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Editorial

25th Anniversary

If you check the cover page of this issue, you will see that it is volume 25 – something to celebrate?

When we started Journal of Optical Communications 25 years ago it was at a time optical communications. fiber and integrated optics were a niche in the research laboratories of the telecommunications industry; at that time I was a professor at the Technical University of Berlin and started to give lectures on topics such as lasers, wave propagation in dielectric media, modulation of optical waves etc for electric engineers! At that time my father was publishing a journal called "Frequenz", which had an old tradition and was not really willing to change from electric fields to optical waves. Therefore we discussed a project "Journal of Optical Communications" with the publisher of the "Frequenz". We decided to try an international. English journal concentrating on the new R&D-topics connected with the ideas to transmit high bit rate information via optical fibers. I still remember the time (some years before) when I was working at Siemens R&D, where we were happy if semiconductor lasers lasted more than an hour

We had the luck to get a very high level scientific editorial board which supported us in the beginning very much. Most of them are retired or are even no more between us.

It was a very exciting time, but it was 25 years ago: Time is running very fast!

Sincerely yours

Ralf Th. Kersten

Editor-in-chief

Contents

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J Opt Commun 25 (2004) 1 6–9
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Optical Communications

An Advanced Direct Detection Receiver Model

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Summary

An advanced receiver model for the optically amplified communication system (based on EDFA, Raman or hybrid Raman/EDF amplifiers) is proposed. It is independent on modulation scheme, pulse shape, type of electrical filter and type of optical filter. The channel is modeled as a stationary additive noise channel with memory (intersymbol interference-ISI), with the noise at its arbitrary statistics. The validity of the model is verified by Monte Carlo simulations for different modulation schemes like non return to zero (NRZ), return to zero (RZ), chirped return to zero (CRZ) and carrier suppressed return to zero (CSRZ) with respect to the amplifier spontaneous emission (ASE) noise. An excellent agreement is obtained for the all considered schemes.

1 Introduction

Modeling of fiber channels is an active area of research in optical communications. Finding an accurate receiver model is a difficult problem due to the inherent complexity of the physical phenomena involved in transmission of a digital signal and a variety of optical technologies and modulation schemes presently under investigation. Recently several receiver models have been proposed [1–7]. In the model presented in [1], the optical filter is omitted, limiting its application to NRZ/RZ signals for a very narrow region of electrical filter bandwidths. Moreover, the derivation of the signal-independent variance term assumes a flat electrical filter transfer function. In [2] and [6–7], a more comprehensive model is proposed. This model, however, lacks generality It is applicable under the white Gaussian noise approximation of the optical filter input noise, despite the fact that the ASE noise is a narrow band process. Also the analysis is restricted to NRZ receivers. The most comprehensive receiver model to date is proposed in [4–5]. But it is applicable for a simple integrate-and-dump type of the electrical filter for the NRZ signals.

Construction of a general model, derived without any constraints with respect to either the choice of the front-end filters and modulation scheme is still an open problem. This paper introduces a more general model, independent on the modulation scheme, pulse shape, optical and electrical filter types. It can be used for arbitrary additive stationary noise including amplifier spontaneous emission (ASE) and multipath interference (MPI). Such a model allows a comparison among different modulation schemes and in identifying the optimum filter bandwidths. Due to the generality with respect to optical and electrical filter choice, this model allows to find the best optical/electrical filter pair with respect to performance of every particular system. Besides the ASE noise, this model allows other types of noise, such as MPI for Raman amplifiers, without adjusting the model.

The model is validated by Monte Carlo simulations. The analysis includes optically amplified CRZ, CSRZ, RZ and NRZ signals, super Gaussian optical filter and Gaussian electrical filter in the presence of ASE noise and under ISI influence.

2 Model description

The block scheme of the receiver is shown in Fig. 1. The observed electrical field coming through the fiber to the optical filter input can be written as

$$r(t) = s(t) + n(t)$$

$$s(t) = \sum_{n=-\infty}^{\infty} \sqrt{b_n P} p_n(t - nT_b) \tag{1}$$

where s(t) is the (optical) amplifier chain output signal field, $p_n(t)$ is the n^{th} bit pulse shape, P the peak power and b_n is the information content $b_n \in \{r,1\}$, with r being the extinction ratio, $0 \le r \le 1$. The additive noise component n(t) is assumed to be a zero-mean wide sense stationary (ASE, MPI, etc.) with the autocorrelation function

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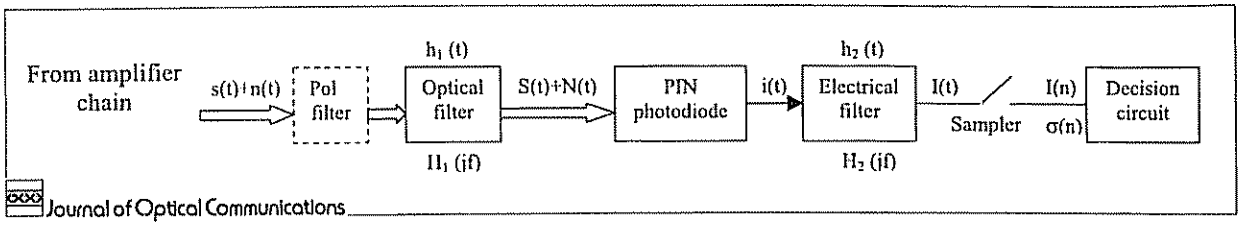


Fig 1: Block scheme of receiver following an amplifier chain

$R_n(\tau)$. In the case of ASE noise, the expression of the auto correlation function is $R_n(\tau) = N_0 R_{EDFA}(\tau)$, where $R_{EDFA}(\tau)$ is the EDFA output optical filter autocorrelation function, and N_0 is the power spectral density of ASE noise per polarization, i.e. $N_0 = n_{sp}(G-1)hf\, N_{amp}$ (within the bandwidth of optical filter in EDFA), with n_{sp} being the spontaneous emission factor, G is an EDFA gain, hf is a photon energy, and N_{amp} is a number of amplifiers. r(t), s(t) and $p_n(t)$ are in fact the complex envelopes of corresponding analytical signals [8].

Let $h_1(t)$ and $h_2(t)$ be the optical filter and electrical filter impulse responses. $h_2(t)$ can be considered as the impulse response of whole receiver electronics, while $h_1(t)$, as the inverse Fourier transform of DEMUX (AWG) transfer function of the observed channel. Since the optical filter is a linear susbsystem, there is no interaction between signal and noise and both the optical filter output signal S(t) and noise N(t) can be written as the convolution of the impulse response and corresponding filter input, that is

$$S(t) = \int_{-\infty}^{\infty} h_1(\tau) s(t-\tau) d\tau$$

and

$$N(t) = \int_{-\infty}^{\infty} h_1(\tau) n(t-\tau) d\tau \qquad (2)$$

The optical filter output noise is a zero-mean process with an autocorrelation function

$$R_N(\tau) = \int_{-\infty}^{\infty} R_{h_1}(t) R_n(\tau - t) dt \qquad (3)$$

where $R_{h_1}(\tau)$ is the autocorrelation function of the optical filter impulse response

$$R_{h_1}(\tau) = \int_{-\infty}^{\infty} h_1(t) h_1^*(t+\tau) dt \qquad (3a)$$

where the asterisk (*) denotes complex conjugate.

The PIN photodiode output can be written as

$$i(t) = R|S(t) + N(t)|^2 = R\left[|S(t)|^2 + |N(t)|^2 + 2R_e\{S(t)N^*(t)\}\right] \qquad (4)$$

R is the photodiode responsivity and will be omitted in further derivations without loss of generality and $R_e\{\}$ denotes the real part of a complex number.

The electrical filter output I(t) is the convolution of the photodiode current i(t) and the electrical filter impulse response $h_2(t)$

$$I(t) = \int_{-\infty}^{\infty} h_2(\tau) i(t-\tau) d\tau \qquad (5)$$

The electrical filter output signal mean follows

$$\overline{I(t)} = \int_{-\infty}^{\infty} |S(\tau)|^2 h_2(t-\tau) d\tau + R_N(0) \qquad (6)$$

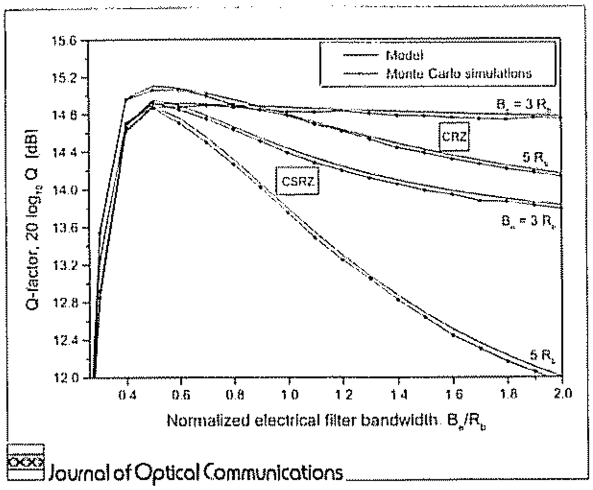
where the overbar denotes the ensemble averaging. In deriving the Equation (6), the following normalization was used

$$\int_{-\infty}^{\infty} h_2(t) dt = H_2(0) = 1 \qquad (6a)$$

with $H_2(j\omega)$ being the electrical filter transfer function.

By using (6a), the electrical filter output signal autocorrelation function can be written as

$$R_I(t_1, t_2) = \int_{-\infty}^{\infty} |S(\tau)|^2 h_2(t_1-\tau) d\tau \int_{-\infty}^{\infty} |S(\hat{\tau})|^2 h_2(t_2-\hat{\tau}) d\hat{\tau} + R_N^2(0) + R_N(0) \int_{-\infty}^{\infty} |S(\tau)|^2 h_2(t_1-\tau) d\tau$$
$$+ R_N(0) \int_{-\infty}^{\infty} |S(\hat{\tau})|^2 h_2(t_2-\hat{\tau}) d\hat{\tau} + 2R_e\left\{\int_{-\infty}^{\infty}\int_{-\infty}^{\infty} S(\tau) S^*(\hat{\tau}) R_N(\tau-\hat{\tau}) h_2(t_1-\tau) h_2(t_2-\hat{\tau}) d\tau d\hat{\tau}\right\} +$$
$$+ 2\int_{-\infty}^{\infty}\int_{-\infty}^{\infty} h_2(t_1-\tau) |R_N(\tau-\hat{\tau})|^2 h_2(t_2-\hat{\tau}) d\tau d\hat{\tau} \qquad (7)$$



Fi 2: Proposed model vs Monte Carlo simulation for two typical modulation schemes for terrestrial long-haul communications (CRZ with modulation depth 1 and phase modulation index 1 rad and CSRZ) and two different optical filter bandwidths (5 R_b and 3 R_b). Bit rate is 10 Gbit/s, extinction ratio 13 dB. Q-factor in absence of ASE noise (determined by transmitter and receiver electronics) is 23 dB Optical signal-to-noise ratio is 15 dB

The first four terms are DC terms, the fifth one is the signal-noise beating term in the photodiode, and the last term comes from noise-noise beating. Note that the mean and the autocorrelation are now functions of time, which means that the electrical filter output process is not sta-ti ;ary any more The reason is that the photodiode is a non-linear element regarding the electrical field

Starting from the expression for the autocovariance

$$L_I(t_1,t_2) = R_I(t_1,t_2) - \overline{I(t_1)}\ \overline{I(t_2)} \tag{8}$$

the variance can be determined by putting $t_1 = t_2 = t$. The derivation of the expression for the variance of the process at the output of an electrical filter is then straightforward, and after inserting the shot noise term $q\overline{I(t)}$, q being electron charge, and the electronic circuitry term σ^2_{elec}, which include both transmitter and receiver electronic noise, also known as "back-to-back" noise, we obtain

$$\sigma^2(t) =$$

$$2R_e\left\{\int_{-\infty}^{\infty}\int_{-\infty}^{\infty} S(\tau)S^*(\hat{\tau})R_N(\tau-\hat{\tau})h_2(t-\tau)h_2(t-\hat{\tau})d\tau d\hat{\tau}\right\}+$$

$$+2\int_{-\infty}^{\infty}\int_{-\infty}^{\infty} h_2(t-\tau)\left|R_N(\tau-\hat{\tau})\right|^2 h_2(t-\hat{\tau})d\tau d\hat{\tau} + q\overline{I(t)} + \sigma^2_{elec} \tag{9}$$

If there is no polarization filter the second term in (9) should be multiplied by factor 2.

Bit error probability can be calculated using the Gaussian approximation [1–2] at the sample time

$$P_e = \frac{1}{N_b}\sum_{n=0}^{N_b-1}\frac{1}{2}\mathrm{erfc}\left(\frac{\left|I(n)-I_{tsh}\right|}{\sigma(n)\sqrt{2}}\right) \tag{10}$$

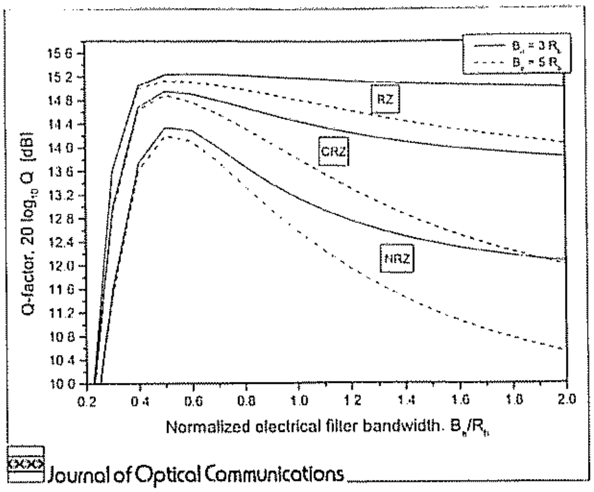


Fig. 3: Comparison among different modulation schemes (NRZ. RZ with duty cycle 0 33. CSRZ) using the proposed method; the set of parameters is the same as in Figure 2

where N_b is the PN generator length The length of a PN generator is to be long enough long so that all possible combinations, that cause ISI, occur at least once. With I(n) and σ(n) as the mean and standard deviation of the n^{th} bit samples determined using the expressions (6) and (9), I_{tsh} as the decision threshold, the error function is defined as

$$\mathrm{erfc}(x) = \frac{2}{\sqrt{\pi}}\int_x^{+\infty} e^{-u^2}du \tag{10a}$$

3 Numerical results

The results of calculation for different modulation schemes, namely NRZ, RZ, CRZ and CSRZ are shown in Figures 2–3. The optical filter is modeled as super Gaussian of order eight, while the electrical filter is modeled as Gaussian. No FEC is applied and no fiber nonlinearities are considered, although the model is general and independent on pulse shape. It is confirmed that the optimum electrical filter bandwidth exists. Results are obtained for optimally chosen threshold, and omitting the polarization filter.

The Q-factor is determined starting from expression (10) by

$$P_e = \frac{1}{2}\mathrm{erfc}\left(\frac{Q}{\sqrt{2}}\right). \tag{11}$$

The Q-factor variation with electrical filter bandwidth B_e, normalized with bit rate R_b, for two different modulation schemes, CRZ and CSRZ, and two different optical filter bandwidths $B_o = 5\ R_b$ and $B_o = 3\ R_b$ is shown in Fig. 2. Curves obtained using the proposed method (full lines) and Monte Carlo simulations ('diamond' marked curves) show excellent agreement. The figure also illustrates as to which modulation scheme is better. For the region of electrical filter bandwidth (0.4 R_b, 0.65 R_b), the CSRZ for both optical filter bandwidths is found to give comparable results to that of CRZ for optical filter bandwidth of $3R_b$ with amplitude modulation depth 1 and phase modulation index 1 rad. Nevertheless, CRZ is least sensitive to the choice of any electrical filter bandwidth. Moreover, for this particular optical filter bandwidth the electrical filter can be omitted without much degradation in performance. CRZ with optical filter bandwidth $5R_b$ is found to be always better.

Figure 3 allows for making comparison among different modulation schemes. It is observed that CSRZ is always better than NRZ, while RZ with duty cycle 0.33 is always better than NRZ and CSRZ.

4 Conclusion

A new receiver model independent on the modulation format, pulse shape, electrical and optical filter choice is proposed. This Model can work for ISI, ASE noise, MPI for terrestrial long-haul communications and other type of noises that can be modeled as stationary processes. The method is compared against Monte Carlo simulations and perfect agreement is found.

This method is applied to compare different modulation schemes (NRZ, RZ, CRZ and CSRZ). RZ (CRZ with phase modulation index 0 rad) is observed to give the best performance. For large phase modulation indices and small optical filter bandwidths, CRZ is found to be less sensitive to the choice of electrical filter bandwidth. But, if the optical filter bandwidth becomes too small electrical filter efficiency decreases.

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