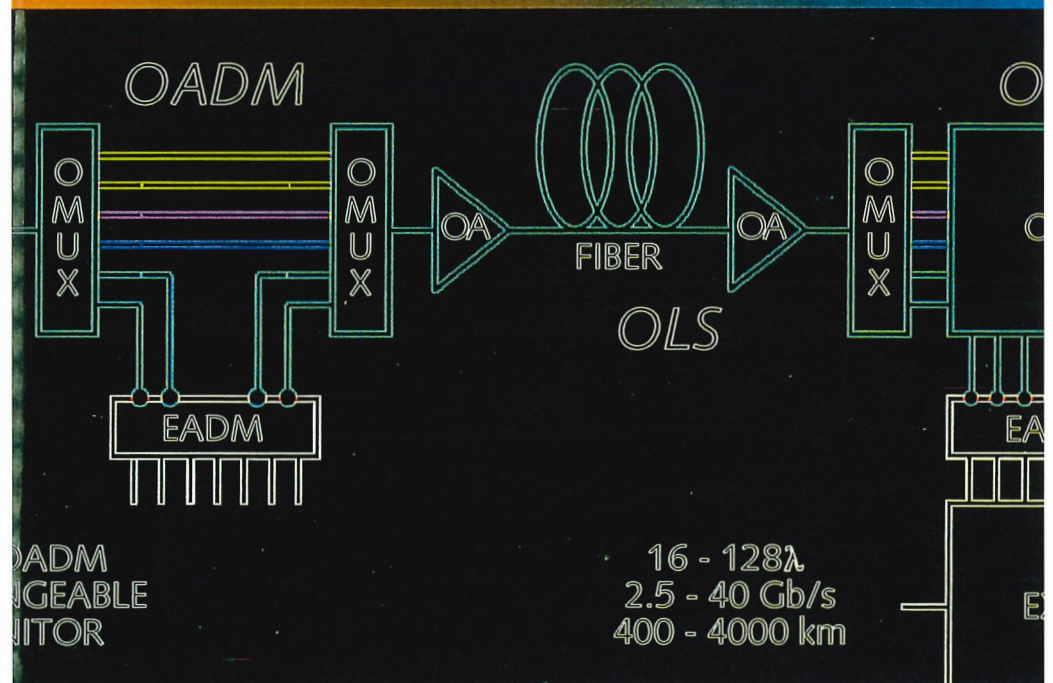


OPTICAL FIBER TELECOMMUNICATIONS SYSTEMS AND IMPAIRMENTS

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OPTICAL FIBER TELECOMMUNICATIONS IV B SYSTEMS AND IMPAIRMENTS

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1. Introduction

As the time-division multiplexed (TDM) bit rates of fiber-optic systems have increased from the early systems operating at DS-3 [1] to modern commercial rates of 9.95328 Gb/s (OC-192/SDH-64) over single-mode fibers, attention has turned to the deleterious effects of chromatic dispersion in nondispersion shifted single-mode fiber (NDSF). For instance, a system operating at 2.48832 Gb/s (OC-48/SDH-16) using a wavelength in the region of minimum loss (~ 1550 nm) sees a chromatic dispersion coefficient close to 17 ps/km-nm, which results in a 1 dB eye-closure penalty at a distance around 700 km when the signal is transmitted chirp free, such as can be accomplished with a LiNbO_3 modulator driven in push-pull configuration. Because the chromatic-dispersion-limited distance varies approximately inversely as the bit rate squared [2], this distance is reduced by a factor of about 16 at OC-192 compared with OC-48. With a bit of prechirping at the transmitter [3], this distance can be improved to ~ 100 km at 10 Gb/s, a distance that is quite comparable to practical loss-limited distances at this bit rate. However, with the advent of the erbium-doped fiber amplifier (EDFA), dramatic increases in loss-limited transmission distances became possible, with the result that most terrestrial long-haul systems have a degree of dispersion-limited performance, which depends on the type of modern fiber used and the nature of the dispersion compensation technique used to reverse the chromatic dispersion of the fiber [see 4,5].

As the push for higher capacity and longer reach continues, TDM rates and the number of wavelength-division multiplexed (WDM) channels continue to increase, causing WDM channel spacings to decrease, thereby placing added emphasis on the impact of dispersion as well as intra- and interchannel nonlinearities. One technology that could improve system performance is the use of transmission formats that provide better immunity to dispersion and/or nonlinear fiber impairments as well as allowing higher channel density or spectral efficiency. The interplay of these items is often somewhat mutually exclusive in that nonlinear impairments as well as spectral filtering requirements in optical demultiplexers get more difficult to handle as channel separation decreases.

If transmission formats can be conceived (or adapted from those that have been used in radio or other wireline communications systems) and implemented that produce a transmission spectrum that is narrower than the commonly used non-return-to-zero (NRZ) format, some dispersion benefit should consequently accrue, either by way of not requiring any dispersion compensation at all or requiring less compensation.

However, there is another, and perhaps more compelling, reason for exploring alternative transmission formats that have reduced spectral occupancy. This is illustrated in Table 16.1, where we indicate the decreasing channel separations and increasing TDM bit rates at which dense WDM (DWDM) systems have evolved, and through which the total system capacities and spectral efficiencies have been dramatically increased. To a great extent the economics of higher and higher bit rate systems is based on increasing total system capacity within the spectrum available from an EDFA, simply by filling up the EDFA spectrum with signal spectrum. The quantitative measure of this is the system spectral efficiency, which is defined as the ratio of the individual channel bit rate to the DWDM channel separation, and it is this quantity that is becoming the fundamental limiting factor in increasing system capacity.

To illustrate this with NRZ transmission in which the individual bits are modeled ideally as rectangular pulses occupying the full time slot of duration $T = 1/B$ where B is the bit rate, the baseband power spectral density (for square NRZ pulses of amplitude A) has a sinc function relationship given by [6]

$$S_a(f) = \frac{A^2 T}{4} \text{sinc}^2(fT) + \frac{A^2}{4} \delta(f) \quad (16.1)$$

and is illustrated in Fig. 16.1a.

Thus the baseband NRZ signal spectrum can be represented in the frequency domain by a power spectral density that has nulls at $\pm NB$, where N is an integer, which upon amplitude modulation of an optical carrier at frequency f_c is frequency translated to be situated on either side of the optical carrier frequency, as illustrated in Fig. 16.1b.

Thus the transmission bandwidth, if measured between the first nulls on either side of the carrier, is twice the bit rate, and is centered on the optical carrier frequency. Normally, the baseband spectrum above $\pm B$ is filtered

Table 16.1 System Capacity Evolution

<i>Bit Rate (Gb/s)</i>	<i>Channel Spacing (GHz)</i>	<i>Spectral Efficiency (%)</i>
2.5	100/50/25	2.5/5/10
10	200/100/50	5/10/20
40	200/100	20/40

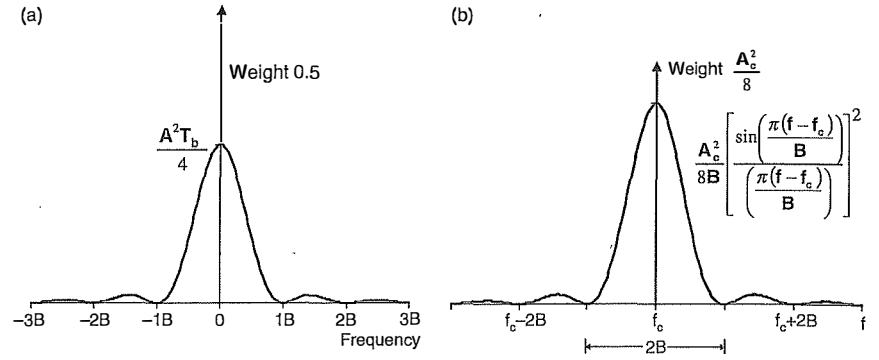


Fig. 16.1 (a) Baseband power spectral density for NRZ and (b) power spectral density of an NRZ signal amplitude modulated onto a carrier at f_c (only positive frequencies shown).

out, which results in practical pulses with nonzero rise and fall times, which also substantially reduces the frequency content of the signal spectrum above the first nulls in the spectrum at the bit rate B . This also reduces spectral overlap between adjacent optical channels in a DWDM configuration, which can be placed at frequency intervals of $2B$ without incurring interchannel interference or cross-talk, and thus the maximum spectral efficiency for NRZ is 50% with this simple approximation. In a practical sense a spectral efficiency of 40% is realistic.

Thus for 10 and 40 Gb/s, the channel separations at 40% spectral efficiency are 25 and 100 GHz, respectively, which falls conveniently on the ITU spectral grid. It is worth noting that increasing the bit rate beyond 40 Gb/s cannot increase system capacity in the sense shown by the evolution in spectral efficiency illustrated in Table 16.1. What does result is a coarser channel plan that could have benefit in DWDM networks since a smaller number of wavelengths need to be managed.

Additionally, because interchannel nonlinear effects such as four-wave mixing (FWM) and cross-phase modulation (XPM) vary inversely with channel separation, reduced interchannel FWM and XPM penalties do occur as the bit rate is increased while keeping spectral efficiency fixed. Going to closer channel spacings at lower bit rates, while keeping spectral efficiencies in the 40% range, presents difficult challenges from both an optical filtering perspective as well as from the standpoint of fiber nonlinearities and will require further research.

Thus, what is required if greater spectral efficiency, and hence system capacity, is to be achieved, are transmission formats that occupy a narrower spectrum while at the same time not causing other system impairments to increase.

Various formats have been explored over the years, largely driven by the wireless business where spectrum is more at a premium than has hitherto been the case for fiber systems. Many communications engineering texts have extensive chapters devoted to this subject [see 6–9]. Those formats that have received some attention for their potential application to optical communications include multilevel signaling or M-ary ASK [10], various forms of duo-binary signaling [11–22] (which is a subset of partial response codes), and Optical Single Sideband (OSSB) [23–28].

Additionally, much attention has been focused recently on the use of return-to-zero (RZ) signaling [29] because of its demonstrated improved immunity to fiber nonlinearities relative to NRZ. If rectangular pulses were generated with a pulse width of half the bit time, their spectral occupancy would be twice that of NRZ, as the first nulls in the frequency domain would occur at $\pm 2B$, relative to the optical carrier, and their potential spectral efficiency would be one half that of NRZ signals. With pulse shaping, however, 40% spectral efficiency at 40 Gb/s has been demonstrated [30]. Therefore, it is fruitful to examine whether the nonlinearity improvements associated with RZ signaling can be combined with more bandwidth-efficient formats to achieve still greater system performance.

The theme of this chapter, then, is to explore and review some of the work on transmission formats that would permit (or inhibit) increasing spectral efficiency, dispersion immunity, and interchannel nonlinearities. Much of this work has now progressed to the point where some reasonable statements as to future directions can and will be made.

The emphasis of this chapter will be on transmission formats that first and foremost have spectral occupancy or a transmission bandwidth on the fiber that is less than that of the true and tried (well loved?) NRZ format. It will be shown that such formats, by and large, are more immune to fiber dispersion, but that other considerations and impairments need close attention. The emphasis will be on the fundamentals of generating and transmitting these signals, with a treatment founded in the theory of linear signals familiar to communications engineers, but possibly less familiar to the optical physics community.

It is also worth noting that in this electrical engineering communications context, fiber-optic systems are classified as classical bandpass systems to which well-developed electrical engineering modulation theory applies. The distinguishing differences between optical and radio/microwave systems lies in the carrier frequency (~ 200 THz vs several GHz) and in the nature of available optical components that can generate or modulate these signals such that a direct comparison with theory can be made.

The balance of this chapter is organized as follows. Section 2 develops some basic background in modulation theory that is needed to understand how different modulation formats have different spectral occupancies and how these may be practically developed, as well as how optical modulators shape or

distort the input signal spectra before they are launched into optical fibers. We also examine how the commonly used PIN photodiode, which responds to the optical power and hence squares the time-dependent modulated electric field, significantly impacts both the time domain waveform and the spectrum of the square-law-detected signal, relative to the modulating signal at the modulator input, and the spectrum of the optical electric field at the detector input, all of which can be different.

Section 3 then examines the spectra associated with several bandwidth-efficient modulation formats; specifically M-ary ASK and poly-binary signaling with an emphasis on duo-binary and Optical Single Sideband formats where experimental work has been done to permit an assessment of their potential. Present-day understanding of the transmission characteristics of these signals, with an emphasis on the performance of single-channel 10-Gb/s transmission over NDSF, in the 1550-nm low-loss spectral region is covered in Section 4, more as an illustration of how formats behave than to claim superiority of a particular format or for a particular optical fiber.

Section 5 looks at DWDM experiments that utilize these formats and also examines various formats that combine RZ with various other forms of signal manipulation, in particular some added phase modulation, to achieve improved transmission performance. Section 6 concludes with some suggestions for potential future research.

As a general rule, papers listed in the References section contain either the first or particularly illustrative results; other papers of interest are listed in the Bibliography.

2. Amplitude Modulation and Detection

2.1 AMPLITUDE-MODULATED SIGNAL REPRESENTATION¹

Fiber-optic transmission systems, where the optical carrier frequency is in the 200-THz region and where the information to be transmitted occupies significantly less bandwidth than the optical carrier frequency, can be classified as bandpass signals. The desired outcome of an amplitude modulation process is to multiply, in the time domain, an information signal $a(t)$ (which may or may not contain a dc component) with a continuous carrier wave represented by $\cos(2\pi f_c t)$ to produce a linearly amplitude-modulated signal given by

$$s(t) = a(t) \cos(2\pi f_c t), \quad (16.2)$$

where f_c is the optical carrier frequency.

¹ Excellent material on the time and frequency domain representation of modulated signals can be found in [6–10].

If we represent $a(t)$ by

$$a(t) = a_0 + a_s(t), \quad (16.3)$$

where a_0 is a dc component and $a_s(t)$ is the information signal, then the Fourier transforms of these signals give the frequency domain representation of the baseband and modulated signals, respectively, as

$$A(f) = A_0\delta(f) + A_s(f) \quad (16.4)$$

and

$$S(f) = \frac{A_0}{2} [\delta(f - f_c) + \delta(f + f_c)] + \frac{1}{2} [A_s(f - f_c) + A_s(f + f_c)]. \quad (16.5)$$

Therefore, the desired outcome of this *linear* modulation process is the up-shifting of the frequency spectrum of the baseband signal to be centered on the optical carrier frequency, as illustrated in Fig. 16.1. Implicit in this is that the optical carrier can be represented by a pure sinusoid such that it is a delta function in the frequency domain, an approximation to which is that the linewidth of the optical source be much narrower than the bandwidth of the signal to be modulated onto the carrier.

It is important to note that it is the optical *electric field* spectrum that is represented by an up-shifted version of the baseband signal. It should also be noted that if no dc component is included in the information signal represented by Eq. 16.2, there is no unmodulated carrier present in the final modulated signal, whereas if a dc component is present, so is an unmodulated carrier. This has particular importance when a PIN photodiode detects the signal.

Distortion of the signal results from a large number of imperfections in an optical system. The most fundamental of these are:

- (1) Modulators that, except in various small signal approximations, are rarely linear in the sense of Eq. 16.2.
- (2) The commonly used PIN semiconductor detector, which responds to the *optical power* and hence it squares the time representation of the modulated optical electric field. This in turn results in time and frequency representations of the detected signal that are quite different from those of the signal applied to the modulator.
- (3) Dispersion in optical fibers, which introduces a differential time delay across the various frequency components of the modulated signal, which in turn is equivalent to a frequency-dependent phase shift [31].

These various impairments make it difficult to advantageously implement modulation formats other than the tried-and-true *intensity modulated* approaches using either NRZ or RZ formats. Progress has nevertheless been

made, and in what follows we discuss this progress and outline where necessary, how current implementation capabilities impose limitations on progress and how improvements can be made.

2.2 MODULATOR CHARACTERISTICS

Of the various approaches to modulating optical signals, really only three have seen significant adoption, namely

- (1) direct modulation of semiconductor lasers by modulating their drive current [32];
- (2) electroabsorption (EA) modulators [33];
- (3) Mach-Zehnder (MZ) modulators [34].

The directly modulated semiconductor laser and the electroabsorption modulator result in the *optical power* being modulated in response to an input modulating signal, usually with some accompanying phase modulation that can be either beneficial or detrimental to the transmission of the signal [35]. The MZ modulator, on the other hand, can be a quite flexible device that is capable of performing a variety of modulation functions on the light, and it has been used successfully in creating a variety of modulated signals that are not purely NRZ binary amplitude or intensity modulated signals.

Specifically, it has been used to create both duo-binary [13–16] and optical single-sideband signals [25,26,36,37], but invariably with some accompanying distortion in the output signal such that the transmitted signal spectrum is not an undistorted, frequency up-shifted version of the spectrum of the baseband signal. This is, undoubtedly, one of the main reasons why the potential of modulation formats other than NRZ have not fulfilled their potential.

Because the modulated output from the semiconductor laser and the EA modulator correspond to intensity modulation (with some added phase modulation), modulation formats requiring electric modulation as described by Eq. 16.2 cannot be used with these devices, and they will not be addressed further. We will examine how the MZ modulator can and has been used to generate modulated signals other than binary amplitude shift keyed (ASK) or intensity modulated (IM) signals.

Figure 16.2 is a schematic representation of a MZ modulator. The input optical electric field is represented as $E_{\text{in}} = |E_o| e^{j\omega_c t}$, where in this instance we adopt the exponential representation of the time-varying part of the optical electric field. This unmodulated signal propagates in a waveguide at the modulator input where it is split into two separate waveguides, with a split ratio that ideally is 50%. Electrodes, to which signal voltages $v_1(t)$ and $v_2(t)$ are applied, modulate the propagation constant within each waveguide, resulting in a phase modulation of the signal propagating in each arm. When these two signals are combined at the modulator output, the result is a modulated output

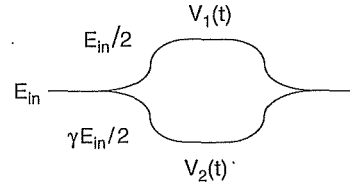


Fig. 16.2 Schematic representation of a Mach-Zehnder modulator.

electric field given by

$$E_{\text{out}}(t) = \frac{E_{\text{in}}}{2} [e^{j\varphi_1(t)} + \gamma e^{j\varphi_2(t)}] = \frac{E_{\text{in}}}{2} [e^{j(\pi v_1(t)/V_\pi)} + \gamma e^{j(\pi v_2(t)/V_\pi)}]. \quad (16.6)$$

The electric field transfer function of the device is

$$\Psi(v_1(t), v_2(t)) \equiv \frac{E_{\text{out}}(t)}{E_{\text{in}}} = \frac{1}{2} [e^{j(\pi v_1(t)/V_\pi)} + \gamma e^{j(\pi v_2(t)/V_\pi)}], \quad (16.7)$$

where v_π is the modulator extinction voltage,

$$\gamma = \frac{\sqrt{\delta} - 1}{\sqrt{\delta} + 1} \quad (16.8)$$

and δ equals the dc extinction ratio of the modulator.

To simplify, consider an ideal modulator with infinite extinction ratio such that

$$\delta \rightarrow \infty \quad \text{or} \quad \gamma = 1.$$

Then,

$$\begin{aligned} \Psi(v_1(t), v_2(t)) &= \frac{1}{2} [e^{j(\pi v_1(t)/V_\pi)} + e^{j(\pi v_2(t)/V_\pi)}] \\ &= \cos \frac{\pi}{2V_\pi} (v_1(t) - v_2(t)) [e^{j(\pi(v_1(t)+v_2(t))/2V_\pi)}]. \end{aligned} \quad (16.9)$$

$\begin{array}{cc} \uparrow & \uparrow \\ \text{Amplitude Modulation} & \text{Phase Modulation (Chirp)} \end{array}$

To avoid chirp, the time-varying portions of v_1 and v_2 are often chosen such that $v_1(t) = -v_2(t)$, and then

$$\Psi(v_1(t), v_2(t)) = \cos \frac{\pi}{2V_\pi} (v_1(t) - v_2(t)), \quad (16.10)$$

and the output optical electric field is

$$E_{\text{out}}(t) = E_{\text{in}} \Psi(v_1(t), v_2(t)) = E_{\text{in}} \cos \frac{\pi}{2V_{\pi}} (v_1(t) - v_2(t)) \quad (16.11)$$

and the output optical power is

$$P_{\text{out}}(t) = P_{\text{in}} \left| \Psi(v_1(t), v_2(t)) \right|^2 = P_{\text{in}} \cos^2 \left\{ \frac{\pi}{2V_{\pi}} (v_1(t) - v_2(t)) \right\}. \quad (16.12)$$

This is the usual way in which MZ modulator transfer functions are represented.

Figure 16.3 shows both the optical power and electric field transfer functions of the MZ modulator.

To look at the effect of a relative bias between the arms retaining chirpless modulation, let $v_2(t) = V_{\text{dc}} - v_1(t)$. Then Eq. 16.9 becomes

$$\Psi(v_1(t), V_{\text{dc}}) = \cos \left[\frac{\pi}{2V_{\pi}} (2v_1(t) - V_{\text{dc}}) \right] e^{j(\pi V_{\text{dc}}/2V_{\pi})} \quad (16.13)$$

$\uparrow$
 Modulation
Term

$\uparrow$
 Constant
Phase Shift

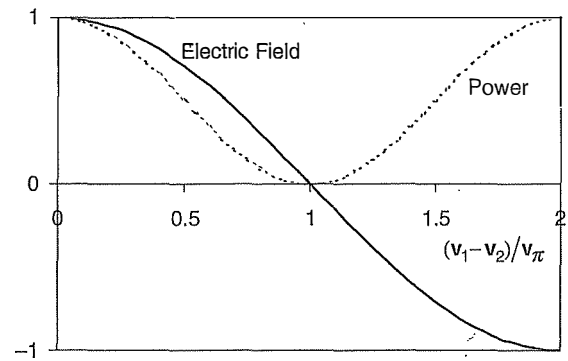


Fig. 16.3 Normalized optical electric field and power transfer functions of a Mach-Zehnder modulator.

(a) “Normal” bias $V_{dc} = V_\pi/2$

$v_1(t) = ma(t)V_\pi$ with $a(t) = +/ -1$ for a digital signal, m = electrical drive amplitude to each arm as a fraction of V_π . Then,

$$\begin{aligned}\Psi(a(t), V_\pi/2) &= \cos \left[\pi \left(ma(t) - \frac{1}{4} \right) \right] e^{j(\pi/4)} \\ &= \cos \left[\pi \left(ma(t) - \frac{1}{4} \right) \right] \quad (\text{Ignoring the constant phase shift}).\end{aligned}\tag{16.14}$$

To simplify let's invoke the small-signal condition such that $\pi ma(t) \ll 1$. Then

$$\begin{aligned}\Psi(a(t), V_\pi/2) &= \cos \left[\pi \left(ma(t) - \frac{1}{4} \right) \right] \\ &= \cos(\pi ma(t)) \cos \frac{\pi}{4} + \sin(\pi ma(t)) \sin \frac{\pi}{4} \\ &\approx \frac{1}{\sqrt{2}} [1 + \pi ma(t)]\end{aligned}\tag{16.15}$$

The output optical electric field then becomes:

$$\begin{aligned}E_{out}(a(t), V_\pi/2) &= E_{in} \Psi(a(t), V_\pi/2) \\ &\approx \frac{E_{in}}{\sqrt{2}} [1 + 2\pi ma(t)] e^{j\omega_o t} \\ &= \underbrace{\frac{E_{in}}{\sqrt{2}} [1] e^{j\omega_o t}}_{\text{Unmodulated Carrier}} + \underbrace{\frac{E_{in}}{\sqrt{2}} [2\pi ma(t)] e^{j\omega_o t}}_{\text{Modulated Carrier}}.\end{aligned}\tag{16.16}$$

The optical power transfer function is then given by

$$\begin{aligned}|\Psi(a(t), V_\pi/2)|^2 &= \cos^2 \left[\pi \left(ma(t) - \frac{1}{4} \right) \right] \\ &= \left[\cos(\pi ma(t)) \cos \frac{\pi}{4} + \sin(\pi ma(t)) \sin \frac{\pi}{4} \right]^2 \\ &\approx \frac{1}{2} [1 + \pi ma(t)]^2.\end{aligned}\tag{16.17}$$

So that in the small-signal limit, the output optical power is

$$\begin{aligned}
 P_{\text{out}}(a(t), V_{\pi}/2) &= \frac{|E_{\text{in}}|^2}{2} [1 + \pi m a(t)]^2 \\
 &= \frac{|E_{\text{in}}|^2}{2} [1 + 2\pi m a(t) + (\pi m a(t))^2] \\
 &\approx \frac{|E_{\text{in}}|^2}{2} [1 + 2\pi m a(t)].
 \end{aligned} \tag{16.18}$$

Thus, *in the small-signal limit* both the electric field and the optical power are linearly modulated by the signal $a(t)$. Also, when the modulator is biased at $V_{\pi}/2$, we have unmodulated carrier, and the modulated signal is represented in the same sense as in Eqs. 16.2 and 16.3, such that the frequency domain representation is given by Eq. 16.5.

In communications engineering parlance, this is called amplitude modulation with large carrier.

(b) Bias at extinction, $V_{\text{dc}} = V_{\pi}$

In this case, Eq. 16.13 becomes (ignoring the constant phase-shift term)

$$\begin{aligned}
 \Psi(a(t), V_{\pi}) &= \cos\left(\frac{\pi m a(t)}{V_{\pi}} - \pi/2\right) \\
 &= -\sin\left(\frac{\pi m a(t)}{V_{\pi}}\right).
 \end{aligned} \tag{16.19}$$

In the small signal approximation this becomes

$$\Psi(a(t), V_{\pi}) \approx -\frac{\pi m a(t)}{V_{\pi}}, \tag{16.20}$$

and the output optical power then also becomes

$$P(a(t), V_{\pi}) \approx \frac{|E_{\sigma}|^2}{2} \left(\frac{\pi m a(t)}{V_{\pi}}\right)^2. \tag{16.21}$$

In this bias condition and in the small-signal approximation, only the electric field is linearly modulated by the signal $a(t)$, without the presence of an unmodulated carrier, while the modulated optical power is proportional to the square of the modulating signal.

Thus, only in the small signal limit do we get the linear electric field amplitude modulation that is required for distortion free modulation.

It is important to note that optical signals are commonly detected by semiconductor PIN or avalanche photodiodes that respond to the optical power or the square magnitude of the incoming time-varying electric-field envelope.

The consequences of this, and the nonideal or nonlinear transfer function of MZ modulators other than in the small-signal approximation, for the transmission and detection of formats other than intensity modulated NRZ and RZ signals will be discussed in subsequent sections.

3. Power Spectral Densities²

As mentioned previously, for purely amplitude modulated signals, the transmitted optical electric field signal is simply a frequency up-shifted version of the baseband signal, and to discuss binary and multilevel ASK systems as well as duo-binary signals, we can equally well focus our discussion on the baseband signal, specifically the power spectral density, or PSD, of the baseband signal, since it represents the spectral occupancy of the information signal we wish to transmit. A digital baseband signal can be represented by a superposition of identically shaped, randomly weighted pulses [38]

$$a(t) = \sum_n I_n g(t - nT), \quad (16.22)$$

where $\{I_n\}$ is the information sequence, T is the time interval between pulses, and $g(t)$ is the pulse shape. The PSD of $a(t)$ can be calculated from

$$S_a(f) = \frac{1}{T} |G(f)|^2 S_i(f), \quad (16.23)$$

where $G(f)$ is the Fourier transform of the pulse shape $g(t)$ and $S(f)$ is the PSD of the information sequence.

To preview the rest of this section, Fig. 16.4 illustrates the time waveforms of the baseband signals for various binary signaling formats, as well as their PSDs. The unipolar NRZ and RZ formats are in common use and are usually implemented as intensity modulated signals. Their baseband PSDs are similar to those of Fig. 16.4, particularly as regards the presence of unmodulated tones that translate upward in frequency to the optical carrier when modulated. The other main difference is that the PSDs of Fig. 16.4b are those associated with the assumed rectangular pulses of Fig. 16.4a—real systems will have PSDs that are additionally filtered by the transfer functions of baseband electronic amplifiers and filters as well as by optical filters and amplifiers that act on the passband or modulated optical signal. The time-domain square-law detection of these signals is represented in the frequency domain by their autocorrelation function [6] which, when a large unmodulated carrier is present, can be represented as the beating of this carrier with the signal spectrum resulting in its down conversion to baseband. The beating description is in reality only an approximation, as other “beatings” occur between other unmodulated tones

² Material for this section is drawn from references [6–8].

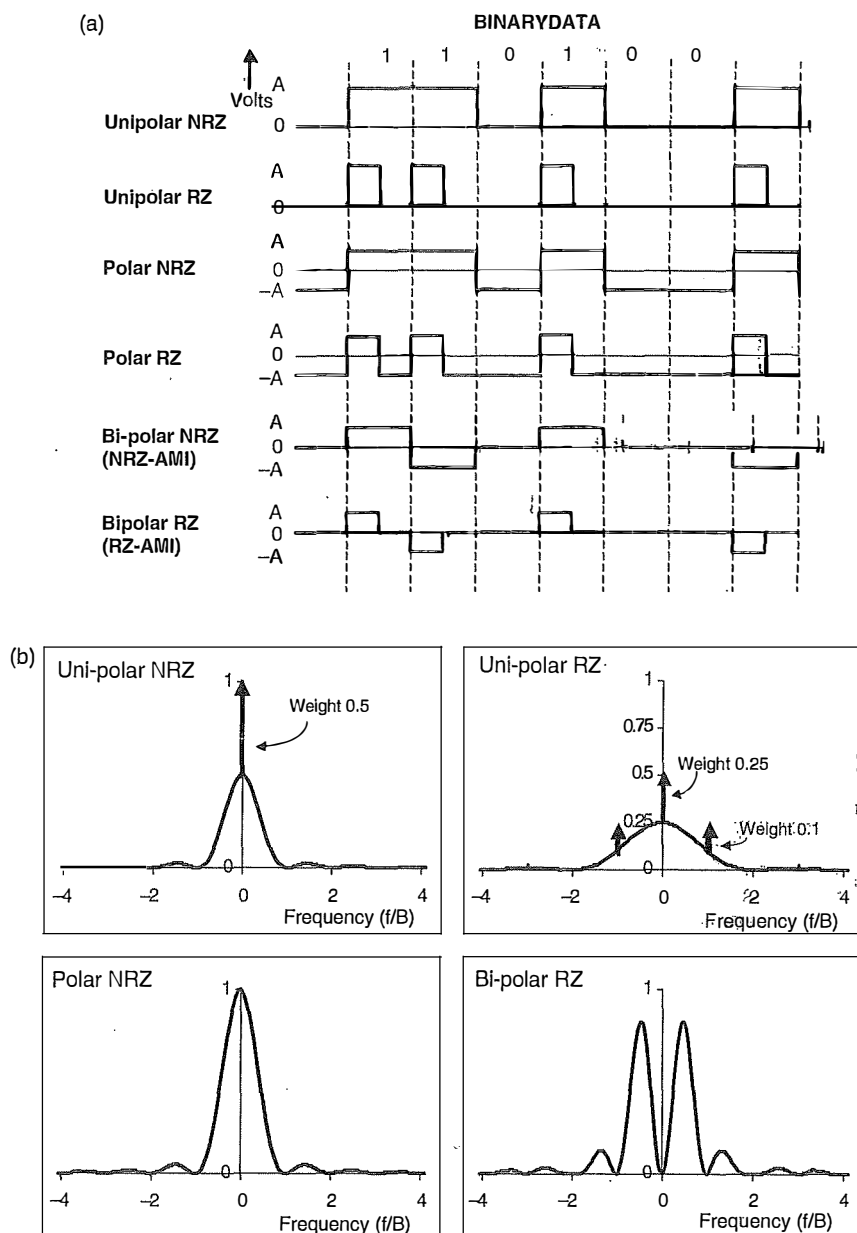


Fig. 16.4 (a) Time domain representation of various binary signal formats and (b) Power spectral density representation of binary signaling formats.

(if present) with the signal spectrum and, of course, the beating of the signal spectrum with itself.

However, this simple description does serve to help with visualizing the process. By way of an added example, if a *rectangular NRZ polar* signal was amplitude modulated onto the optical electric field using a MZ modulator biased at V_π , a purely phase modulated or phase shift keyed (PSK) signal would result. Square-law detection by a PIN photodiode would produce an unmodulated dc photocurrent. The use of coherent detection through the introduction of a local oscillator at the receiver that is both phase and frequency locked to the signal carrier would result in an amplitude shift keyed photocurrent from a PIN diode detector. Thus the presence or absence of an unmodulated carrier (often as implemented through the bias point of a MZ modulator) has major impact on the square-law-detected signal.

3.1 BINARY ASK

For a binary ASK system, the baseband signal is switched between two states (logical ONEs and logical ZEROS). For NRZ binary signaling, the information sequence consists of either a 1 or a 0, occupying the full time slot $T = 1/B$, where B is the bit rate, and these symbols occur with equal probability. If we represent the rectangular pulse shape as having amplitude A during the full time interval T the PSD becomes

$$S_{\text{NRZ},bb}(f) = \frac{A^2 T}{4} \text{sinc}^2(fT) + \frac{A^2}{4} \delta(f) = \frac{A^2}{4B} \text{sinc}^2\left(\frac{f}{B}\right) + \frac{A^2}{4} \delta(f), \quad (16.24)$$

which is illustrated in Fig. 16.1a.

When this baseband signal is used to amplitude modulate the electric field of an optical carrier, the PSD of the transmitted optical electric field becomes

$$S_{\text{NRZ},e}(f) = \frac{A^2}{8B} \left[\text{sinc}^2\left(\frac{f-f_c}{B}\right) + \text{sinc}^2\left(\frac{f+f_c}{B}\right) \right] + \frac{A^2}{8} [\delta(f-f_c) + \delta(f+f_c)], \quad (16.25)$$

which is illustrated in Fig. 16.1b.

For unipolar RZ signals with pulse width equal to one half the bit period, the PSD can be written as

$$S_{\text{RZ},bb}(f) = \frac{A^2}{4B} \text{sinc}^2\left(\frac{f}{2B}\right) \left[1 + B \sum_{n=-\infty}^{\infty} \delta(f - nB) \right]. \quad (16.26)$$

This PSD has the same sinc-function shape as the NRZ PSD, but with nulls at integer multiples of twice the bit rate, plus a dc term similar to NRZ. Additionally, there are other tones at integer multiples of the bit rate whose

amplitudes are determined by the sinc function, resulting in tones at odd integer multiples of the bit rate. This is the PSD of RZ signals with ideal-square pulses, but with spectral shaping through various modulator stages, this spectrum can be substantially reduced and shaped.

3.2 MULTILEVEL AMPLITUDE SHIFT KEYING (*M*-ary ASK) [10]

In this format, binary signal bits are converted into a binary code word where N bits are converted into a multilevel signal whose number of amplitude levels is $M = 2^N$.

If we use the aggregate bit rate B of the final signal as our bit-rate reference, then effectively, instead of transmitting bits of a single amplitude of time duration T at a rate B , we transmit symbols of time duration $T = N/B$ at a rate B/N with amplitudes ranging from zero to $M = 2^N$.

A generalization of Eq. 16.1 for the baseband power-spectral density of a square-pulse M -ary signal is

$$\begin{aligned} S_a(f) &= \frac{(M^2 - 1)A^2T}{12} \text{sinc}^2(fT) + \frac{(M - 1)^2A^2}{4} \delta(f) \\ &= \frac{(M^2 - 1)A^2T}{12} \text{sinc}^2\left(\frac{f}{B/\log_2(M)}\right) + \frac{(M - 1)^2A^2}{4} \delta(f). \end{aligned} \quad (16.27)$$

For a given level of encoding, the amplitudes of the symbols are given by the equivalent binary representation of the word it represents, and each symbol carries $N = \log_2(M)$ bits of information. Since the symbol time has been extended from $T = 1/B$ to $T = N/B$, the transmission bandwidth (measured between the first nulls in the spectrum) has been reduced from $2B$ to $2B/N = 2B/\log_2(M)$. Table 16.2 is a simple illustration of the diminishing return.

In order to construct and to detect these signals, complex linear analog circuitry is required. A relatively simple implementation for a 4-ary system is shown in Fig. 16.5.

This signal format is best transmitted as a four-level intensity-modulated signal, in that if equal amplitudes were used to modulate the electric field, the

Table 16.2 Equivalent Transmission Bandwidth of M -ary ASK

<i>Number of Levels (M)</i>	<i>Equivalent Transmission Bandwidth</i>
2	$+/-B$
4	$+/-B/2$
8	$+/-B/3$
16	$+/-B/4$
32	$+/-B/5$
64	$+/-B/6$

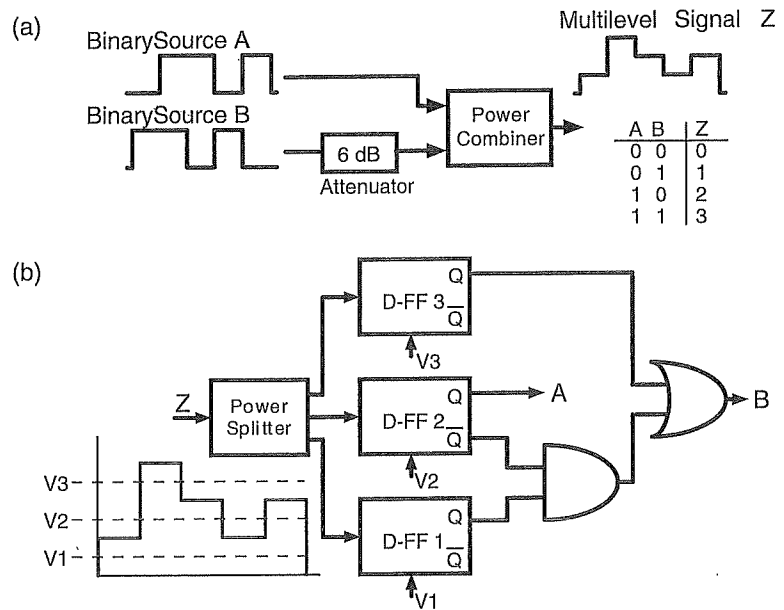


Fig. 16.5 (a) Multilevel symbol generator and (b) multilevel decoder [10].

optical power levels would have a squared (0,1,4,9) relationship—clearly not desirable.

In this case, four signal levels are created with a consequent reduction in the transmission bandwidth of a factor of two and an increase in the symbol time by a factor of two. While a beneficial factor of two is achievable, this does not translate quite so simply into an increase in the dispersion-limited transmission distance due to the fact that we now have a four-level power signal or eye, which inevitably suffers additional eye closure due to the added levels, as illustrated in Fig. 16.6. Although shown to provide additional dispersion immunity for a single-channel 10-Gb/s system over NDSF [10], the additional optical power that this requires is such that in a multichannel DWDM configuration excessive penalties due to fiber nonlinearities are likely to preclude the use of M-ary ASK signaling in long-haul DWDM systems. The potential application of M-ary ASK to transmission over dispersion-limited multimode fiber (such as for 10-Gb/s Ethernet) is a potential, but to the author's knowledge, an untried application.

3.3 DUO-BINARY SIGNAL GENERATION

Duo-binary signaling is one of a class of codes known as partial response codes in which filters are introduced to reduce the signal bandwidth but that also create a correlation between the signal sent in consecutive time slots [39–42].

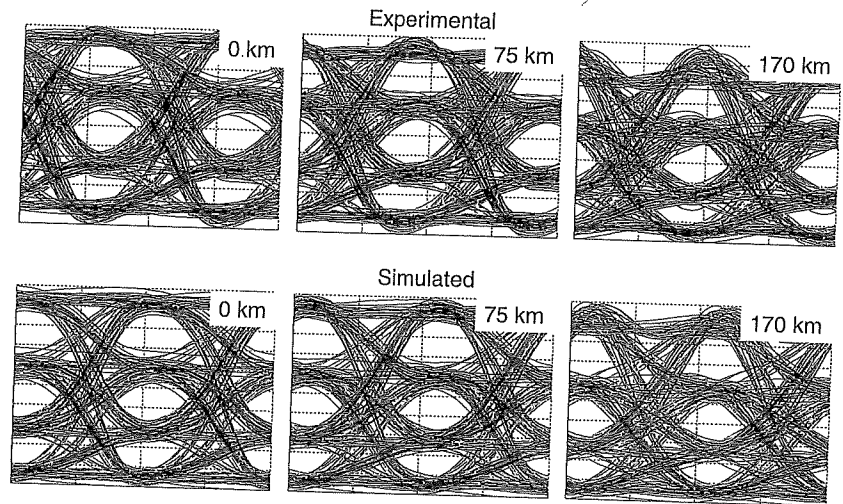


Fig. 16.6 Experimental and simulated eye diagrams for 4-ary ASK. Reprinted from [10] with permission. Copyright © 1999 IEEE.

The principle behind a poly-binary signal generator is shown in Fig. 16.7, which shows how a signal with b levels is generated by a series of delay and add circuits. A duo-binary signal is generated using only a single delay stage of duration T , which creates a three-level signal with first nulls in the spectral density at $B/2$. References [10,40] give the power-spectral density of a poly-binary signal:

$$S(f) = \frac{(b-1)^2 A^2 T}{4} \text{sinc}^2[(b-1)fT], \quad (16.28)$$

and is illustrated in Fig. 16.8 for $b = 2$ (binary), 3 (duo-binary), 5, and 7.

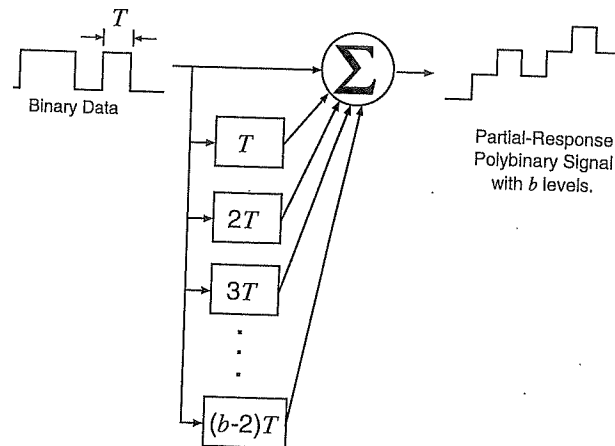


Fig. 16.7 Poly-binary signal generator (delay and add circuit).

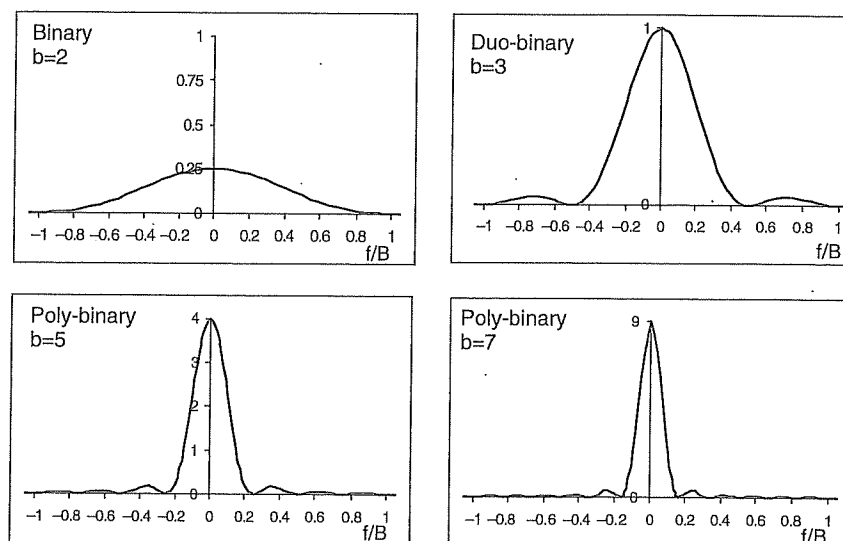


Fig. 16.8 Power spectral density for poly-binary signals.

Given that the spectral width decreases with increasing number of signal levels b , as measured to the first nulls in the PSDs, it is tempting to conclude that this form of signaling would be progressively more immune to dispersion as b is increased. That this is not the case relates to the fact that for poly-binary signaling, the symbol period is equal to $T = 1/B$, *regardless of the number of levels*. Nyquist's criterion that the transmission bandwidth be greater than $1/(2T)$ for zero intersymbol interference must still be satisfied, and this criterion relates the required transmission bandwidth to the symbol period, not the bandwidth of the PSD. Thus, the most sensible and useful subset of poly-binary signaling is the duo-binary version.

The duo-binary signal can be applied to an optical modulator to create a three-level intensity signal, as shown in Fig. 16.9a. However, a much better and also more subtle method [13–16] was implemented by creating a *three-level optical electric-field signal* using an MZ modulator that is biased at extinction, as shown in Fig. 16.9b. The effect is to create a suppressed carrier signal in the sense of Eqs. 16.1–16.3 with $a_0 = 0$ and $A_0 = 0$.

The three electric field levels are 0 and $\pm E$, which translates to only two optical intensity levels. Thus in the optical electric field domain, the signal spectrum is halved with respect to the equivalent binary signal, and it has no unmodulated carrier. When detected with a PIN photodiode that acts as a square-law detector on the electric field, two current levels result, which is a consequence of the absence of the carrier. The autocorrelation of the electric field (square-law detection) expands the signal spectrum at baseband to have first nulls at the bit rate. Because of the π phase shift between the two nonzero

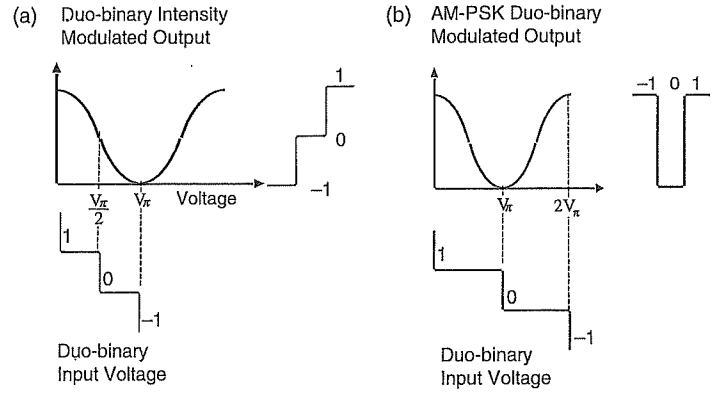


Fig. 16.9 Mach-Zehnder modulator bias and drive conditions for generating (a) a duo-binary intensity modulated signal and (b) an AM-PSK duo-binary signal.

electric field levels, this form of signaling is sometimes referred to as Amplitude Modulated-Phase Shift Keyed (AM-PSK) duo-binary.

Precoding of the input bitstream to the duo-binary encoder is required to ensure that the two detected optical power levels correspond to the original signal bits.

3.4 OPTICAL SINGLE SIDEBAND (OSSB) GENERATION³

When transmitting amplitude modulated signals in NRZ or RZ form, the transmitted spectrum is double-sided, with redundant information carried in the upper and lower sidebands on either side of the carrier. This has, of course, been recognized in cable TV transmission, where the lower sideband is partially filtered out, resulting in the commonly used amplitude modulation vestigial sideband, or AM-VSB, format. Similar methods have been illustrated in the optical domain [23,24], where a sharp cutoff filter was used to partially remove one of the sidebands at the transmitter followed by transmission on NDSF. A similar filtering approach has more recently been very successfully applied in a DWDM experiment, but in this case the optical filtering takes place at the receiver [27]. A single sideband signal can also be created using Hilbert transforms [6–9] for which the bandpass representation of the signal is

$$s_{\text{SSB},u/l}(t) = a_s(t) \cos(2\pi f_c t) \mp \hat{a}_s(t) \sin(2\pi f_c t), \quad (16.29)$$

where the upper (minus) and lower (plus) signs generate the upper and lower sidebands, respectively. $\hat{a}_s(t)$ is the Hilbert transform of the modulating signal

³ The fundamentals of single sideband theory can be found in [6–9], with much added detail in various original papers listed in the bibliography.

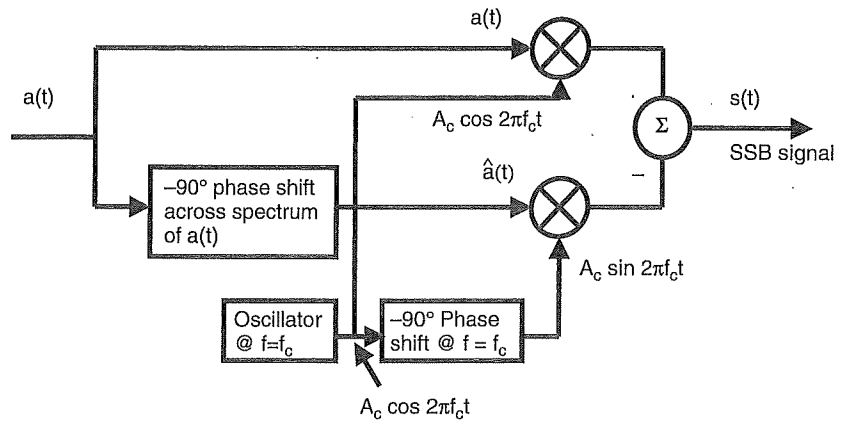


Fig. 16.10 Hartley modulator structure for generating single sideband signals.

and imparts a 90-degree phase shift to all frequencies in the information signal. Figure 16.10 shows a block diagram for the creation of the signal of Eq. 16.29.

This mathematical construct can be implemented for both broadband baseband signals, as well as for sub-carrier RF and microwave signals.

3.4.1 Subcarrier OSSB

Transmission of microwave subcarrier modulation in single-sideband form is both highly beneficial from the standpoint of fiber dispersion and relatively easy to implement [36,37].

In conventional optical subcarrier modulation, the subcarrier appears in the optical frequency domain on both sides of the optical carrier, as illustrated in Fig. 16.11. Square-law detection of this signal causes the optical carrier to beat with both the upper and lower sidetones, and as long as there is no phase difference between these sidetones, the signal is detected efficiently because the two resultant baseband-signal components are in phase. Fiber dispersion causes a phase shift between these sidetones, and when each is shifted in opposite directions by 90° from the carrier, and this carrier beats against these sidetones, the result is total cancellation of the signal [43,44].

Mathematically, the optical electric-field transfer function due to first-order dispersion in a fiber can be represented by

$$H(f, z) = \exp \left(j \frac{\pi D \lambda_0^2 f^2}{c} z \right), \quad (16.30)$$

where D is the fiber dispersion parameter, λ_0 is the wavelength, and f is the frequency offset from the optical carrier.

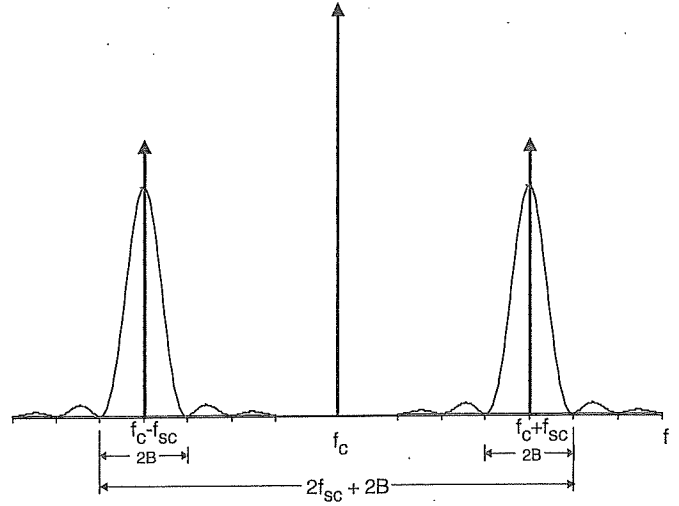


Fig. 16.11 Illustration of double sidebands associated with subcarrier modulation (only positive frequencies shown).

If a double sideband sinusoidal signal at modulation frequency f_{sc} according to Eq. 16.3 propagates down a fiber whose transfer function is given by Eq. 16.30, the frequency domain representation of the optical electric field at distance z down the fiber is

$$E(f, z) = [2\pi\delta(f - f_0) + m\pi \{ \delta[f - (f_0 - f_{sc})] + \delta[f - (f_0 + f_{sc})] \}] \times \exp\left(j\frac{\pi D\lambda_0^2 f^2}{c}z\right). \quad (16.31)$$

The square-law-detected signal is the Fourier transform, F , of the squared envelope operator in the time domain (i.e., the autocorrelation of the Fourier transform of the electric field) and is given by

$$F(|E(t)|^2) = \int_{-\infty}^{\infty} F_E(f)F_E^*(f + \xi) df = A_{ff}(\xi). \quad (16.32)$$

If we consider only the fundamental term in A_{ff} at f_{sc} resulting from Eq. 16.32, we get:

$$A_{ff}(f_{sc}) = [2\pi\delta(f - f_{sc})m] \left[\exp\left(j\frac{\pi D\lambda_0^2 f^2}{c}z\right) + \exp\left(-j\frac{\pi D\lambda_0^2 f^2}{c}z\right) \right]. \quad (16.33)$$

If we take the inverse Fourier transform of Eq. 16.33 and then take the real part of this, we get the frequency dependence of the detected photocurrent as

$$\begin{aligned} I(t) &\propto 2\pi^2 \left[\cos \left(2\pi f_{sc} t + \pi \frac{\pi D \lambda_0^2 f_{sc}^2}{c} z \right) + \cos \left(2\pi f_{sc} t - \pi \frac{\pi D \lambda_0^2 f_{sc}^2}{c} z \right) \right] \\ &\propto \cos(2\pi f_{sc} t) \cos \left(\pi \frac{\pi D \lambda_0^2 f_{sc}^2}{c} z \right). \end{aligned} \quad (16.34)$$

Thus the *detected RF power* at the subcarrier frequency will vary with distance and subcarrier frequency according to

$$P(f, z) \propto I^2(f, z) \propto \cos^2 \left(\frac{\pi D \lambda_0^2 f^2}{c} z \right). \quad (16.35)$$

If one of the sidebands is removed, as in single-sideband modulation, the detector output current becomes

$$I(t) \propto \cos \left(2\pi f_{sc} t + \pi \frac{\pi D \lambda_0^2 f_{sc}^2}{c} z \right) \text{ or } \cos \left(2\pi f_{sc} t - \pi \frac{\pi D \lambda_0^2 f_{sc}^2}{c} z \right), \quad (16.36)$$

depending on which sideband is removed. Note that both terms in Eq. 16.36 are constant in power as a function of fiber length.

Since most microwave subcarrier transmission occurs with relatively narrowband information signals, the result is that only dispersion across the narrow information spectrum is of relevance, and this is small for the distances that subcarrier systems have been implemented.

To implement an RF OSSB system using the modulator structure of Fig. 16.10, the sine and cosine carriers can be simply generated in an MZ modulator through the application of a differential dc bias voltage equal to $V_\pi/2$ to the two arms, which creates a 90-degree phase shift between the optical carriers in the two arms of the MZ modulator. The two arms of the MZ modulator are then modulated by the subcarrier, with a differential time delay between the drives to the arms corresponding to a $\pi/2$ phase shift of the subcarrier frequency. Given that the information signal is narrow band, this also adequately Hilbert transforms the information signal.

Two illustrations of the impact of OSSB on the dispersion-limited transmission of a 17-GHz subcarrier on NDSF are illustrated in Fig. 16.12 [37]. Figure 16.12a compares the detected RF power of a 17.16-GHz subcarrier when transmitted at 1550 nm over NDSF using conventional double sideband modulation and OSSB as implemented previously. Figure 16.12b illustrates performance at 17.16 GHz when using the MZ modulator in two significantly nonlinear configurations.

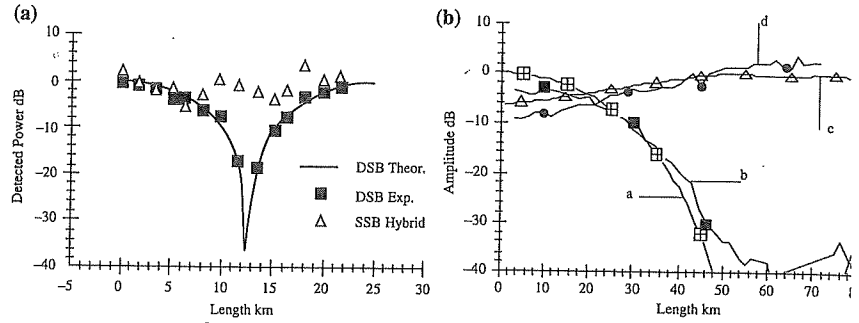


Fig. 16.12 (a) Normalized detected RF power at 17.16 GHz vs fiber length. The RF power is normalized relative to the received optical power, and is given in arbitrary dB units. Reprinted from [37] with permission. Copyright © 1998 IEEE. (b) Change in detected RF power at 17.16 GHz (relative to the received dc optical power) vs transmission distance. (a) Experiment and (b) simulation for the harmonically upconverted double sideband spectrum. (c) Experiment and (d) simulation for the harmonically upconverted single sideband spectrum [37].

Reference to additional work on various modulator structures that implement microwave subcarrier transmission in OSSB form or that use the nonlinear transfer function of the MZ modulator combined with the square-law-detection properties of PIN photodiodes to harmonically upconvert microwave signals without significant fiber dispersion can be found in the bibliography.

3.4.2 Broadband OSSB

The same Hartley modulator principle can be used to generate a broadband, digital OSSB signal. However, as was shown earlier, for chirp-free transmission, the MZ amplitude modulator needs to be driven in a symmetric push-pull fashion to avoid chirping the amplitude modulated output, thereby incurring excess dispersion penalty on transmission. Therefore, a more appropriate and chirp-free implementation for broadband OSSB transmission is required.

An amplitude-modulated single-sideband signal can be obtained using the complex envelope of the baseband signal, given by [6,45–47]:

$$g(t) = a_s(t) \pm j\hat{a}_s(t). \quad (16.37)$$

This signal gives rise to the bandpass SSB signal of Eq. 16.29, but which can be shown to have both amplitude and phase modulation.

The amplitude modulation component is the real part of the envelope:

$$R(t) = |g(t)| = \sqrt{a_s^2(t) + \hat{a}_s^2(t)}, \quad (16.38)$$

whereas the phase modulation component is

$$\varphi(t) = \angle g(t) = \tan^{-1} \left[\frac{\pm \hat{a}_s(t)}{a_s(t)} \right]. \quad (16.39)$$

This is equivalent to cascading a chirp-free amplitude modulator, driven by the real part of the signal envelope, with a phase modulator that is driven by $\phi(t)$. The original signal will be recovered on *coherent* detection.

In the noncoherent-detection optical case, however, we are detecting the signal power envelope, for which a reasonable approximation to an OSSB signal, which can be envelope detected, results if it is generated according to

$$g_r(t) = a(t) \exp(j\hat{a}(t)), \quad (16.40)$$

where the signal is amplitude modulated as usual (without chirp) and phase modulated with the Hilbert transform of the information signal, as illustrated in Fig. 16.13. This requires ideal amplitude modulation, which does not take place in an MZ modulator (except as in the small-signal approximation), nor is it a simple matter to generate a Hilbert transform of a broadband signal that spans the bandwidth required by, for example, a $2^{23} - 1$ or a $2^{31} - 1$ pseudo-random binary sequence as required for SONET system characterization.

A reasonable, vestigial sideband signal was generated in [48] using a tapped-delay-line filter approximation to a Hilbert transformer. Figure 16.14 shows the measured PSDs of a 10-Gb/s signal in double-sideband form (before applying the Hilbert transform) and in single-sideband form after applying the approximate Hilbert transform to a cascaded AM + PM OSSB modulator.

It should be noted that with this OSSB modulator the signal power of the unwanted sideband is transferred to the wanted sideband, whereas with the optical filtering method it is lost.

If a relatively small modulating signal is applied to the amplitude modulator, we meet the criteria for Eq. 16.40. Furthermore, the square-law-detected signal will then also be a close approximation to the received electric field, as noted earlier in Section 2.2, with the result that the phase characteristics of the signal are transferred to the detected signal photocurrent. This will permit dispersion compensation in the electrical domain at the output of the linear

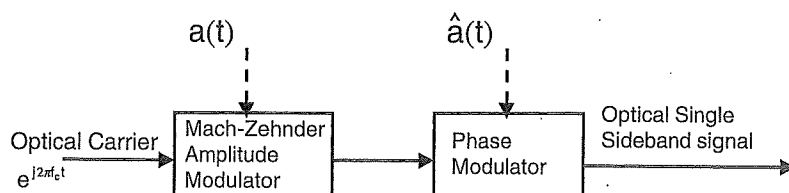


Fig. 16.13 Cascaded Mach-Zehnder amplitude and phase modulator for generating OSSB.

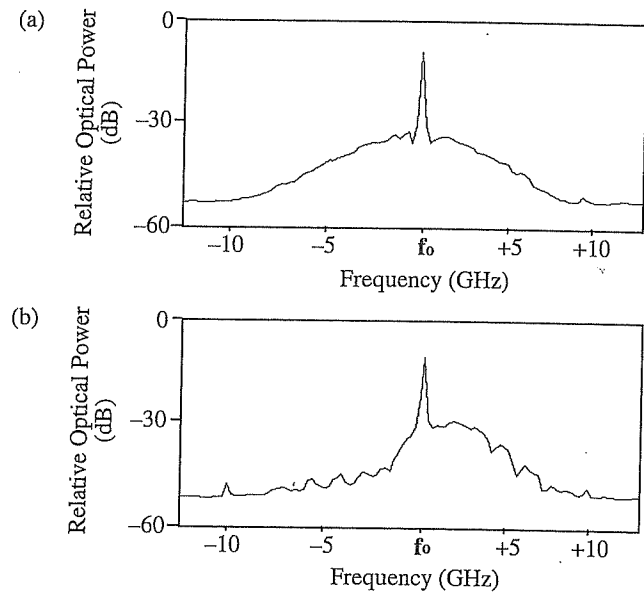


Fig. 16.14 Experimental optical spectra for (a) DSB and (b) OSSB signals at 10 Gb/s [48].

portion of the optical receiver. In a sense, we have created a “self-homodyne” coherent detection system, and it is interesting to contemplate whether a full coherent detection system would permit large signal modulation at the transmitter with a reduction in the initial eye opening penalty associated with the small modulation depth. Alternatively, if a suppressed carrier OSSB signal is transmitted, and this signal is coherently detected at the receiver, it should be possible to perform dispersion compensation quite accurately at the output of the linear portion of the receiver, provided the received signal has not been corrupted by nonlinear effects in the fiber.

4. Single-Channel Transmission Results

4.1 *M*-ary ASK TRANSMISSION [10]

The simulations in Fig. 16.15 show how the dispersion-induced reduction in receiver sensitivity (at a bit error ratio of 10^{-9}) varies with fiber length for several ASK formats for NDSF at an aggregate bit rate of 10 Gb/s at a wavelength in the 1550-nm window. Clearly, the dispersion limitation has been improved in all cases. In the case of the 4-ary signal, because of the need to detect the signal with a given bit error ratio (BER), the signal-to-noise ratio within each of the three eyes needs to be essentially the same as for the equivalent two-level

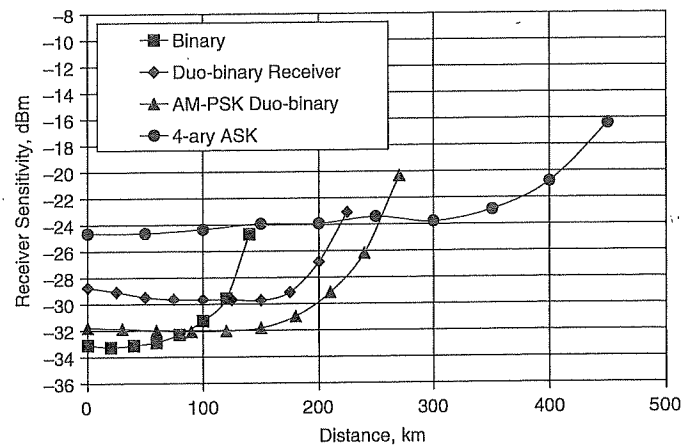


Fig. 16.15 Comparison of dispersion-induced receiver sensitivity degradation for NRZ, duo-binary, and M-ary signaling at 10 Gb/s. Reprinted from [10] with permission. Copyright © 1999 IEEE.

or binary signal. This requires significantly greater optical power, which is the reason for the offset of the M-ary ASK curve in Fig. 16.15. More significant, however, is the increase in launched power that is necessary to achieve a given BER after a number of fiber links that have EDFAs along the way. This is illustrated in Fig. 16.16, where we show the launched power necessary to achieve a BER of 10^{-9} for 4-ary and 8-ary signaling with the NDSF link configurations as given in the figure captions.

Clearly, this large increase in launched power precludes the use of M-ary ASK for multichannel DWDM long-haul systems as, at the power levels shown, huge interchannel nonlinearity problems can be expected to occur due to FWM and XPM. Some degradation in the single-channel case can also be expected due to self-phase modulation (SPM).

Thus, while not practical for long-distance transmission, an unexplored possible application of M-ary ASK is for moderate-distance extension of transmission distances over improved multimode fiber, such as for 10-Gb/s Ethernet.

4.2 DUO-BINARY TRANSMISSION

Duo-binary intensity modulation has been used to extend 10-Gb/s transmission on NDSF to 140 km [12], but this three-level intensity format will have the same optical power and nonlinearity issues as associated with M-ary ASK signals. An alternative approach to improving dispersion-limited transmission via duo-binary signaling is to transmit the signal in binary form, as usual, but to convert it to a duo-binary signal using electrical filters after detection in

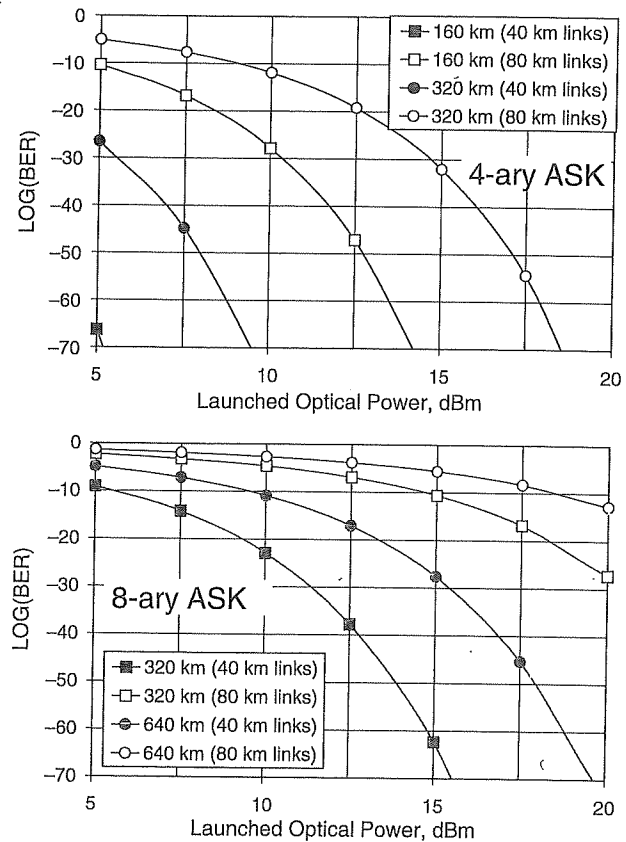


Fig. 16.16 Launched power necessary to achieve a 10^{-9} BER for various M-ary ASK formats at 10 Gb/s over NDSF. Reprinted from [10] with permission. Copyright © 1999 IEEE.

the receiver. Improvement in transmission distance at 10 Gb/s on NDSF to ~ 160 km was achieved by this means [11].

The best duo-binary single-channel transmission results have been obtained using AM-PSK duo-binary signaling where distances up to ~ 225 km at 10 Gb/s at 1550 nm over NDSF have been reported [13–16].

One important item to note is that in order to achieve improved dispersion-limited transmission with this format, it is necessary to provide extra filtering after the delay-and-add filter in order to remove those parts of the baseband signal spectrum that occur at frequencies above $B/2$ [17]. This is because the delay-and-add filter is in fact periodic in frequency, and does not continuously attenuate frequencies above $B/2$. These higher-frequency components of the signal, which are present at sufficient power, undergo sufficient dispersion to substantially reduce the signal eye opening, as illustrated in Fig. 16.17. The

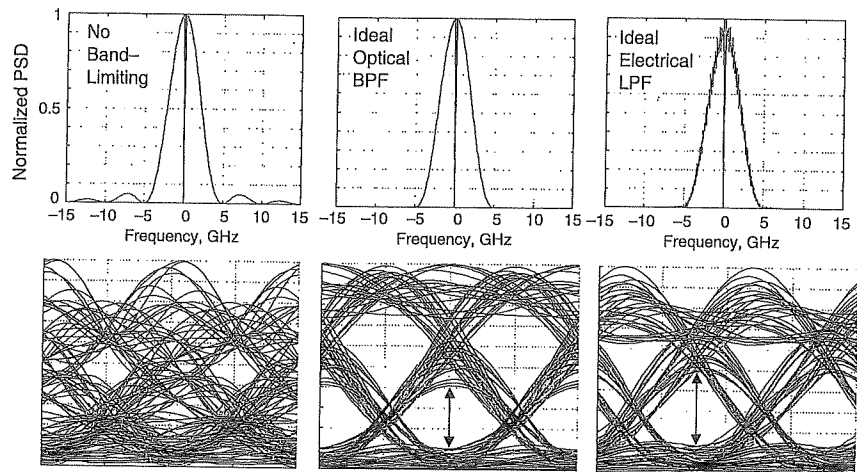


Fig. 16.17 10 Gb/s power spectral density of the optical electric field (*top*) and corresponding filtered receiver eye diagrams (*bottom*) after transmission over 200 km of NDSF for an optical AM-PSK duo-binary signal [10].

removal of the higher frequencies can be achieved by a combination of the delay-and-add filter followed by a sharp cutoff filter whose bandwidth is $\sim B/2$ or an approximation to this combined filter can be implemented with a single filter whose bandwidth is $\sim B/4$. The sensitivity of the dispersion immunity to the details of the filtering has been discussed in [17], and likely is why a great variability in experimentally extending single-channel transmission distance has been reported. Whether this filtering is necessary in systems that compensate for dispersion using optical dispersion compensation techniques has not yet been investigated.

4.3 OSSB TRANSMISSION WITH HILBERT-TRANSFORMED SIGNALS

In the first such effort [25,26,48], the focus was on obtaining maximum dispersion-limited transmission distance without optical dispersion compensation by focusing on a combination of the lower transmission bandwidth or spectral occupancy of the OSSB transmission spectrum, combined with the prospect of doing postdetection dispersion compensation using electrical methods at the output of the linear stages of the receiver.

Significant improvement in the dispersion-limited transmission distance for a single OC-192 channel operating at 1550 nm over NDSF was obtained by this method, as illustrated in Fig. 16.18. A degree of electrical dispersion compensation was also achieved with a dispersive co-planar waveguide at the output of the linear portion of the receiver. This is possible because a quasi-homodyne detection system was created by using a moderate extinction ratio

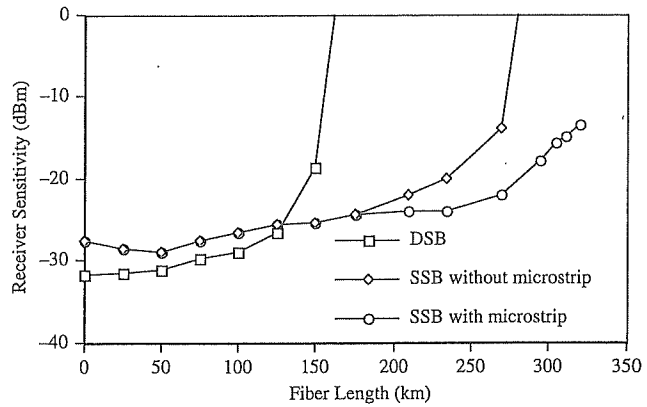


Fig. 16.18 Experimental receiver sensitivity vs fiber length for 10 Gb/s with six EDFAs at a BER of 10^{-9} . Reprinted from [48] with permission. Copyright © 1999 IEEE.

at the transmitter, the result of which is that the output of the PIN detector, which is always the autocorrelation of the optical electric field, is dominated by the product of the unmodulated optical carrier and the single sideband optical electric field spectrum, effectively translating this spectrum to baseband. Since dispersion in the fiber causes pure phase distortion (in the absence of fiber nonlinearities), the dispersion can be reversed with an appropriate phase filter at the receiver output.

5. Multichannel Transmission

5.1 DUO-BINARY DWDM

Recognizing the potential dispersion and spectral efficiency benefits of AM-PSK duo-binary signaling, significant progress has been reported in using this format for very dense WDM transmission, as well as for transmission over very long distances [49–53].

Conventional NRZ duo-binary DWDM transmission has been reported with 100% spectral efficiency in a single span of 100 km. Long distance systems have not been reported, as progress in suppressed carrier/alternating phase RZ transmission has led to this being a more nonlinearity-immune transmission format.

5.2 OSSB DWDM

Several multichannel WDM OSSB experiments have recently been reported that have resulted in significantly improved spectral efficiency. In one particular experiment at 40 Gb/s, a carefully designed channel plan was implemented in

which the signals were transmitted in double-sideband form using RZ modulation, allowing overlap of the transmitted signals to occur [27]. Portions of the overlapping spectra were subsequently filtered out at the receiver by optical means to create alternating upper and lower vestigial sideband signals. Spectral efficiencies of 66% were obtained with transmission over 300 km of NZ-DSF. Doing the filtering at the transmitter does not provide as much immunity to nonlinear impairments as filtering at the receiver, as nonlinear products apparently fill the filtered spectrum. Whether performing such filtering at both the transmitter and at the receiver would provide benefit has not yet been explored.

As pointed out in [48], electrical dispersion compensation with OSSB detection is possible, but the effectiveness depends on the modulation depth of the OSSB signal in that it relies on the creation of what was called a “self-homodyne” detection capability. Combining this capability with VSB generation through optical filtering at the receiver also remains unexplored.

A recent comparison [54] of nonlinear impairments of NRZ, RZ, and OSSB (as implemented using Hilbert transforms) indicated no strong advantage for OSSB generated in this way. This would appear to be quite consistent with the underlying physical mechanisms that have resulted in significant progress made to indicate that chirped RZ and alternating phase RZ do indeed seem to be the formats that are most immune to nonlinearities that have been implemented and tested to date.

5.3 MODIFIED RZ DWDM

Of recent efforts, probably the most significant single practical advance in the use of modulation techniques has been in the application of added phase modulation in conjunction with RZ transmission. Many groups have reported [29] on the application of added phase modulation by the application of a half-bit-rate clock to a second MZ modulator biased at V_π , which is added in series to a conventional NRZ data modulator. This results in the generation of RZ bits with alternating phase, and it also results in suppressing the unmodulated carrier. It also adds tones at odd integers of $\pm B/2$ away from the carrier frequency. The formalisms developed earlier in this chapter are directly applicable to analyzing such cascaded structures.

Although improvements in signal quality only in the 1-dBQ range have been achieved, this is very significant in ultra-long-haul systems, particularly when used in conjunction with forward error correction (FEC). Figure 16.19 illustrates the power spectral densities obtained in this way, for which a 0.6-dBQ improvement was reported [55]. In a comparison of NRZ, RZ, and chirped RZ at 12.3 Gb/s, Bakhshi and colleagues at Tyco [56] concluded that chirped RZ with a phase modulation index close to $\pi/2$ at a phase modulation frequency equal to the bit rate provided improvement over RZ in a very long system, but that careful balancing of nonlinear effects and spectral broadening from the chirping was essential.

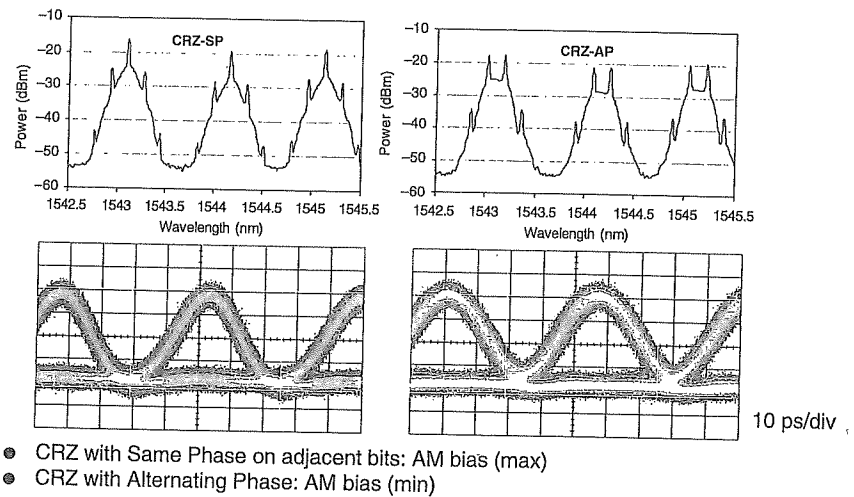


Fig. 16.19 Chirped RZ signals with same phase and alternating phase on adjacent bits [55]. Slide courtesy N. G. Bergano.

More recently, Cheng and Conradi [57] compared RZ, RZ with alternating phase, and two forms of RZ duo-binary at 40 Gb/s and concluded that of these, RZ with alternating phase and a modified form of RZ duo-binary, created by a delay-and-subtract as opposed to a delay-and-add filter, would perform best, with the modified duo-binary signaling having nearly a 1 dBQ advantage over the alternating phase RZ. Although the simulation comparisons were carried out over a very limited range of parameters, and the need for more work is apparent, the conclusion could nevertheless be drawn that the suppression of the optical carrier (common to both formats) and the suppression of the other tones at $\pm N \cdot B/2$ from the carrier (which is the case for the modified duo-binary format) can *both* help to reduce *interchannel* FWM and the impact of ghost pulses resulting from *intrachannel* FWM. The modified duo-binary signal also has a significantly reduced and reshaped PSD, as illustrated in Fig. 16.20, and this too will alleviate the impact of chromatic dispersion.

While it is often difficult to elucidate the physical mechanism responsible for any particular improvement in the transmission performance resulting from the use of a particular transmission format, the conclusions on the effect of the presence or absence of carriers and tones, and the effect that alternating the phase of sequential bits has on the width and shape of the transmitted signal spectrum, does provide useful guidance for future work. In particular, it is worth noting that for decades the copper wireline T1 system at 1.544 Mb/s has been using what is variably known as bipolar RZ or Alternate Mark Inversion (AMI), in which sequential marks or ONE bits are transmitted as +ve and -ve voltages [6]. When transmitted as a bandpass optical signal, this will result in alternating the phase of sequential marks or ONE bits, not the phase

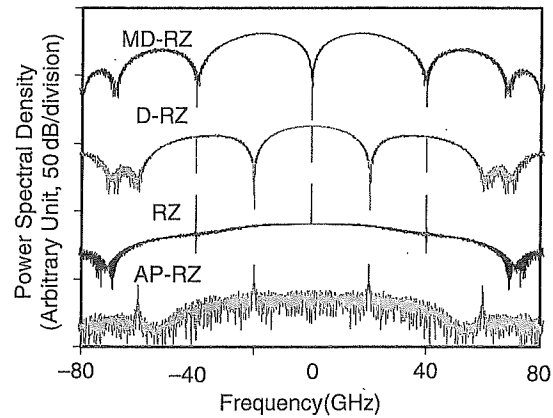


Fig. 16.20 Simulated power spectral densities of various RZ formats (after [57]).

of sequential bits. In Fig. 16.4 we illustrated the PSDs of various RZ formats, which shows that AMI has (a) no tones in the baseband spectrum that would result in unmodulated tones in the transmitted spectrum; (b) a first null at the bit rate, resulting in a similar PSD to the modified (delay and subtract) duo-binary signal [57]; and (c) significantly reduced low-frequency content. The detection of this signal with a square law (i.e., PIN detector) produces a unipolar RZ photocurrent version of the original bit sequence. A transmission spectrum such as this could result in low impact due to dispersion and fiber nonlinearities, and potentially could be easy to generate in single-sideband form by optical filtering.

6. Discussion

In this chapter we have attempted to bring to the fore some theoretical and implementation issues surrounding the possible advantageous application of modulation formats other than unipolar NRZ and RZ to fiber-optic communication systems. This has been illustrated with examples from the literature to elucidate the properties of signal formats in various single and multichannel optical transmission configurations, and how the work may be extended to further improve system performance.

Multilevel modulation is capable of significant increases in spectral efficiency in radio systems where noise in signal (shot noise, signal-spontaneous beat noise) and power-dependent channel nonlinearities are absent. In optical systems where noise in signal is the limiting noise source, multilevel intensity modulation, we believe, will not have practical application in long-haul systems due to the excessive power required that would induce much larger interchannel nonlinearity penalties than binary signaling formats. This would then preclude the use of multilevel *amplitude* modulation formats that could

otherwise improve spectral efficiency beyond what has already been demonstrated through the combined use of polarization-division multiplexing and such formats as chirped or alternating phase RZ, modified duo-binary RZ, OSSB, and potentially AMI.⁴ The potential application of multilevel signaling to data transmission systems on multimode fiber remains an open issue.

However, it is intriguing to speculate whether with improvements in optical source stability now being required by closely spaced DWDM channels, if coherent transmission might just become practical. If this were practical from the standpoint of optical source stability, both for transmission lasers as well as for the phase and/or frequency-locked local oscillator lasers needed at the receiver, then a whole host of additional possibilities open up. For instance, with the lower launched powers that distributed Raman amplifiers offer, the fiber portion of the transmission link is now very much closer to operating free of nonlinear distortions due to SPM, FWM, and XPM. Then the combination of OSSB with coherent detection might permit adaptive electrical dispersion compensators to be used based on, for example, electrical tapped delay line filters. If this were the case, then most of the needed dispersion compensation could be implemented using the usual very broadband optical means such as dispersion-compensating fiber, but that fine-tuning of the signal eye could be done electrically.

Additionally, coherent detection of *phase modulated* channels might take advantage of the constant envelope of phase modulated channels to reduce the impact of nonlinear impairments resulting from FWM and XPM since amplitude fluctuations, other than those induced by dispersion, would be absent. This too, might allow the application of multilevel PSK formats that have narrow signal spectra, thereby permitting the increased spectral efficiency denied by fiber nonlinearities in multilevel ASK formats. (Note the power levels in Fig. 16.16.)

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⁴ This is in contradiction to a recent paper in *Nature*. P. P. Mitra and J. B. Stark, "Nonlinear limits to the information capacity of optical fibre communications," *Nature*, vol. 411, pp. 1027–1030, 28 June 2001.

and Sing Cheng contributed original work on the various modulation formats that is summarized here and in their publications. Additionally, I would like to acknowledge Prof. David Dodds of TRILabs and the University of Saskatchewan for his contributions and collaboration in the evolution of the work on optical single-sideband modulation.

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