OSNR-Aware Digital Pre-Emphasis for High Baudrate Coherent Optical Transmissions

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Abstract: We derive and experimentally verify an analytical expression for adaptive digital pre-emphasis in coherent optical transmission systems with severe bandwidth limitation by considering both the transmitter response and optical link OSNR.

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1. Introduction

The capacity of coherent optical transponders has been doubled almost every two years by employing more bandwidth efficient modulation formats [1] as well as higher symbol rates [2]. State-of-the-art transponders operate at ~ 90 Gbaud, offering a capacity up to 800 Gb/s by employing PCS-64 QAM format [3]. The next generation transponders are expected to operate beyond 100 Gbaud (e. g. 130 Gbaud). At these high symbol rates, overcoming high-frequency gain reduction under severe component bandwidth limitations while reducing receiver equalization noise enhancement requires a careful design of the Tx digital pre-emphasis (DPE).

In several prior works, DPE is applied for fully pre-compensating the transmitter response to flatten the optical spectrum [1-4]. This technique minimizes the inline amplified spontaneous emission (ASE) noise enhancement due to the Rx equalization. However, it has been recognized recently that full DPE is suboptimal for high baudrate transmission systems with severe bandwidth limitation [5, 6] because it significantly reduces the signal power at the output of the digital to analog converter (DAC) and enhances the impact of quantization noise, driver's noise and ASE noise of the Tx EDFA. In this case, the Tx noise can eventually become the dominant noise source of the system, especially in single-span applications.

In fact, the optimum DPE for a coherent transmission system depends on the transceivers' characteristics (e. g. the Tx response, Tx noise, Rx response and Rx noise), the inline ASE, and the Rx equalizer. In other words, for achieving the best transmission performance, DPE should be adaptively applied based on the link OSNR. Designing DPE considering all the above factors is a complicated task. Recently, there has been an attempt for tackling this problem through link modelling [6]. This approach, however, relies on brute force optimization of DPE within the link model, and thus it is not efficient and does not provide insight into the optimal DPE.

In this paper, we present an accurate model for coherent transmission systems with high order QAM and severe bandwidth limitation. Based on this model, an exact analytical expression for DPE that maximizes the effective Rx SNR is derived, giving insight into how the optimal DPE relates to the system parameters. The proposed DPE technique is verified in a 100 Gbaud PCS-256 QAM transmission, showing from 0.4 dB to 0.8 dB SNR benefit over the full DPE for inline OSNR from 25 dB to 37 dB.

2. Link model description

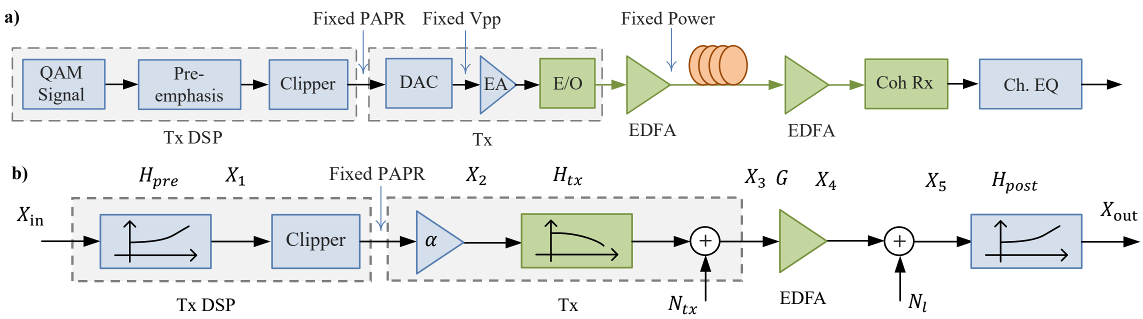

Fig. 1a) – Simplified block diagram of a coherent transmission system; b) – An equivalent model for coherent transmission system with high order QAM and severe bandwidth limitation; EA – Electrical amplifier; E/O – Optical modulator

A basic block diagram of a coherent transmission system is depicted in Fig. 1a. The Tx DSP generates QAM modulated signal using a pulse-shaping technique, such as RRC pulse-shaping, and then DPE is performed. For optimally utilize the physical resolution of the DAC, a clipper is placed at the output of the DPE. The output signal of the DAC is then amplified before being fed to an optical modulator. The optical signal is then amplified by an EDFA before being launched into a fiber link. At the receiver, the signal is received by a coherent receiver and then is fed to an equalizer. To build an equivalent mathematical model of this systems, we assume: ***i***) – the clipper operates such that its output signal has a fixed PAPR, regardless of the DPE and the clipping noise is relatively small compared to DAC quantization noise and driver noise. This assumption is valid for systems with high order QAM formats and small RRC roll-off because DPE does not have a significant impact on the signal's PAPR for such systems (Fig. 2); ***ii***) – The EA operates at a a fixed gain; ***iii***) – Tx EDFA amplifies the signal to a fixed power (optimum launch power) and ***iv***) – The Rx' noise is small compared to the Tx noise and inline ASE noise and Rx equalizer is a zero-forcing equalizer.

Based on these assumptions, an equivalent system model is shown in Fig. 1b. For deriving the optimum DPE filter, we adapt a frequency domain analysis. Assuming the input signal is $X_{in}(f)$, the signal after DPE is $X_1(f) = X_{in}(f)H_{pre}(f)$. As the signal PAPR after the clipper is fixed and both the Vpp of the DAC and EA's gain are fixed, the combination of DAC, EA and modulator can be modelled as a scaler (α) and a low-pass filter as depicted in Fig. 1b where the signal power at the output of the scaler is also fixed. Without loss of generality, we assume: $|X_{in}(f)| = 1$, and $\int |X_2(f)|^2 df = \int \left|\alpha X_{in}(f)H_{pre}(f)\right|^2 df = 1$ where the integral is taken within the interval [-0.5, 0.5] and the signal bandwidth is normalized to 1. Based on that, one can obtain:

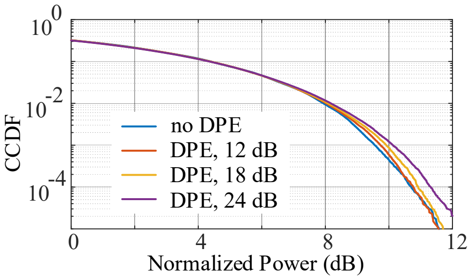

Fig. 2. CCDF of the signal power normalized to the averaged power for 100 Gbaud PCS-256 QAM with 5% RRC roll-off and various DPE levels (at 50 GHz)

$$\alpha^2 = 1/\int \left|H_{pre}(f)\right|^2 df = 1/\int H_{Pre}^2(f)df \tag{1}$$

Under perfect zero-forcing equalization, we can drop the phase responses and focus only on the amplitude responses in further analyses. The signal before the first EDFA is $X_3(f) = X_2(f)H_{tx}(f) + N_{tx}(f)$, where $N_{tx}(f)$ is the accumulated Tx noise density, including quantization noise, EA's noise and the ASE noise of the Tx EDFA. We assume that N_{tx} is white over the signal bandwidth. After the Tx EDFA we have $X_4(f) = GX_3(f) = G\big(X_2(f)H_{tx}(f) + N_{tx}(f)\big)$, where G is the EDFA gain such as its output power is 1 (optimum launch power), e. g. $G^2 \int X_3^2(f)df = 1$. Then one has:

$$G^2 \int X_3^2(f)df \sim G^2 \int X_2^2(f)H_{tx}^2(f)df = G^2\alpha^2 \int X_{in}^2(f)H_{Pre}^2(f)H_{tx}^2(f)df = 1, \tag{2}$$

or $G^2 = \int H_{Pre}^2(f)df \,/ \int H_{Pre}^2(f)H_{tx}^2(f)df$. The signal at the Rx is

$$X_5(f) = X_4(f) + N_l(f) = G\big(X_2(f)H_{tx}(f) + N_{tx}(f)\big) + N_l(f), \tag{3}$$

where $N_l(f) \;=\; N_l$ is the link's ASE noise density. The signal after Rx zero-forcing equalizer is:

$$X_{out}(f) = \frac{X_5(f)}{H_{tx}(f)H_{pre}(f)} = G\alpha X_{in}(f) + \frac{GN_{tx}(f) + N_l(f)}{H_{tx}(f)H_{pre}(f)} \tag{4}$$

One can see that the equalized signal contains two parts.

The first part is the useful signal with power of:

$$P_s = G^2\alpha^2 \int X_{in}^2(f)df = 1/\int H_{Pre}^2(f)H_{tx}^2(f)df \tag{5}$$

The second part is the noise with enhanced power of (under the white noise assumption):

$$P_n = \int \frac{\big(GN_{tx}(f) + N_l(f)\big)^2}{H_{Pre}^2(f)H_{tx}^2(f)} df \;= (G^2N_{tx}^2 + N_l^2)\int \frac{1}{H_{Pre}^2(f)H_{tx}^2(f)} df \tag{6}$$

We can now calculate the noise to signal ratio as:

$$\begin{aligned} \frac{1}{SNR} = \frac{P_n}{P_s} &= (G^2N_{tx}^2 + N_l^2)\int H_{Pre}^2(f)H_{tx}^2(f)df \int \frac{1}{H_{Pre}^2(f)H_{tx}^2(f)} df \\ &= \left(N_{tx}^2 \int H_{Pre}^2(f)df + N_l^2 \int H_{Pre}^2(f)H_{tx}^2(f)df\right)\int \frac{1}{H_{Pre}^2(f)H_{tx}^2(f)} df \\ &= \int H_{Pre}^2(f)\big(N_{tx}^2 + N_l^2H_{tx}^2(f)\big)df \int \frac{1}{H_{Pre}^2(f)H_{tx}^2(f)} df \end{aligned} \tag{7}$$

From Eq. (7), using the integral form of the Cauchy–Schwarz inequality, one can easily prove that:

$$SNR \leq SNR_{max} = 1/\left[\int\left(\frac{\sqrt{N_{tx}^2 + N_l^2H_{tx}^2(f)}}{H_{tx}(f)}\right)df\right]^2 \tag{8}$$

The SNR is maximized ($SNR = SNR_{max}$) when:

$$H_{pre}(f) = c/\sqrt[4]{H_{tx}^2(f)\left(1 + \frac{N_l^2}{N_{tx}^2}H_{tx}^2(f)\right)}, \tag{9}$$

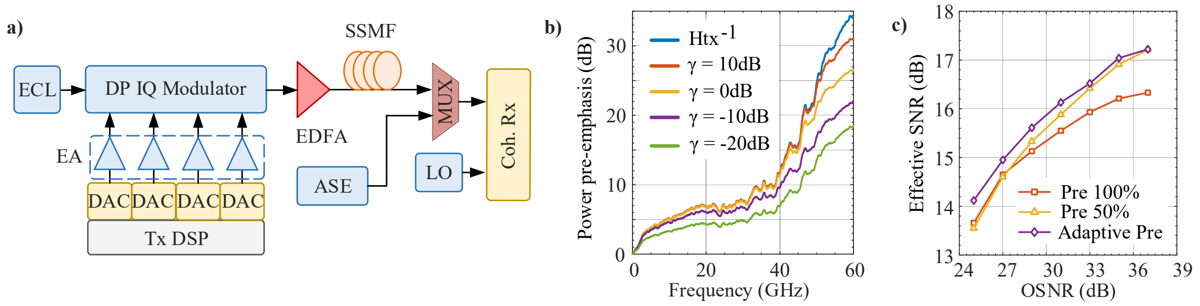

Fig. 3a) – Experimental transmission setup for 100 Gbaud PCS-256 QAM; b) – Adaptive DPE for various values of $\gamma = SNR_{tx}/SNR_l$; c) B2B measurement comparison between 100% DPE, 50% DPE and adaptive DPE. ECL – external cavity laser; LO – local oscillator

where c is an arbitrary constant. Eq. (9) is the main result of this paper, which shows the optimum, adaptive DPE considering the link OSNR. Now we shall consider two special cases:

i) when the inline ASE noise is dominant, i.e. $N_l^2 \gg N_{tx}^2$, we have: $H_{pre}(f) \sim 1/H_{tx}(f)$

ii) when Tx noise is dominant, i.e. $N_{tx}^2 \gg N_l^2$, one has $H_{pre}(f) \sim 1/\sqrt{H_{tx}(f)}$

One can see that when the ASE noise is dominant, the optimum DPE is the inverse of the Tx response, which is the conventional full DPE. However, when the Tx noise is dominant, the optimum DPE is the square-root of the inverse of the Tx response, instead.

Now, we shall discuss how to implement Eq. (9). Applying full DPE in B2B (no inline ASE noise), one has:

$$\frac{1}{SNR_{tx}} = \int H_{Pre}^2(f)\left(N_{tx}^2 + N_l^2 H_{tx}^2(f)\right)df \int \frac{1}{H_{Pre}^2(f) H_{tx}^2(f)} df = N_{tx}^2 \int \frac{1}{H_{tx}^2(f)} df = N_{tx}^2 \left\langle \frac{1}{H_{tx}^2(f)} \right\rangle \quad (10)$$

This SNR_{tx} value can be measured at the manufacturing stage.

Measuring N_l^2 is simpler as it directly relates to the signal to in-band ASE noise ratio (SNR_l) and OSNR as:

$$N_l^2 = \frac{1}{SNR_l} = \frac{B}{OSNR \cdot 12.5\ \mathrm{GHz}}, \quad (11)$$

Where B is the baudrate. From Eqs. (9) - (11), one can implement the proposed adaptive DPE as:

$$H_{pre}(f) = 1/\sqrt[4]{H_{tx}^2(f)\left(1 + \frac{SNR_{tx}}{SNR_l} H_{tx}^2(f) \left\langle \frac{1}{H_{tx}^2(f)} \right\rangle\right)} = 1/\sqrt[4]{H_{tx}^2(f)\left(1 + \gamma H_{tx}^2(f) \left\langle \frac{1}{H_{tx}^2(f)} \right\rangle\right)} \quad (12)$$

where $\gamma = SNR_{tx}/SNR_l$ which shows how much stronger the inline ASE noise is than the Tx noise.

3. Experimental verification for 100 Gbaud PCS-256 QAM

The experimental setup is shown in Fig. 3a, where 100 Gbaud PCS-256 DP-QAM signal (with entropy of 6 bits/2D symbol) with RRC pulse-shaping with roll-off of 5% is generated using 120 GS/s DACs. The signal is then amplified by 4 EAs with ~ 23 dB of gain before being fed to a DP I/Q modulator. The inverse of the Tx response is depicted in Fig. 3b, showing ~ 34 dB of attenuation (in power) at the Nyquist frequency (60 GHz). Fig. 3b shows various DPE filters calculated by Eq. (12) for different values of $\gamma = SNR_{tx}/SNR_l$. The SNR_{tx} measured with full DPE (100% DPE) was ~ 16.35 dB. One can note in Fig. 3b that the shape of DPE filter depends on the OSNR, showing the adaptation of the proposed approach to the link condition. A stronger DPE is applied when OSNR is lower and vice versa. Fig. 3c shows that the proposed adaptive DPE scheme outperforms both 100% DPE and 50% DPE for all considered OSNR values. At a high OSNR value, adaptive DPE achieves a performance improvement comparable to 50% DPE and outperforms 100% DPE by ~ 0.8 dB. At ~25 dB OSNR, this gain reduces to ~0.4 dB. Adaptive DPE converges to 100% DPE when OSNR is further reduced. Overall, Fig. 3c confirms the effectiveness and adaptability of the proposed scheme.

4. Conclusion

We have proposed the first analytical expression for adaptive DPE by considering both the Tx response and the link OSNR, which is very effective for high baudrate coherent transmissions with severe bandwidth limitation. The proposed scheme offers 0.8 dB and 0.4 dB SNR gains at 38 dB and 24 dB of OSNRs for a 100 Gbaud PCS-256 QAM transmission.

5. References

[1] F. Buchali et al, "Rate Adaptation and Reach Increase by Probabilistically Shaped 64-QAM: An Experimental Demonstration," JLT
[2] F. Buchali et al., "1.52 Tb/s single carrier transmission supported by a 128 GSa/s SiGe DAC," in OFC 2020
[3] H. Sun et al., "800G DSP ASIC Design Using Probabilistic Shaping and Digital Sub-Carrier Multiplexing," in JLT., vol. 38, 2020
[4] A. Ghazisaeidi et al., "65Tb/s Transoceanic Transmission Using Probabilistically-Shaped PDM- 64QAM," in ECOC 2016
[5] D. Rafique et al, "Digital pre-emphasis based system design trade-offs for 64 Gbaud coherent data center interconnects," ICTON
[6] Yu Zhao et al, "A Novel Analytical Model of the Benefit of Partial Digital PreEmphasis in Coherent Optical Transponders" ECOC