Performance optimization of optically preamplified receivers for return-to-zero and non return-to-zero coding

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Dedicated to Prof. Walter R. Leeb on the occasion of his 60th birthday

Abstract **We determine optimum bandwidths for optical and electrical filters in optically preamplified receivers, both for NRZ coding and RZ coding. Our simulations clearly reveal the trade-offs to be made when optimizing bandwidths: NRZ is typically limited by intersymbol interference, while RZ is limited by energy truncation. Thus, RZ allows for tighter filtering, leading to near quantum limited performance. Further, RZ systems are less susceptible to suboptimum filtering. We also show that the use of RZ with duty cycles below 33% only leads to minor additional receiver sensitivity improvements at the expense of impractically higher receiver bandwidths. Employing an RZ duty cycle of 33%, we achieved a receiver sensitivity of 52 ppb at a data rate of 10 Gbit/s, which is only 1.4 dB off the quantum limit.**

Keywords **direct detection receivers, optical preamplification, RZ, NRZ, optimum filter bandwidths.**

1. Introduction

The design of highly sensitive optical receivers is of major interest for optical communication systems. This is especially true for optical intersatellite links [1, 2, 3, 4, 5], characterized by the inherent impossibility of deploying optical in-line amplification. Since in these systems the available transmit power as well as the mass and size of the optical terminals (including its optical antennas) are tightly bounded design parameters due to technological restrictions, optimizing receiver sensitivity, i.e. minimizing the optical receiver input power required to achieve a certain bit error ratio (BER) becomes a prime goal [6, 7, 8].

The class of receivers one may think of first when high receiver sensitivity is desired is probably coherent reception, i.e. optoelectronic detection of the signal beating with the field of a local oscillator laser (see, e.g., [9]). Indeed, the best receiver sensitivity involving classical receiver concepts is realized by the most sophisticated representative of this class, the optical homodyne receiver employing phase shift keying (PSK). It has a theoretical sensitivity limit of 9 photons/bit (ppb) at a bit error ratio (BER) of 10^{-9}, with technically achieved 20, 72, and 101 ppb at data rates of 565 Mbit/s [10], 4 Gbit/s [11], and 10 Gbit/s [12]. With this performance, homodyne PSK receivers outperform all other (coherent as well as direct-detection (DD)) receivers by at least 3 dB, however, at the cost of exceedingly high technological complexity. Optically preamplified DD receivers, on the other hand, have been shown [13] to yield the same theoretical sensitivity as heterodyne receivers. Recent technological advances in the wavelength range around 1550 nm (as widely deployed in terrestrial fiber communication systems) have led to experimental demonstrations of optically preamplified receiver sensitivities approaching the quantum limit to within 0.5 dB (i.e. 43 ppb) at 5 Gbit/s for intensity modulation (on/off keying, OOK) [7], and to within 2 dB (i.e. 30 ppb) at 10 Gbit/s for differential phase-shift keying (DPSK) [14]. These values outperform all comparable results achieved by coherent reception so far, and thus make optically preamplified receivers a serious contender to coherent receivers for free-space optical communication systems [15, 16]. An analysis of commercial availability of components, as well as on their robustness, mass, size, and cost, further emphasizes the superiority of optically preamplified receivers for space-borne communication links [17, 18], especially when using the return-to-zero (RZ) data format rather than the more conventional non return-to-zero (NRZ) format.

The optimum optically preamplified receiver uses a matched optical filter (i.e. an optical filter with a transmission spectrum equal to the optical pulse spectrum) and no postdetection electrical filtering [19, 7]. However, since the convenience of an inherently narrow-band matched optical filter is doubtful because of wavelength stability reasons, and since electrical filtering cannot be avoided due to fundamental bandwidth limitations of the receiver electronics, one rather has to optimize optical and electrical bandwidths for given, technologically available filter characteristics [20].

In this paper, we present both numerical simulations and experimental results for the design of optically preamplified direct detection receivers, both for intensity modulated NRZ and RZ. We investigate the dependence of receiver sensitivity on the signal's pulse shape as well as on the receiver bandwidths in the optical and electrical domain. Our simulations result in design guidelines leading to near quantum limited performance: Using solely off-the-shelf commercially available components, we experimentally demonstrate record receiver sensitivities of 1.4 dB off the quantum limit at a data rate of 10 Gbit/s.

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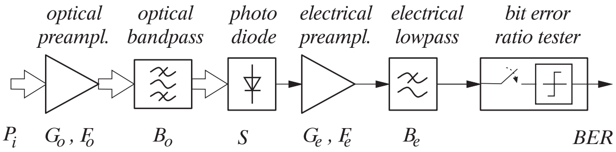

Fig. 1. System setup.

2. Receiver setup

Figure 1 shows the structure of the optically preamplified receiver underlying our simulations and measurements. The intensity modulated optical signal at the receiver input, both in RZ and NRZ format, carries the information of a pseudo random bit sequence (PRBS) of length $2^7 - 1$. In our simulations, the optical pulses are modeled with $\cos^2$-shaped edges, as defined in [20]; using a pulse shape parameter of $\alpha = 0.8$ for NRZ and $\alpha = 1$ for RZ coded signals, we obtained good agreement with the pulse shapes measured in the experimental setup. For RZ-coding the pulse duration is a fraction d of the bit duration, $T_{pulse} = d \cdot T_{bit}$, where d is the duty cycle. We employ data rates of up to 10 Gbit/s, and an optical carrier wavelength of 1550 nm. The receiver input power P_i is defined at the input to a high-gain ($G_o = 38$ dB), low-noise (noise figure $F_o = 3.3$ dB) Erbium-doped fiber amplifier (EDFA). The amplified spontaneous emission (ASE) generated by the EDFA is modeled as a circularly symmetric, white Gaussian noise process [21]. To limit the ASE power reaching the detector, narrow-band fiber Bragg gratings of various 3 dB-bandwidths B_o are used as optical filters, represented by their complex transfer functions [22]. The insertion loss of these filters is 5 dB. The broadband pin photodiode (bandwidth 40 GHz) is followed by a broadband electrical preamplifier (bandwidth 18 GHz) to yield an overall conversion gain of 1350 V/W as well as a noise voltage density of $40\,\mathrm{nV}/\sqrt{\mathrm{Hz}}$. The transfer characteristic of the receiver electronics is set by 5th-order Bessel electrical lowpass filters of various 3 dB-bandwidths B_e. The BER is determined by feeding the signal to a bit error ratio test set (BERT).

3. Simulation method

To determine BER, we use a quasi-analytical simulation method [23]. Mean $s(t)$ and variance $\sigma^2(t)$ of the signal after electrical filtering are calculated separately using the expressions given in [24]. Regarding the noise statistics, we employ an advanced Gaussian method based on *exact* expressions for the noise variances and on the assumption of a Gaussian distribution of the photocurrent. The sensitivities obtained using this method are an excellent estimate of the *exact* values, derived taking into account the correct statistics of the photocurrent. The difference lies within 0.4 dB over a wide range of receiver parameters [25]. Since both signal shot noise and ASE shot noise are negligible in a typical optically preamplified receiver scenario, the total variance of the electrically filtered signal is the sum of the signal-ASE beat noise, the ASE-ASE beat noise, and the thermal noise of the receive electronics,

$$\sigma^2(t) = \sigma^2_{s-ASE}(t) + \sigma^2_{ASE-ASE} + \sigma^2_{therm}. \quad (1)$$

Note the time-dependence of the signal-ASE beat noise variance, expressing the non-stationary nature of this noise source.

From the signal mean and variance, the BER is calculated by averaging over the error probabilities for each bit in the PRBS, i.e.

$$BER = \frac{1}{N} \sum_{k=1}^{N} BER_k, \quad (2)$$

with BER_k being the error probability associated with the detection of one individual bit,

$$BER_k = \mathcal{Q}\left(\frac{\pm(s(kT + t_s) - s_{th})}{\sigma(kT + t_s)} \right). \quad (3)$$

In this equation, $\mathcal{Q}$ is the complementary error function

$$\mathcal{Q}(x) = \frac{1}{2\pi} \int_x^\infty e^{-\alpha^2/2} d\alpha, \quad (4)$$

s_{th} is the decision threshold, t_s is the sampling instant relative to the bit frame, and $T = 1/R$ denotes the bit duration. The "+" sign in Equ. (3) has to be used when detecting a logical '1', whereas the "−" sign applies if the bit under consideration is a '0'. Sampling instant and decision threshold are optimized by minimizing the overall BER (Equ. (2)) over s_{th} and t_s. Both in simulation and experiment, no clock recovery is used; the clock signal is assumed to be known to the receiver.

The quantity of ultimate interest when designing a high performance receiver is its sensitivity. We define the sensitivity n_s as the average number of photons per bit (ppb) necessary at the input of the optical preamplifier to achieve $BER = 10^{-9}$. Most of the results in this paper are given in terms of a sensitivity penalty γ_q relative to the quantum limit n_q,

$$\gamma_q = 10 \log(n_s/n_q) \quad [dB]. \quad (5)$$

For optically preamplified direct detection receivers employing the Gaussian approximation n_q equals 41 ppb at a BER of 10^{-9} [26, 20]. However, an exact calculation yields 38 ppb, which was used throughout this paper as a common reference, since it represents the physical limit for the measurements.

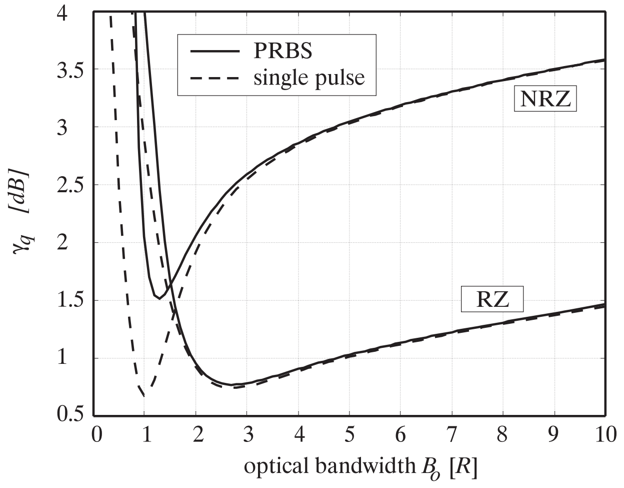

Fig. 2. Sensitivity penalty vs. optical filter bandwidth for NRZ and RZ. Solid lines represent results for a PRBS, while dashed lines stand for the single-pulse case. The electrical bandwidth is kept constant at $B_e = 0.8R$.

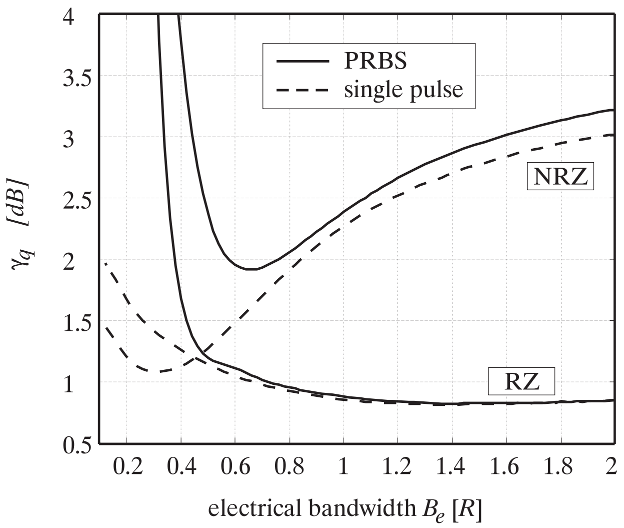

Fig. 3. Sensitivity penalty vs. electrical filter bandwidth for NRZ and RZ. Solid lines represent results for a PRBS, while the dashed lines stand for the single-pulse case. The optical bandwidth is kept constant at $B_o = 2R$.

4. Bandwidth optimization and design tradeoffs

As pointed out in the introduction, the theoretically optimum optically preamplified receiver structure – consisting of a matched optical filter and no post-detection electrical filtering – is impractical, mainly for reasons of component availability, design tolerances, and parameter drift. The main task in receiver design it thus to optimize performance for given optical pulse shapes as well as optical and electrical filter characteristics by varying the filter bandwidths and/or the pulse duration.

As discussed in [20], the main objective of the optical filter is to reduce the amount of (broadband) ASE reaching the detector, which in turn lowers the ASE-ASE beat noise[1]. Until recently, this well-established design rule led to the choice of as narrow as possible optical bandpass filters. However, the increase of data rates to ranges beyond 10 Gbit/s combined with the availability of optical filters with passbands on the order of 10 GHz has led to technically well realizable situations in which narrowband optical filters start to deteriorate receiver performance. For RZ, narrowband optical filters mainly reduce the signal energy, while for NRZ they significantly increase inter-symbol interference (ISI).

Figure 2 illustrates the trade-offs to be made when optimizing the *optical* filter bandwidth: The receiver sensitivity penalty γ_q is given as a function of the optical bandwidth B_o for both NRZ-coding and RZ-coding with 33% duty cycle ($d = 0.33$). The electrical filter bandwidth is kept constant at $B_e = 0.8R$. The dashed curves apply to the ISI-free single-pulse case, which is obtained by sampling only a single, isolated "1"-bit and a single, isolated "0"-bit. Decreasing the optical bandwidth from $10R$ initially improves receiver sensitivity, both for RZ and NRZ, due to reduced ASE-ASE beat noise. If chosen below $2.7R$ for RZ (below $1R$ for NRZ), the penalty increases significantly. Since we are considering the single-pulse case, this increase can be attributed solely to energy truncation brought by the optical filter [20, 22]. To quantitatively assess the influence of ISI on system performance, we now compare the single-pulse case to the results for a PRBS (solid lines in Figure 2). At high optical bandwidths, the ISI-free curves are identical to the curves for the PRBS, indicating the absence of ISI. Moving towards smaller bandwidths, the NRZ curves start to separate at a bandwidth of about $B_o = 3R$, which is clearly attributable to ISI only. For NRZ, the optimum optical bandwidth in the presence of ISI ($B_o = 1.3R$) is higher than for the ISI-free case ($B_o = 1R$), which shows that a trade-off between ISI and ASE-ASE beat noise has to be made for NRZ. For RZ-coding, on the other hand, there is almost no difference between the results for the PRBS and the single-pulse case for $B_o \geq 1.5R$, leading us to the conclusion that ISI plays a minor role in optimizing the optical bandwidth for RZ. In this case, it is rather the energy reduction accompanying too narrow optical filtering that has to be compromised with noise; RZ pulses with a duty cycle of 33% roughly correspond to a spectral width of $3R$, which is close to the optimum optical bandwidth of $B_o = 2.7R$ (for both PRBS and single pulse case).

[1] To a good approximation [27], the ASE-ASE beat noise is linearly proportional to the optical bandwidth, while the signal-ASE beat noise is independent of the optical bandwidth.

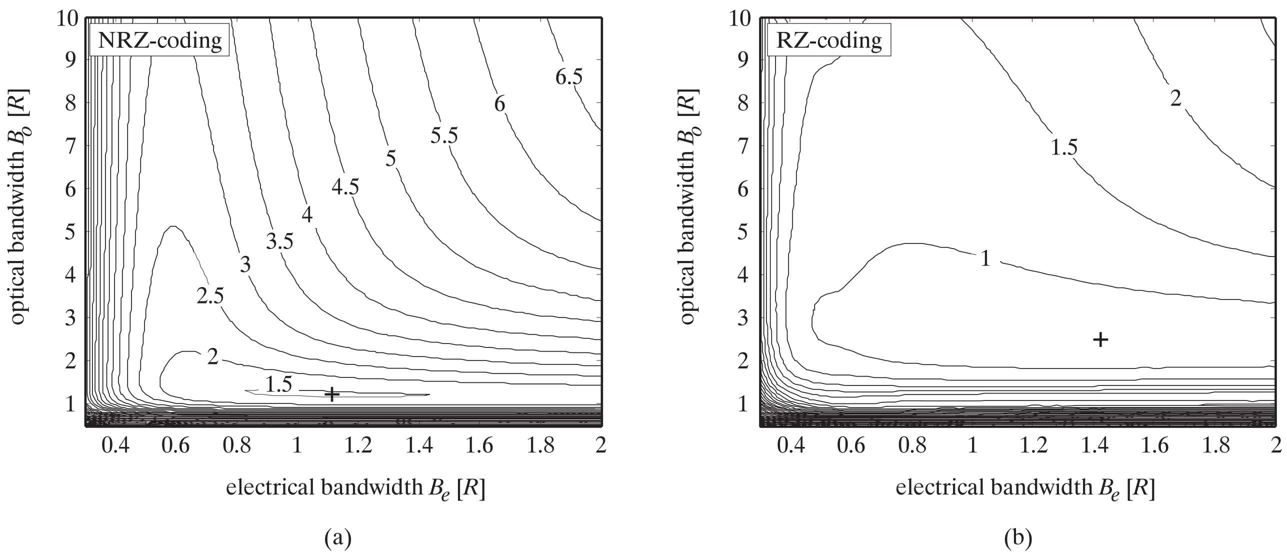

Fig. 4. Receiver sensitivity penalty γ_q relative to the quantum limit as a function of optical and electrical bandwidth, B_o and B_e for NRZ-coding (a) and RZ-coding (b). The contour lines are separated by $0.5dB$.

To better understand the trade-offs to be made when optimizing the *electrical* filter bandwidth, Fig. 3 shows the dependence of the sensitivity penalty γ_q on B_e for constant optical bandwidth ($B_o = 2R$). As in Fig. 2, the dashed curves apply for the single-pulse case, while the solid lines take into account ISI. The difference between the ISI-free case and the PRBS case for NRZ at high B_e reflects the ISI brought by the optical filter and should be compared to the ISI penalty for NRZ at $B_o = 2R$ in Fig. 2. In the case of NRZ, decreasing B_e from $2R$ improves sensitivity due to reduced detection noise, until at $B_e = 0.65R$ the sensitivity starts to degrade because of ISI, marked by the beginning separation of the dashed and solid NRZ-curves at $B_e = 0.8R$. At $B_e = 0.3R$, the ISI-free curve also starts to degrade, which can be attributed to the signal-independent noise terms (ASE-ASE beat noise and thermal noise) overtaking the signal-dependent noise term (signal-ASE beat noise) [28]. For RZ, ISI is seen to set in only at $B_e = 0.5R$, which is significantly lower than the optimum value ($B_e = 1.5R$). Thus, as in the case of the optical bandwidth, the receiver sensitivity for NRZ coding is limited by ISI, while for RZ, the optimum receiver performance is found by trading detection noise against pulse energy reduction due to too narrowband electrical filtering.

Note also that ISI is the main reason, why RZ outperforms NRZ by 1-2 dB at optimized bandwidths; for the ISI-free case, the optimum NRZ performance is much closer to that of RZ. This finding is specific to receivers limited by signal-dependent noise, and contrasts the behavior of receivers limited by signal-independent noise, since the latter show significant RZ gain even for the ISI-free case [25].

Having shed some light on the trade-offs to be made when optimizing an optically preamplified receiver, the question arises how the optimum optical and electrical filter bandwidths depend on each other, and what the tolerances to suboptimum bandwidth choices are. In other words, it is interesting to know the optimum optical bandwidth, given some fixed electrical bandwidth, and vice versa. Figure 4 displays the sensitivity penalty γ_q vs. both B_o and B_e for NRZ (a) and RZ ($d = 0.33$) (b). The contour lines (lines of constant γ_q) are separated by 0.5 dB in both plots. The plots illustratively show the optimum bandwidth constellations. They also provide an overview of system tolerances regarding bandwidth variations, which are considerably different for NRZ and RZ: In general, RZ is much more tolerant to suboptimum receiver bandwidths than NRZ. Further, both for RZ and NRZ, choosing bandwidths larger than the optimum has less consequences on system performance than excessive narrowband filtering, as is commonly known [20, 29, 30, 31, 32, 33]. For NRZ coding, Fig. 4 (a) shows that the choice of the electrical bandwidth B_e becomes more critical if broadband, suboptimum optical filters are used. In this case, the optical filter is of little influence, regarding both signal distortion (ISI) and ASE noise reduction. Under these circumstances, B_e is the only parameter to affect ISI and, consequently, has to be chosen carefully. Using an electrical filter of about $B_e = 0.6R$, the influence of B_o on receiver performance is rather weak.

Note that over a wide range of electrical bandwidths (from $0.5R$ to $2R$) the optimum optical bandwidth is quite independent of the actual B_e employed. However, to maximize system performance, the electrical bandwidth B_e has to be significantly increased as B_o approaches its optimum. This owes to the fact

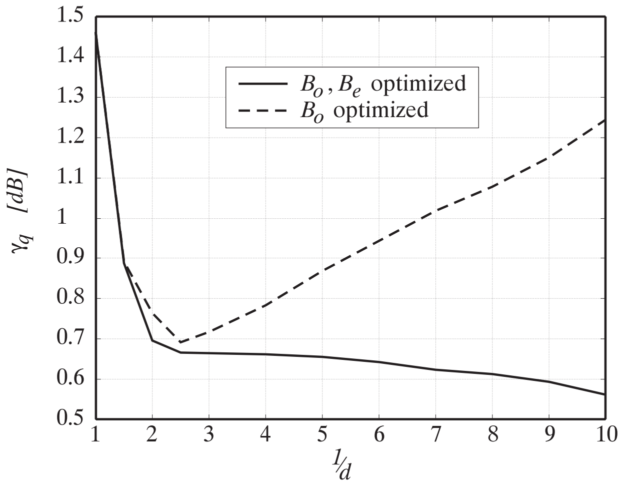

Fig. 5. Sensitivity penalty vs. RZ duty cycle d. The solid curve represents the results for optimized optical and the electrical filter bandwidths, while the dashed curve applies for fixed electrical filter bandwidth, kept at $B_e = 1R$, the optimum value for NRZ ($d = 1$).

that detection noise due to ASE is already reduced to a large extent by optical filtering. Since the beating of signal-ASE and ASE-ASE are by far the dominating noise sources in our receiver, any further reduction of noise power by means of electrical filtering is of little effect on performance. Reducing B_e will rather introduce unwanted degradations due to ISI. Thus, the electrical bandwidth primarily has to be chosen to avoid ISI (and *not* for noise reduction). As a consequence, the influence of B_e is less pronounced with optimum optical filtering, provided that B_e falls within a range of about $0.8R$ up to $2R$. To put it another way, filtering in the optical domain is more effective than filtering in the electrical domain, as this reduces ASE bandwidth *and* ASE power, whereas electrical filtering only acts on the receiver noise bandwidth [30].

5. Choice of RZ duty cycle

In the previous section, a duty cycle of $d = 0.33$ was used for all simulations involving RZ-coding. Here, we investigate the influence of the RZ duty cycle on receiver sensitivity. This duty cycle dependence of the RZ gain is particularly interesting for optically preamplified receivers, where – in contrast to receivers limited by signal-independent noise – the RZ gain tends to a finite limit as the receiver sensitivity approaches the quantum limit [25].

Figure 5 shows the receiver sensitivity penalty with respect to the quantum limit, γ_q, as a function of the inverse duty cycle $1/d$ for raised cosine RZ pulses, a fiber Bragg grating as an optical filter, and a 5th-order Bessel electrical lowpass filter. The solid curve represents the results obtained when jointly optimizing optical and electrical filter bandwidths for each duty cycle, while the dashed curve applies for optimized optical bandwidth only, keeping the electrical bandwidth constant at its optimum value for NRZ ($d = 1$), $B_e = 1R$. It can be seen that by optimizing optical and electrical bandwidths for each input pulse duration, the receiver sensitivity can be steadily increased when reducing the pulse duty cycle. The receiver sensitivity then asymptotically approaches the quantum limit, however, with the serious drawback of significantly increased demands on electrical receiver bandwidth for little gain in receiver performance. Choosing, for example, $d = 0.1$ increases the performance by just 0.1 dB as compared to $d = 0.3$, and at the same time requires an electrical bandwidth of about $B_e = 5R$. This represents an extremely broadband receiver which is difficult to realize, especially at data rates exceeding 10 Gbit/s. Additionally, optimum sampling of such short pulses is a formidable task and will place heavy demands with respect to allowable timing jitter. On the other hand, if the electrical receiver bandwidth is kept fixed at a technologically reasonable value (e.g. at the optimum bandwidth for NRZ reception), and only B_o is optimized as d is decreased, best performance is achieved for duty cycles between 0.5 and 0.33 (dashed line in Fig. 5). Adjusting the optical filter bandwidth B_o is no problem in practice, since broad optical bandpass filters are easily available. Thus, a RZ duty cycle around 33% is optimum from a receiver point of view. Luckily, this number coincides with the duty cycle inherent to the most widely established RZ pulse source, a sinusoidally driven Mach-Zehnder modulator.

6. Experiment

Using the experimental setup described in Sec. 2, we measured the receiver sensitivity n_s for various combinations of optical and electrical filter bandwidths. The results are shown in Fig. 6, together with the corresponding simulations, by means of sensitivity penalty relative to the quantum limit, γ_q, being 38 ppb in our case.

In (a), we present the receiver sensitivity for NRZ and RZ coding as a function of the optical filter bandwidth B_o at a data rate of $R = 10$ Gbit/s. The electrical bandwidth was kept constant at $B_e = 0.6R$ for NRZ coding, and $B_e = 0.9R$ in case of RZ coding. These values of B_e apply to an effective electrical bandwidth, since it turned out that the finite bandwidth of the BERT input network could not be neglected. Relying on data provided by the manufacturer of the BERT, we modeled the BERT as a 1st order lowpass with a cut off frequency of $f_c = 13.5$ GHz in our simulations. With this adjustment, the agreement between measurement and simulation could be considerably improved. Note the excellent match between experimental results (bullets, triangles) and simulation (curves). In particular, we experimentally confirm the fairly moderate sensitivity

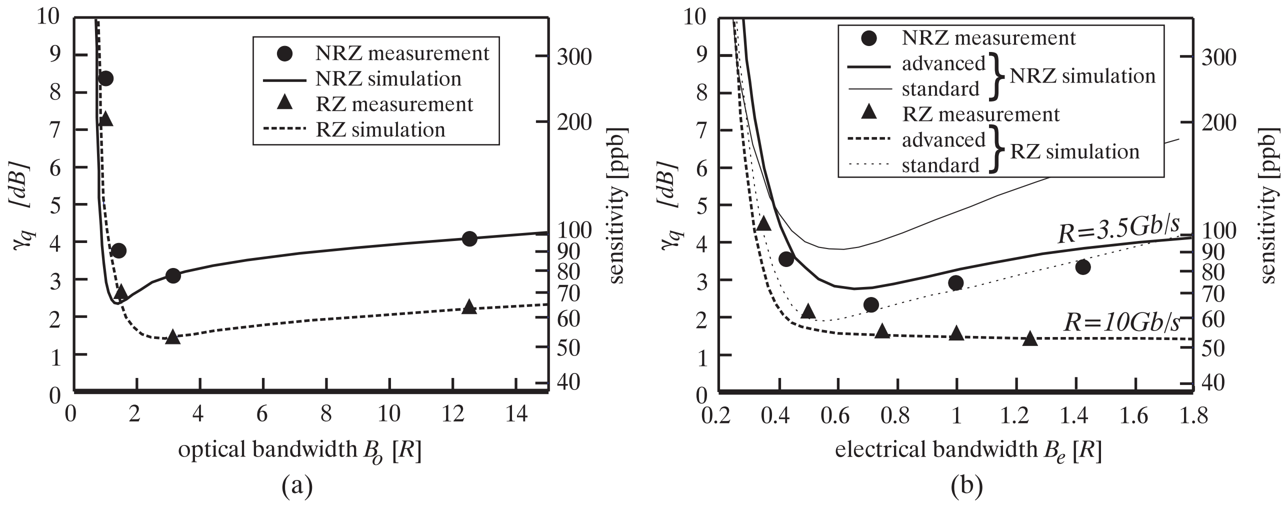

Fig. 6. Measured (bullets, triangles) and simulated (curves) dependence of receiver sensitivity for NRZ and RZ coding on optical (B_o) and electrical (B_e) bandwidth. B_o and B_e are normalized to the data rate R. In (a), results are plotted as a function of the optical bandwidth at a constant electrical bandwidth of $B_e = 0.75R$ (NRZ) and $B_e = 0.9R$ (RZ). In (b), the electrical bandwidth is varied at a constant optical bandwidth of $B_o = 2.86R$ (NRZ) and $B_o = 3.12R$ (RZ)

degradation for higher-than-optimum optical bandwidths, and the severe penalties for too narrow filter choices. For RZ coding, we achieved a sensitivity of 52 ppb (± 2 ppb measurement uncertainty) at the optimum optical bandwidth, but using a slightly suboptimum electrical filter. This is about 1.4 dB off the quantum limit and represents, to our knowledge, the best reported receiver sensitivity at a data rate of 10 Gbit/s. Note, that the optimum optical filter bandwidth might change when a DWDM system is considered.

In Fig. 6 (b), we plot sensitivity versus (effective) electrical bandwidth B_e at a constant optical bandwidth of $B_o = 2.86R$ for NRZ and $B_o = 3.12R$ for RZ, respectively. Additionally, we show simulation results using advanced Gaussian noise statistics (thick lines) and commonly used noise approximations [27] (thin lines). To get around the above mentioned problems associated with the limited bandwidth of the BERT input network, the measurements had to be taken at a reduced data rate of $R = 3.5$ Gbit/s for NRZ, since the delicate influence of ISI on NRZ performance can only be accurately accounted for in simulation if the overall impulse response of the electronics is known sufficiently well. In contrast, the prediction of RZ performance is far less influenced by slightly unknown electrical filter characteristics, since it is an energy truncation effect rather than ISI that limits its performance. Note that while the simulations using advanced Gaussian noise statistics (thick lines) excellently reproduce the measured results, the simulations using the frequently employed approximations for signal-ASE beat noise and ASE-ASE beat noise may be off by up to 3 dB, especially at large bandwidths B_e (thin lines).

7. Conclusion

We have shown by means of numerical simulations that the main reason for enhanced receiver sensitivities when using RZ instead of NRZ in a well designed optically preamplified receiver is intersymbol interference (ISI). A typical RZ coding gain of 1-2 dB can be expected, provided that the receiver setup is optimized with respect to optical and electrical filter bandwidths. The RZ gain can considerably increase if suboptimum receiver bandwidths have to be employed, which could be dictated by technological constraints such as limited electronic bandwidth or laser frequency stability. Concerning the joint optimization of optical and electrical filter bandwidths, RZ proves to be more tolerant to suboptimum filtering.

We have demonstrated that the use of RZ with duty cycles below 33% only leads to minor additional receiver sensitivity improvements, however, at the expense of significantly higher receiver bandwidths. Thus, going beyond 33% duty cycle seems impractical.

Our experimental results accurately confirm the predictions of our simulations. In particular, our measurements prove that too large an optical bandwidth leads only to moderate degradation of receiver sensitivity, whereas severe penalties are encountered if the optical bandwidth is chosen below its optimum. Comparing measurements and simulations, we also prove that simulation errors of several dB may result from employing commonly used, but frequently oversimplified noise formulae.

Employing an RZ duty cycle of 33%, we achieved a receiver sensitivity of 52 ppb at a data rate of 10 Gbit/s, which is only 1.4 dB off the quantum limit, and represents the best reported sensitivity at this data rate.

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