandom sequence of electrical pulses which represent 1 and 0 bits. The length $N$ of the pseudorandom bit sequence determines the computing time and should be chosen judiciously. Typically, $N = 2^M$, where $M$ is in the range 6–10. The optical bit stream can be obtained by solving the rate equations that govern the modulation response of semiconductor lasers (see Section 3.5). The equations governing the modulation response should be used if an external modulator is used. Chirping is automatically included in both cases. Deformation of the optical bit stream during its transmission through the optical fiber is calculated by solving the NLS equation (5.3.1). The noise added by optical amplifiers should be included at the location of each amplifier.

The optical signal is converted into the electrical domain at the receiver. The shot and thermal noise is adding through a fluctuating term with Gaussian statistics. The electrical bit stream is shaped by passing it through a filter whose bandwidth is also a design parameter. An eye diagram is constructed using the filtered bit stream. The effect of varying system parameters can be studied by monitoring the eye degradation or by calculating the $Q$ parameter given in Eq. (4.5.11). Such an approach can be used to obtain the power penalty associated with various mechanisms discussed in Section 5.4. It can also be used to investigate trade-offs that would optimize the overall system performance. An example is shown in Fig. 5.12, where the dependence of the calculated system penalty on the frequency chirp and extinction ratio is found. Numerical simulations reveal the existence of an optimum extinction ratio for which the system penalty is minimum.

Computer-aided design has another important role to play. A long-haul lightwave system may contain many repeaters, both optical and electrical. Transmitters, receivers, and amplifiers used at repeaters, although chosen to satisfy nominal specifications, are never identical. Similarly, fiber cables are constructed by splicing many different pieces (typical length 4–8 km) which have slightly different loss and dispersion characteristics. The net result is that many system parameters vary around their nominal values. For example, the dispersion parameter $D$, responsible not only for pulse broadening but also for other sources of power penalty, can vary significantly in different sections of the fiber link because of variations in the zero-dispersion wavelength and the transmitter wavelength. A statistical approach is often used to estimate the effect of such inherent variations on the performance of a realistic lightwave system [146]–[150]. The idea behind such an approach is that it is extremely unlikely that all system parameters would take their worst-case values at the same time. Thus, repeater spacing can be increased well above its worst-case value if the system is designed to operate reliably at the specific bit rate with a high probability (say 99.9%).

The importance of computer-aided design for fiber-optic communication systems became apparent during the 1990s when the dispersive and nonlinear effects in optical fibers became of paramount concern with increasing bit rates and transmission distances. All modern lightwave systems are designed using numerical simulations, and several software packages are available commercially. Appendix E provides details on the simulation package available on the CD-ROM included with this book (Courtesy OptiWave Corporation). The reader is encouraged to use it for a better understanding of the material covered in this book.
Problems

5.1 A distribution network uses an optical bus to distribute the signal to 10 users. Each optical tap couples 10% of the power to the user and has 1-dB insertion loss. Assuming that the station 1 transmits 1 mW of power over the optical bus, calculate the power received by the stations 8, 9, and 10.

5.2 A cable-television operator uses an optical bus to distribute the video signal to its subscribers. Each receiver needs a minimum of 100 nW to operate satisfactorily. Optical taps couple 5% of the power to each subscriber. Assuming 0.5 dB insertion loss for each tap and 1 mW transmitter power, estimate the number of subscribers that can be added to the optical bus?

5.3 A star network uses directional couplers with 0.5-dB insertion loss to distribute data to its subscribers. If each receiver requires a minimum of 100 nW and each transmitter is capable of emitting 0.5 mW, calculate the maximum number of subscribers served by the network.

5.4 Make the power budget and calculate the maximum transmission distance for a 1.3-µm lightwave system operating at 100 Mb/s and using an LED for launching 0.1 mW of average power into the fiber. Assume 1-dB/km fiber loss, 0.2-dB splice loss every 2 km, 1-dB connector loss at each end of fiber link, and 100-nW receiver sensitivity. Allow 6-dB system margin.

5.5 A 1.3-µm long-haul lightwave system is designed to operate at 1.5 Gb/s. It is capable of coupling 1 mW of average power into the fiber. The 0.5-dB/km fiber-cable loss includes splice losses. The connectors at each end have 1-dB losses. The InGaAs p–i–n receiver has a sensitivity of 250 nW. Make the power budget and estimate the repeater spacing.

5.6 Prove that the rise time $T_r$ and the 3-dB bandwidth $\Delta f$ of a RC circuit are related by $T_r \Delta f = 0.35$.

5.7 Consider a super-Gaussian optical pulse with the power distribution

$$P(t) = P_0 \exp\left[-(t/T_0)^{2m}\right],$$

where the parameter $m$ controls the pulse shape. Derive an expression for the rise time $T_r$ of such a pulse. Calculate the ratio $T_r/T_{\text{FWHM}}$, where $T_{\text{FWHM}}$ is the full width at half maximum, and show that for a Gaussian pulse ($m = 1$) this ratio equals 0.716.

5.8 Prove that for a Gaussian optical pulse, the rise time $T_r$ and the 3-dB optical bandwidth $\Delta f$ are related by $T_r \Delta f = 0.316$.

5.9 Make the rise-time budget for a 0.85-µm, 10-km fiber link designed to operate at 50 Mb/s. The LED transmitter and the Si p–i–n receiver have rise times of 10 and 15 ns, respectively. The graded-index fiber has a core index of 1.46, $\Delta = 0.01$, and $D = 80$ ps/(km-nm). The LED spectral width is 50 nm. Can the system be designed to operate with the NRZ format?
5.10 A 1.3-µm lightwave system is designed to operate at 1.7 Gb/s with a repeater spacing of 45 km. The single-mode fiber has a dispersion slope of 0.1 ps/(km-nm²) in the vicinity of the zero-dispersion wavelength occurring at 1.308 µm. Calculate the wavelength range of multimode semiconductor lasers for which the mode-partition-noise power penalty remains below 1 dB. Assume that the RMS spectral width of the laser is 2 nm and the mode-partition coefficient $k = 0.7$.

5.11 Generalize Eq. (5.4.5) for the case of APD receivers by including the excess-noise factor in the form $F(M) = M^4$.

5.12 Consider a 1.55-µm lightwave system operating at 1 Gb/s by using multimode semiconductor lasers of 2 nm (RMS) spectral width. Calculate the maximum transmission distance that would keep the mode-partition-noise power penalty below 2 dB. Use $k = 0.8$ for the mode-partition coefficient.

5.13 Follow the rate-equation analysis of Section 3.3.8 (see also Ref. [84]) to prove that the side-mode power $P_s$ follows an exponential probability density function given by Eq. (5.4.8).

5.14 Use Eq. (5.4.14) to determine the maximum transmission distance for a 1.55-µm lightwave system operating at 4 Gb/s such that the chirp-induced power penalty is below 1 dB. Assume that $C = -6$ for the single-mode semiconductor laser and $\beta_2 = -20$ ps²/km for the single-mode fiber.

5.15 Repeat Problem 5.14 for the case of 8-Gb/s bit rate.

5.16 Use the results of Problem 4.16 to obtain an expression of the reflection-induced power penalty in the case of a finite extinction ratio $r_{ex}$. Reproduce the penalty curves shown in Fig. 5.13 for the case $r_{ex} = 0.1$.

5.17 Consider a Fabry–Perot interferometer with two surfaces of reflectivity $R_1$ and $R_2$. Follow the analysis of Ref. [128] to derive an expression of the relative intensity noise $\text{RIN}(\omega)$ of the transmitted light as a function of the linewidth of the incident light. Assume that $R_1$ and $R_2$ are small enough that it is enough to consider only a single reflection at each surface.

5.18 Follow the analysis of Ref. [142] to obtain an expression for the total receiver noise by including thermal noise, shot noise, intensity noise, mode-partition noise, chirp noise, and reflection noise.

References


REFERENCES


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