Bit Error Probability Reduction in Direct Detection Optical Receivers Using RZ Coding

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***Abstract*—We analyze the bit error probability reduction for direct detection ON–OFF keying optical receivers using return-to-zero (RZ) coding instead of the nonreturn-to-zero (NRZ) format. For the same average optical power, RZ is shown to outperform NRZ, even when employing the same receiver bandwidth. Results are given for receivers whose noise variance is i) dominated by a signal-*independent* term (e.g., simple pin diode receivers), ii) dominated by a signal-*dependent* term (e.g., optically preamplified receivers), and iii) made up of two equally important contributions [e.g., avalanche photodiode (APD) receivers]. Based on semianalytic simulations including intersymbol interference, we show that the achievable RZ sensitivity gain is typically less for dominating signal-independent noise than for dominating signal-dependent noise, where it amounts to about 3 dB. We also quantitatively discuss the influence of the optical pulse shape on the achievable RZ coding gain, and show that finite extinction ratios can significantly reduce that gain, especially when the RZ signals are produced by direct-modulation methods.**

***Index Terms*—Extinction ratio, impulsive coding, noise, nonreturn-to-zero (NRZ), ON–OFF keying, optical receiver, optimum bandwidth, receiver sensitivity, return-to-zero (RZ).**

I. Introduction

RECENT experimental results [1]–[3] have shown that the sensitivity of direct detection optical receivers can be improved using return-to-zero (RZ) coding instead of nonreturn-to-zero (NRZ) as a modulation format, even if the receiver bandwidth is optimized for the NRZ case. This effect is observed, provided that the average optical power at the receiver remains constant, and may lead to significant system improvements, especially for free-space transmission where dispersion does not set a lower limit on the RZ pulse duration [4], [5].

First signal-to-noise ratio (SNR)-based analyses of potential receiver sensitivity improvements using RZ coding date back to the work of Personick [6] and have been mentioned in several text books since (e.g., [7], [8]). Recently, the subject has been taken up again on a more general level [1], [9], although still based on SNR considerations and without quantifying the effect of intersymbol interference (ISI). To the best of our knowledge, this important effect has so far only been included by Boivin [10] for an optically preamplified receiver in a similar context. Because ISI turns out to have different impact on RZ and NRZ, its inclusion is indispensable for quantitative comparisons of the two modulation schemes. Based on numerical simulations of the bit error probability (BEP) including ISI, it is the aim of this paper to give quantitative predictions of the potential RZ sensitivity gain for various types of receivers (with different types of noise), different optical pulse shapes, and different duty cycles.

In Section II, we outline the receiver structure under consideration and address the modeling of signal and noise underlying our simulations. Section III provides a brief description of the simulation method. Simulation results are presented in Section IV. We show that—as noted in [9]—the improvement of RZ coding depends on whether receiver noise is dominated by a signal-*dependent* or a signal-*independent* component. In contrast to the SNR-based predictions, however, the case where receiver noise is dominated by a signal-dependent term will be seen to yield the higher RZ gain. In Section V, we discuss the influence of various pulse shapes on the receiver performance. Section VI then covers the influence of a finite extinction ratio. For this purpose, we distinguish between two different methods of generating pulses, which are not equally sensitive with respect to an imperfect extinction ratio.

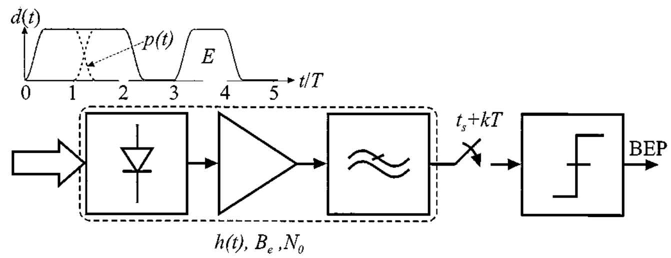

Fig. 1. General receiver structure. The optical input signal $d(t)$ is composed of individual pulses $p(t)$, each of energy E. It is detected and undergoes appropriate electronic filtering [impulse response $h(t)$, bandwidth B_e] before being sampled at $t_s + kT$ (T being the bit duration, $k \in \mathbb{Z}$, and t_s a suitable sampling delay). A threshold decision circuit is then used to regenerate the binary data sequence.

II. Modeling of Signal and Noise

For BEP evaluation, we assume a receiver structure as shown in Fig. 1. The shape of the incoming optical data signal $d(t)$ is specified in front of the photodiode; it comprises all influences of the channel on the transmit signal as well as possible distortions caused by optical prefiltering. For a fair comparison between RZ- and NRZ-coded systems, it is important to keep constant the *average* optical input power to the receiver. As a consequence, the energy E of a single pulse with optical power $p(t)$ (representing a single "1" bit) is independent of the pulse duration. After conversion to an electric current, the detected

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TABLE I
SUMMARY OF THE CONSTANTS K, C, AND $N(B_e)$ FOR VARIOUS RECEIVER TYPES

	pin receiver	optically preamplified receiver	APD receiver
$K[AW^{-1}]$	S	S	SM
$C[A^2sW^{-1}]$	Se	$2N_{ASE}S^2 + Se$	SeM^2F_{APD}
$N(B_e)[A^2s]$	N_0^2 +dark current terms	$2S^2N_{ASE}^2(2B_o - B_e) + N_0^2$	N_0^2 +dark current terms

signal is amplified and low-pass filtered. The impulse response of the entire electronics is denoted $h(t)$ and normalized to unit area $\int_{-\infty}^{\infty} h(t)\,dt = 1$. Electronic noise is accounted for by an additive equivalent noise current density N_0. After filtering, the signal is sampled and applied to a threshold decision circuit.

The mean of the photocurrent at the decision gate is given as (see, e.g., [11])

$$i(t) = K(d * h)(t) \tag{1}$$

where the symbol $*$ denotes a convolution and K is an appropriate constant, specified in more detail below. The photocurrent's variance can generally be written as

$$\sigma^2(t) = \sigma_{\text{dep}}^2(t) + \sigma_{\text{indep}}^2 \tag{2}$$

with the signal-dependent part $\sigma_{\text{dep}}^2(t)$ and the signal-independent part σ_{indep}^2. The total receiver noise $\sigma^2(t)$ is assumed to be additive and Gaussian-distributed, which leads to computationally simple expressions that are reasonably accurate [12].

From [9] and [11], we know that the signal-dependent term in (2) can generally be expressed as

$$\sigma_{\text{dep}}^2(t) = C(d * h^2)(t) \tag{3}$$

C being an appropriate constant. The signal-independent noise contribution, on the other hand, can always be written as

$$\sigma_{\text{indep}}^2 = N(B_e)B_e \tag{4}$$

where B_e stands for the noise equivalent bandwidth of the receiver electronics. With $H(f)$ being the Fourier transform of $h(t)$, B_e is defined as

$$B_e = \int_0^{\infty} |H(f)|^2\, df. \tag{5}$$

The (possibly bandwidth-dependent) equivalent input noise power density is represented by $N(B_e)$. By a proper choice of K, C, and $N(B_e)$, our model can be adapted to various receiver types (cf. Table I).

In case of a *pin photodiode* as a detector, K equals S, the detector sensitivity, C is given by [11] Se (e being the elementary charge), and $N(B_e)$ represents the noise power density of the electrical amplifier N_0^2, plus some dark-current shot noise density produced by the detector. For most of this work, we assume N_0 to be independent of B_e in order to show the basic properties of this type of receiver. Especially at high data rates, however, electrical circuit noise will increase with receiver bandwidth, the impact of which we will also address in our discussions. For an *optically preamplified receiver*, consisting of an optical amplifier with power gain G followed by an optical filter to reduce the amplified spontaneous emission (ASE) power at the (pin) detector, K equals S, if we consistently define $d(t)$ as the (already amplified) signal at the photodiode.[1] The signal-dependent noise is made up of the beat noise between signal, ASE, and a (usually negligible) shot noise term, leading to [9] $C = 2N_{\text{ASE}}S^2 + Se$, with N_{ASE} being the ASE power spectral density per (polarization) mode. The noise density $N(B_e)$ is given by the sum of the ASE-ASE beat noise term [13] and N_0^2, $N(B_e) = 2S^2N_{\text{ASE}}^2(2B_o - B_e) + N_0^2$, where B_o denotes the bandwidth of the optical filter. A receiver employing an *avalanche photodiode (APD)*, characterized by a multiplication factor M and an excess noise figure F_{APD}, is accounted for by setting [11] $K = SM$, $C = SeM^2F_{\text{APD}}$, and $N(B_e)$ equal to N_0^2 plus some dark current noise terms.

III. SIMULATION METHOD

In the quasianalytical method [14] employed here, BEP estimates do not rely on the occurrence of errors but rather on the calculation of error probabilities. Given a pseudonoise (PN) bit sequence of length[2] $2^m - 1$ (which contains 2^{m-1} ones and $2^{m-1} - 1$ zeros), we evaluated the waveforms $i(t)$ and $\sigma^2(t)$ at the sampling instants $t = t_s + kT$ (T being the bit period) according to (1) and (2). (Note that owing to the signal-dependent part of $\sigma^2(t)$, ISI influences not only the signal but also—differently—the noise.) Thus, we get a total of 2^{m-1} pairs $\{i_{1,j}, \sigma_{1,j}\}$ representing the statistics of the "1" bits, and $2^{m-1} - 1$ pairs $\{i_{0,j}, \sigma_{0,j}\}$ for all "0" bits in the PN sequence. Assuming a decision threshold i_{th} and Gaussian photocurrent statistics, we then obtain BEP values for *each* bit using the complementary error function (erfc).

An estimate for the *average* BEP is found by averaging over the individual BEP values

$$\text{BEP} = \frac{1}{2^m - 1}\left\{\sum_{j=1}^{2^{m-1}} \frac{1}{2}\,\text{erfc}\left[\frac{i_{1,j} - i_{th}}{\sqrt{2}\,\sigma_{1,j}}\right] + \sum_{j=1}^{2^{m-1}-1} \frac{1}{2}\,\text{erfc}\left[\frac{i_{th} - i_{0,j})}{\sqrt{2}\,\sigma_{0,j}}\right]\right\}. \tag{6}$$

[1] The optical power at the input to the entire receiver, which is the quantity of prime technical interest, is, in this case, given by $d(t)/G$.

[2] For a reliable estimate of the average BEP, the sequence length should be long enough so that all possible combinations of bits effectively causing ISI occur at least once; if the number of bits that significantly influence a sample is n, there are 2^n different ISI combinations. In order to account for all of them, a PN sequence with $m \geq n + 1$ should be used.

Optimum sampling instant t_s and decision threshold i_{th} are adjusted by minimizing BEP.

IV. RZ Coding Gain

A. Filter Characteristics and Signal Waveforms

For the results presented here, we assumed a receiver with a fifth-order Bessel electrical filter [9]. This type of receive filter is widely used in optical receivers, even at data rates in the multi-Gb/s regime, because it has only little overshoot.

The optical power waveform for a single one-bit $p(t)$ is specified within the time interval $[0,\,(1+\alpha)T_p]$ as

$$p(t) = \begin{cases} \frac{E}{2T_p}\left[1-\sin\left(\frac{\pi}{\alpha T_p}\left(\left|t-(1+\alpha)\,\frac{T_p}{2}\right|-\frac{T_p}{2}\right)\right)\right], \\ \qquad\qquad t\in\{[0,\,\alpha T_p]\vee[T_p\,,\,(1+\alpha)T_p)]\} \\ \frac{E}{T_p}, \qquad\quad t\in[\alpha T_p,\,T_p]. \end{cases} \tag{7}$$

T_p is the effective pulse duration

$$T_p = \frac{\int_{-\infty}^{\infty} p(t)\,\mathrm{d}t}{\max_t\{p(t)\}} = \frac{E}{\max_t\{p(t)\}} \tag{8}$$

and the parameter α specifies the pulse shape. Varying α from 1 to 0, the pulse changes from $\cos^2(t)$-like to rectangular. (As an example, we used pulses with $\alpha = 0.4$ to illustrate the data signal $d(t)$ in Fig. 1.) To further characterize RZ pulses, we also introduce the RZ factor D as the reciprocal of the duty cycle

$$D = T/T_p, \tag{9}$$

with T being the bit duration.

The optical signal power at the receiver, $d(t)$, is composed of individual pulses according to

$$d(t) = \sum_{k=1}^{2^m-1} a_k p(t-kT), \tag{10}$$

with $a_k \in \{0,\,1\}$ representing the PN bit sequence. Note that the summation of optical *power* waveforms instead of optical fields within our model is fully justified by the fact that, for $D \geq 2$, no temporal pulse overlap (and thus no interference effects) can occur. Typical NRZ waveforms can be obtained using this model for $D = 1$, owing to the particular shape of the pulse edges (cf. Fig. 1). The model does not cover the region $1 < D < 2$, which is of no practical interest anyway.

B. BEP as a Function of Receiver Bandwidth

Fig. 2 shows the simulated BEP for constant average receive power as a function of the (normalized) electrical bandwidth

$$b_e = B_e/R, \tag{11}$$

with $R = 1/T$ being the data rate. The curves are shown for RZ factors $D = 1,\,2,\,3$ and for a pulse shape characterized by $\alpha = 1$. In (a), the receiver noise [see (2)] is dominated by the signal-independent term $\sigma^2_{\mathrm{indep}}$; in (b), the signal-dependent noise term $\sigma^2_{\mathrm{dep}}(t)$ dominates; in (c), both noise contributions are of equal importance. These three situations correspond to the following:

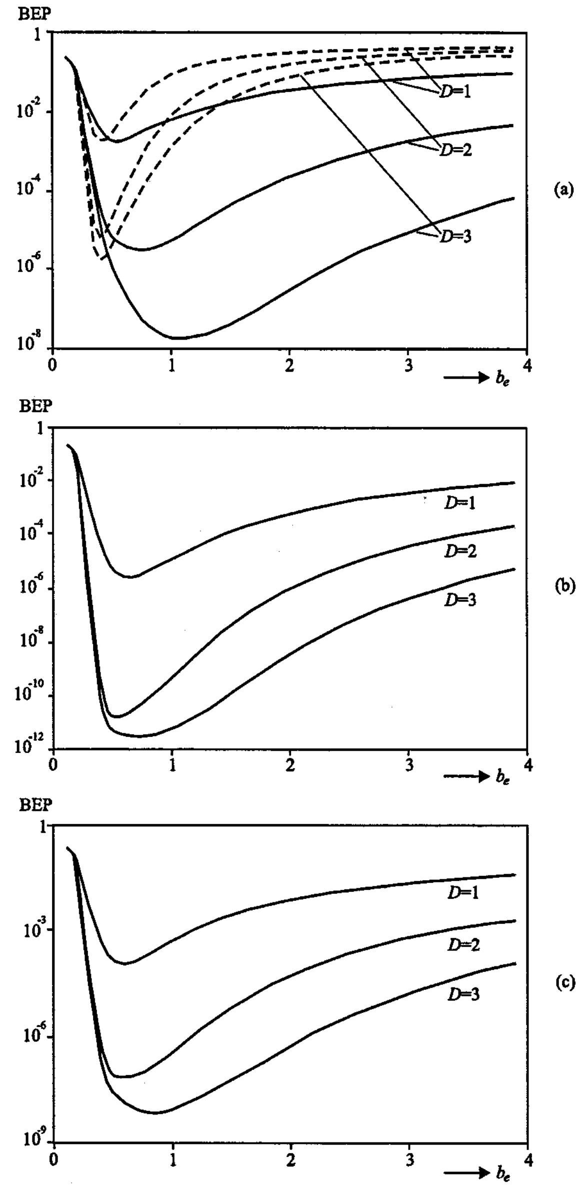

Fig. 2. BEP as a function of receiver bandwidth $b_e = B_e/R$ for (a) dominating signal-independent noise, (b) dominating signal-dependent noise, and (c) a mixture of both. The solid lines in (a) correspond to a receiver with constant noise power density, whereas the dashed lines account for the case where the noise power density increases with receiver bandwidth. An RZ factor of $D = 1$ yields an NRZ signal. $D = 3$ corresponds to a duty cycle of 33%. In all cases, the pulse shape factor is $\alpha = 1$. The pulse energy corresponds to (a) $n = 3125$, (b) $n = 72\,000$, and (c) $n = 400$ photons/bit. Further numerical values are given in Table II.

a) typical pin diode receivers, where the electrical circuit noise will always far exceed the signal dependent shot noise term.
b) optically preamplified receivers, where the signal–ASE beat noise term dominates all other noise contributions.
c) receivers employing an APD, where the signal-dependent multiplication noise is on the order of the electrical circuit noise plus dark current shot noise.

The relevant receiver parameters are summarized in Table II, together with the equivalent simulation parameters K, C, and $N(B_e)$.

TABLE II
RECEIVER PARAMETERS UNDERLYING THE SIMULATIONS OF FIGS. 2 AND 3

pin receiver (a)	optically preamplified receiver (b)	APD receiver (c)
$R = 10Gbit/s$	$R = 10Gbit/s$	$R = 2.5Gbit/s$
	$G = 30dB$	$M = 20$
	$F = 3.5dB$	$F_{APD} = 7$
	$B_0 = 0.4nm(@1.55\mu m)$	
	2 polarization modes	
$S = 0.8A/W$	$S = 0.8A/W$	$S = 1A/W$
$i_d = 10nA$ (negligible)	$i_d = 10nA$ (negligible)	$M \cdot i_d = 200nA$
$N_0 = 12pA/\sqrt{Hz}$	$N_0 = 12pA/\sqrt{Hz}$	$N_0 = 12pA/\sqrt{Hz}$
$K = 0.8A/W$	$K = 0.8A/W$	$K = 20A/W$
$C = 1.28 \cdot 10^{-19} A^2 s/W$	$C = 1.84 \cdot 10^{-16} A^2 s/W$	$C = 4.48 \cdot 10^{-16} A^2 s/W$
$N(B_e) = 1.44 \cdot 10^{-22} A^2/Hz$	$N(B_e) = 2.50 \cdot 10^{-21} A^2/Hz$	$N(B_e) = 1.53 \cdot 10^{-22} A^2/Hz$

For all three cases, optimum values for b_e exist. For $D = 1$, these optimum values represent a compromise between signal degradation due to ISI at low b_e and noise enhancement at high b_e. As expected, the optimum electrical bandwidth for NRZ reception is in the range of 0.6 times the data rate.[3]

By comparing our results for a PN data sequence with those for a single optical pulse (i.e., without ISI effects), we found that the optimum receiver bandwidth for $D \geq 2$ and dominating signal-independent noise is *not* determined by ISI, but rather by signal energy reduction caused by too narrow electrical filtering. (For RZ coding, there is no noticeable ISI degradation for electrical bandwidths above 0.5 R.)

As can be seen from Fig. 2(a), the optimum bandwidth markedly depends on the RZ factor for the case of dominating signal-independent noise if the noise power density of the receiver electronics is assumed to be constant (solid lines). In contrast, if the noise power density considerably increases with receiver bandwidth, the optimum bandwidth becomes almost independent of the RZ factor, as evidenced by the dashed lines in Fig. 2(a), where $N(B_e) = 0.5 \cdot 10^{-22} A^2 s + 4.3 \cdot 10^{-42} A^2 s^3 B_e^2$ was used.[4] If the noise is dominated by the signal dependent contribution [see Fig. 2(b)], the optimum receiver bandwidth is, too, almost the same for all RZ factors.

To give an explanation, we first consider the electrical signal power $i^2(t)$ due to a single optical pulse at the output of the electrical filter (i.e., without the influence of ISI, which is justified for RZ if $b_e \gtrsim 0.5$). At a receiver bandwidth b_e considerably smaller than the spectral width of $p(t)$, the optical pulse can be approximated by the Dirac impulse $E\delta(t)$, resulting in a current [cf. (1)]

$$i(t) \approx KEh(t). \quad (12)$$

Because the height of $h(t)$ is proportional to b_e, the electrical signal power $i^2(t)$ is proportional to b_e^2 in the low bandwidth regime. It is further *independent* of D, as E is considered constant. For dominating signal-independent noise and *constant* noise power density N_0^2, the noise variance (4) depends linearly on b_e, and, thus, system performance improves as long as increasing b_e still leads to a superlinear growth of $i^2(t)$. Any further enlargement of b_e, then, results in a BEP degradation due to noise enhancement without accompanying increase of signal power. Thus, the optimum b_e depends on the spectral width of $p(t)$ (and, hence, on the RZ factor). As long as the receiver noise density does not increase significantly with b_e, increasing D (and adjusting b_e accordingly) steadily improves receiver performance.

If the receiver is dominated by signal-dependent noise, the noise variance, like the signal power, is proportional to b_e^2 at low receiver bandwidths [cf. (4)]. Thus, neglecting ISI, system performance would improve steadily with decreasing b_e, reaching the *same* low-bandwidth limit for all D. Due to ISI, however, the BEP dramatically starts to increase for $b_e \lesssim 0.5$, always yielding an optimum receiver bandwidth in the range of $b_e \approx 0.5$. The same arguments apply to a bandwidth dependent noise power density in a receiver dominated by signal independent noise.

The mixed case, where neither the signal-independent, nor the signal dependent noise contribution clearly dominates [Fig. 2(c)], results in a BEP characteristic very similar to that found for the dominating signal-independent case, as can be seen by comparing with Fig. 2(a).

C. BEP as a Function of Input Power

For *all* cases shown in Fig. 2—no matter which noise contribution dominates or whether receiver noise depends on the electrical bandwidth or not—RZ coding leads to lower BEP than NRZ at any receiver bandwidth b_e. In this subsection, we assume b_e to be fixed at $0.6R$. Fig. 3 shows the BEP as a function of the average received optical power (expressed in terms of an average number of photons/bit at the detector, n) for var-

[3]For dominating signal-independent noise a), our simulations yield an optimum NRZ bandwidth of $0.54R$. For dominating signal-dependent noise b), we find $0.63R$. For no clear noise class dominating c), we have 0.57 R.

[4]The excess noise power of a bipolar transistor amplifier is proportional to b_e^2 [7], [8], and the excess noise power of a transimpedance amplifier is proportional to b_e^3 [15].

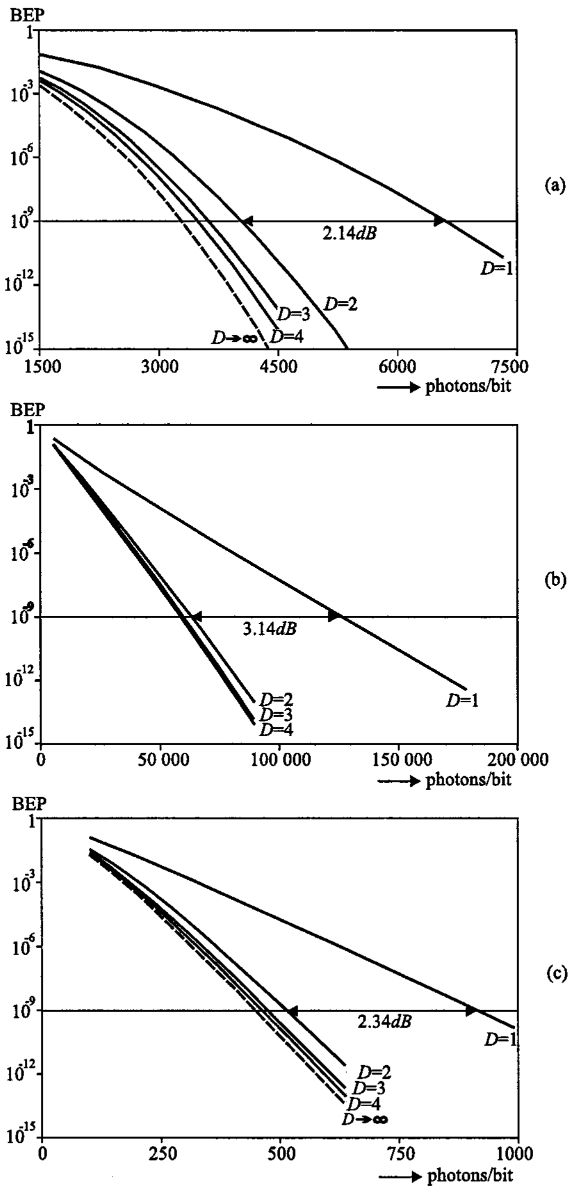

Fig. 3. BEP as a function of average number of photons/bit at the detector for (a) dominating signal-independent noise, (b) dominating signal-dependent noise, and (c) a mixture of both. Note that for the optically preamplified case (b), the number of photons at the receiver input is less than the displayed number at the detector by the optical amplifier gain ($G = 1000$). The pulse shape factor $\alpha = 1$, the receiver bandwidth equals $b_e = 0.6$. Numerical values for the simulation are given in Table II.

ious RZ factors. Again, Fig. 3(a) gives the results for dominating signal-independent noise, (b) for dominating signal-dependent noise,[5] and (c) for a mixture of both noise terms, corresponding to the three receiver types according to Table I.

The marked *qualitative* difference in the behavior of the curves in Fig. 3(a) and (b) can be explained on an SNR basis neglecting ISI. For reasonably high SNR, the BEP can be excellently approximated by the improved Chernoff bound [12]

$$\mathrm{BEP} \approx \frac{1}{Q\sqrt{2\pi}} \exp(-Q^2/2) \tag{13}$$

[5]Note that for the optically preamplified case, the number of photons at the receiver input is less than the displayed number at the detector by the optical amplifier gain ($G = 1000$).

with the parameter Q defined as

$$Q = \frac{i_1 - i_0}{\sigma_1 + \sigma_0}. \tag{14}$$

The indexes 0 and 1 denote the values of the mean signal and its standard deviation for a logical zero and one, respectively. For perfect extinction ratio ($i_0 = 0$) and signal-independent noise ($\sigma_1 = \sigma_0$), we have $Q = \sqrt{\mathrm{SNR}}/2$ and, because for this case SNR $\propto n^2$, we obtain $\log(\mathrm{BEP}) \propto n^2$, which represents a parabola [cf. Fig. 3(a)]. For dominating signal-dependent noise, on the other hand, SNR $\propto n$, and we arrive at $\log(\mathrm{BEP}) \propto n$, i.e., at a straight line [cf. Fig. 3(b)].

Comparing the number of photons necessary to achieve a certain BEP (e.g., 10^{-9}), we notice a clear sensitivity enhancement of RZ coding over NRZ. For dominating signal-independent noise (a), we find a sensitivity gain of 2.14 dB when going from NRZ to RZ with $D = 2$. Further increasing the RZ factor yields only moderate additional gain. The asymptotic gain potential with $D \to \infty$ is 3.01 dB. The fact that (for fixed b_e) an asymptotic sensitivity gain exists for $D \to \infty$, and that moderate RZ factors suffice to nearly exhaust that gain can be readily explained both in the frequency domain and in the time domain. Increasing the RZ factor broadens the spectral width of the RZ pulses $p(t)$, while the low-frequency part of $P(f)$ [the Fourier transform of $p(t)$] remains clamped to E, owing to the constant average optical power constraint. The change of $P(f)$ over the fixed band within which $H(f)$ is significant, therefore, becomes less as a larger D is chosen. [For $D = 4$, $P(f)$ already has a spectral width of about $2R$, which is significantly broader than the typical NRZ bandwidth of $h(t)$.] Thus, the spectrum of the mean photocurrent [which is the multiplication of $P(f)$ with $H(f)$, cf. equation (1)] will be more or less independent of D for RZ factors higher than about 4. In the temporal domain, the explanation runs as follows. For D high enough to let $p(t)$ look like a Dirac impulse when compared to $h(t)$, the convolution (1) reproduces the impulse response of the electronics, $h(t)$, independent of D [cf. (12)]. As this behavior is reached at fairly moderate RZ factors, one does not see significant changes in BEP as D is increased beyond about 4.

For dominating signal-dependent noise (b), a sensitivity gain of 3.14 dB is found when going from NRZ to RZ with $D = 2$, with a total gain of some 3.53 dB for $D = 4$. This amount of RZ gain also represents the overall RZ improvement potential for high D, if the optical bandwidth is increased accordingly; however, for fixed optical bandwidth, the sensitivity starts to deteriorate for higher D due to spectral truncation of the increasingly broader RZ pulses, as discussed in [16]. Unlike that predicted in an SNR-based analysis [9], the observed gain is *higher* for signal-dependent noise than it is for signal-independent noise. It should be noted at this point that the RZ gain achievable in an optically preamplified receiver mainly depends on how well an optical matched filter [17], [18] could be realized (and was practical) for NRZ. If such a filter was employed, no sensitivity enhancement could be achieved by RZ coding, as demonstrated in [10].

If both noise contributions are of considerable influence [see Fig. 3(c)], the achievable sensitivity gain lies between the two

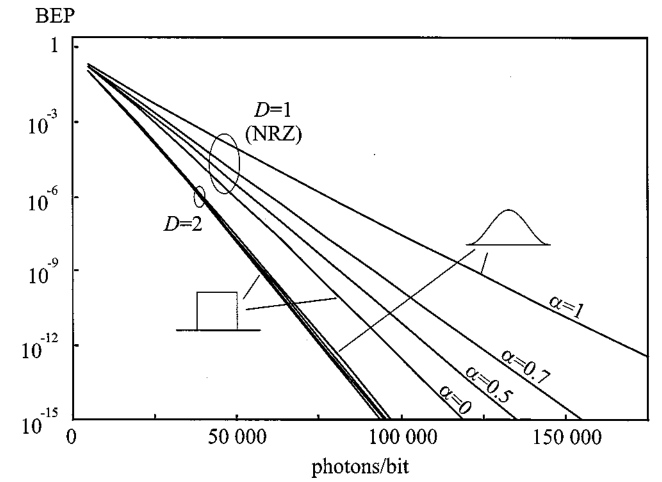

Fig. 4. BEP as a function of average number of photons/bit for dominating signal-dependent noise [as in Table II, column (b)]. The pulse shapes, parameterized by $\alpha = 0, 0.5, 0.75, 1$ range from rectangular to $\cos^2$-like; their influence on NRZ transmission is significantly stronger than for RZ coding.

limiting cases. For RZ coding with $D = 2$, we find a sensitivity improvement of 2.34 dB.

V. Influence of Different Pulse Shapes on BEP

As explained previously, it is the spectral width of the optical pulses in relation to the bandwidth of the receive filter that determines receiver sensitivity and, thus, the RZ gain, as well. In case the spectral width of $p(t)$ is low (i.e., primarily for NRZ), we additionally have to expect a considerable influence of the *pulse shape* on the sensitivity. To illustrate the shape-dependent behavior of the RZ sensitivity gain, we evaluated BEP for the case of dominating signal-dependent noise [setup (b) of Table I] for a normalized receiver bandwidth of $b_e = 0.6$ and various pulse shapes (7), parameterized by $\alpha = 0, 0.5, 0.7, 1$. The results are shown in Fig. 4 as a function of the average number of photons/bit at the detector. Owing to their wider spectra, the rectangular pulses yield the lowest BEPs, both for $D = 1$ and $D = 2$. Differences in BEP with varying α are most pronounced for NRZ because, in this case, the spectrum $P(f)$ is mainly concentrated in the interval $[0, 0.5R]$, i.e., well within the pass-band of $H(f)$; thus, spectral changes brought by different values of α have the strongest impact on the convolution (1) for NRZ. Impulsive coding with $D = 2$ essentially doubles the spectral width of $p(t)$; no significant changes of the pulse spectrum occur within $[0, 0.5R]$ when varying α, and receiver performance becomes less dependent on the particular pulse shape used. Therefore, the RZ sensitivity gain becomes less if NRZ is already realized with comparatively wide-band rectangular pulses.

VI. Influence of the Extinction Ratio

Up to this point, we have assumed pulse sources with perfect extinction ratio, i.e., vanishing optical power for logical zeros. When considering practical systems with finite extinction ratios, we may distinguish between two options for the generation of RZ signals. The first possibility corresponds to a

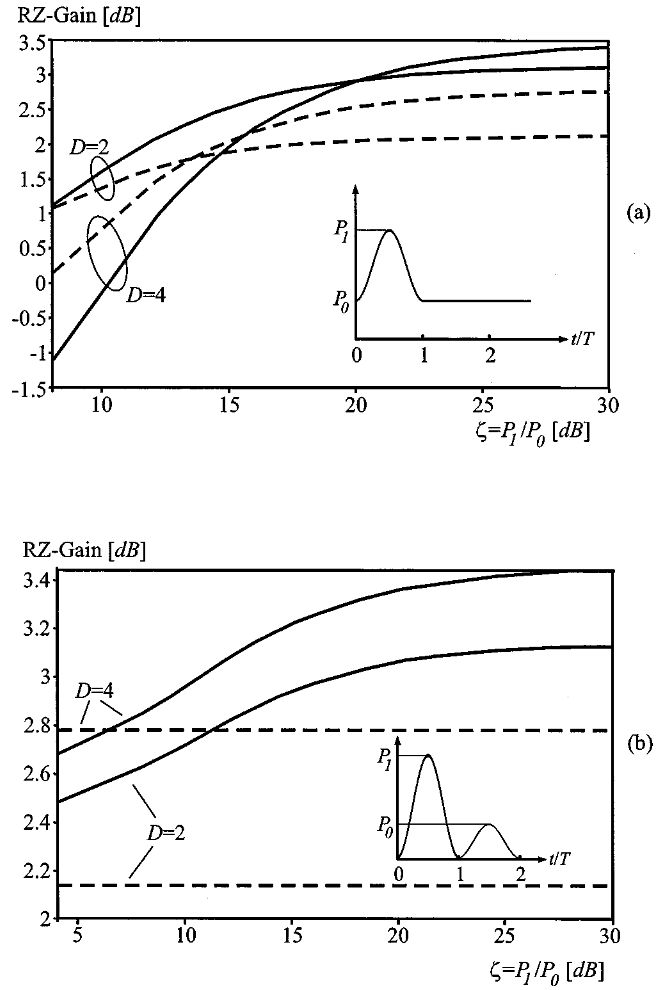

Fig. 5. RZ sensitivity gain as a function of extinction ratio for (a) direct modulation RZ signals and (b) pulse train modulation. The dashed lines represent a receiver with dominating signal-independent noise. The solid lines are valid for dominating signal-dependent noise. Note the different ordinate scales in (a) and (b).

continuous-wave (cw) laser with subsequent external RZ intensity modulation or, equivalently, to a laser biased slightly above threshold that is directly modulated by the RZ signal. For logical zeros, thus, we obtain a temporally constant power level P_0 at the receiver. The second possibility is based on a primary source (e.g., a mode-locked laser or a cw-fed sinusoidally driven nonlinear modulator) producing a continuous train of optical pulses. These are subsequently intensity modulated using the NRZ data signal. As the second data modulator will not completely suppress the pulses of the primary source at a logical zero, a sequence of strong and weak pulses will appear at the transmitter output. In this case, we define P_0 as the peak power of the zero-bit optical pulses at the receiver. With P_1 being the received peak power of the one-bit pulses, we define the extinction ratio ζ as

$$\zeta = P_1/P_0. \quad (15)$$

Fig. 5 shows the sensitivity gain of RZ coding (both $D = 2$ and $D = 4$) over NRZ for $\alpha = 1$ as a function of ζ. The dashed curves represent a receiver with dominating signal-independent noise, whereas the solid lines apply for dominating signal-dependent noise. [The underlying parameters are those from Table II, columns (a) and (b), together with a receiver bandwidth of $b_e = 0.6$.] As indicated by the figure insets, Fig. 5(a) shows the results for the direct modulation option, and Fig. 5(b) shows the results for pulse train modulation.

In the case of *direct modulation* [cf. Fig. 5(a)], the performance of RZ-coded systems dramatically depends on ζ, making an extinction ratio as high as 20 dB necessary to exploit the potential RZ sensitivity gain. For small ζ, impulsive coding may even perform worse than NRZ [cf. RZ coding with $D = 4$ and dominating signal-dependent noise in Fig. 5(a)]; the higher the RZ factor, the worse will be the performance at low ζ, leading to intersections of the RZ gain for $D = 2$ and $D = 4$. The physical reason for this behavior can readily be explained by considering how the electrical eye is formed. The optical power for a "1" bit and for a "0" bit is, respectively, given by

$$p_1(t) = (P_1 - P_0)\tilde{p}(t) + P_0 \tag{16}$$

and

$$p_0(t) = P_0 \tag{17}$$

where $\tilde{p}(t)$ denotes the pulse (7) normalized to unit amplitude. Using (1), we find, for the electrical eye

$$i_1 - i_0 = K(p_1 * h)(t_s) - KP_0 = KP_1 A(D)(1 - 1/\zeta) \tag{18}$$

where $A(D)$ accounts for the attenuating influence of $h(t)$ on $\tilde{p}(t)$; for fixed b_e, $A(D)$ decreases monotonically with D. Thus, the *peak power* P_1 required for a given eye opening increases proportional to $1/A(D)$ for RZ coding as compared to NRZ. To arrive at a meaningful RZ gain, however, we have to compare *average* power levels. For direct RZ modulation the average power is given by

$$\overline{P} = P_1 \left(\frac{1}{\zeta} + \left(1 - \frac{1}{\zeta}\right) \frac{1}{2D} \right) \tag{19}$$

which, in the ideal limit $\zeta \to \infty$, approaches $P_1/2D$, and we arrive at

$$\frac{\overline{P}_{\mathrm{NRZ}}}{\overline{P}_D} \geq DA(D) \tag{20}$$

for the ratio of average power levels required to achieve a certain eye opening. Equation (20) represents a *gain* of RZ over NRZ, because $A(D)$ decreases[6] slower than $1/D$. For poor extinction ratios approaching the limit $\zeta \to 1$, we find from (19) $\overline{P} \to P_1$, independent of D; this leads to the limiting form

$$A(D) \leq \frac{\overline{P}_{\mathrm{NRZ}}}{\overline{P}_D} < 1 \tag{21}$$

for low extinction ratios where impulsive coding degrades receiver sensitivity. Because $A(D)$ decreases monotonically with D, performance gets worse the higher an RZ factor is chosen. This, in turn, indicates that an intersection of RZ gain curves for different RZ factors has to be expected, i.e., that RZ coding with *lower* RZ factor is more advantageous if a certain extinction ratio requirement cannot be met. For the parameters underlying Fig. 5 ($\alpha = 1$, $b_e = 0.6$), we have $A(2) = 0.66$ and $A(4) = 0.38$; using (18) and (19), this leads to a predicted intersection at $\zeta = 13.3$ dB, which corresponds well with that found by our simulations for dominating signal-independent noise [dashed curves in Fig. 5(a)]. In the case of dominating signal-dependent noise (solid lines), the intersection shifts to higher values of ζ. The reason is an additional, signal-dependent "0"-bit noise term due to the constant power level P_0, which is not attenuated by $h(t)$.

For *modulated pulse-train* RZ signals, Fig. 5(b) shows that finite extinction ratios are of no influence on the RZ sensitivity gain at all if signal-independent noise dominates (dashed lines). The *absolute* sensitivity, of course, decreases for finite extinction ratios, but—unlike for the direct modulation option—by the same amount for NRZ and RZ signals. For dominating signal-dependent noise (solid lines), we find a rather moderate reduction of the RZ gain as ζ decreases, such that impulsive coding will always improve system sensitivity compared to NRZ.

This feature can also be explained looking at the electrical eye; "1"-bit pulses and "0"-bit pulses are equally attenuated by the electrical filter $h(t)$, leading to exactly the same expression for the electrical eye opening that was found in the previous case [see (18)]. The difference for the modulated pulse-train option is the calculation of the average power, which we find to be

$$\overline{P} = P_1(1 + 1/\zeta)/2D. \tag{22}$$

Not surprisingly, the RZ gain in the limit of $\zeta \to \infty$ is equal to that for the direct modulation option, as seen in (20). But, as is obvious from (22), the ratio of the average powers $\overline{P}_{\mathrm{NRZ}}$ and $\overline{P}_D$ is now *independent* of the extinction ratio, resulting in a constant RZ gain, at least for dominating signal independent noise. The moderate decrease of RZ gain at low extinction ratios for dominating signal-dependent noise [solid lines in Fig. 5(b)] can again be attributed to a signal-dependent noise contribution due to the pulse $p_0(t)$ for logical zeros, which, in contrast to the direct modulation option, is affected by the attenuating influence of $h(t)$. The weak dependence on ζ indicates that RZ and NRZ coded systems suffer from "0"-bit noise in roughly the same way.

From the results of Fig. 5(b), it becomes apparent that, in the case of pulse train modulation, the extinction ratio is of little influence. Generating RZ signals this way will, thus, from an extinction-ratio point of view, be the preferred modulation technique for practical RZ coded systems.

VII. Conclusion

Using semianalytical simulations, we showed that the BEP of optical direct detection receivers can be reduced if RZ coding is employed instead of NRZ and if the average receive power is kept constant, even if the receiver is optimized for the (more narrow-band) NRZ reception. We pointed out that the RZ-coding gain considerably depends on whether the dominating noise term is signal-dependent, as is the case in a well-designed optically preamplified receiver, or signal-independent, as in a pin receiver. For dominating signal-independent

[6]For $D \to \infty$, i.e., $p(t)$ can be approximated by a Dirac pulse and $(p * h)(t)$ gives the impulse response of the filter, $A(D)$ is proportional to $1/D$; (20) monotonically approaches an asymptotic limit as D is increased.

noise, we find a typical RZ gain of 2.14 dB when using RZ pulses of half the bit duration, with an asymptotic potential of 3 dB that is nearly exhausted for RZ pulses of 1/4 times the bit duration. For dominating signal-dependent noise, an RZ gain of about 3.14 dB can be expected. Depending on the optical filter used for NRZ reception, the latter receiver may offer a higher RZ gain potential.

A quantitative discussion of the influence of the optical pulse shape showed that for NRZ coding, more broad-band pulses (e.g., rectangular) lead to a better performance than smoother ones (e.g., $\cos^2$-like); this difference was not found for RZ coding.

An analysis of the influence of finite extinction ratios on the achievable RZ coding gain revealed that, for direct RZ-modulation, an extinction ratio of about 20 dB is required to still have significant RZ gain. If, on the other hand, the RZ signal is generated by modulating a pulse train, the extinction ratio of the data modulator is not a limiting factor. An optical pulse source modulated externally by an intensity modulator, therefore, suggests itself for practical implementations.

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