Modulator Nonlinearity Compensation by Joint Companding/Predistortion for COOFDM

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Abstract: We show that joint digital companding/predistortion for modulator nonlinearity mitigation improves OSNR headroom of coherent optical OFDM systems by up to 2.2 dB (4.4 dB) compared to conventional (no) predistortion.

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1. Introduction

Coherent optical orthogonal frequency division multiplex (COOFDM) can be considered a promising candidate for optical communication with bitrates towards 400 Gbit/s, as it exhibits a simple channel equalization scheme and enables software-defined bitrate and bandwidth making it a potential technology for flexible bandwidth networks. On the downside, the performance of OFDM systems is degraded by nonlinear devices one of which being the external optical inphase/quadrature modulator (IQM). The increasing capacity demand is expected to lead to higher spectral efficiencies and consequently higher order quadrature amplitude modulation (QAM) formats like 64- or even 256-QAM [1]. Especially these modulation formats suffer from transmitter nonlinearity [2]. One method for reducing the impact of the IQM nonlinearity is to apply a digital predistortion to the time-domain OFDM signal. This technique has been shown to improve performance of COOFDM systems using 16-QAM and beyond [3]. However, some distortion might still occur as the modulator transfer characteristic is non-monotonic and its inverse is consequently ambiguous. Since the real/imaginary part of a time-domain OFDM signal is approximately Gaussian distributed, its peak amplitude can be relatively high compared to its mean power. This high peak-to-average power ratio (PAPR) causes clipping in the predistortion device and results in quantization errors raising the required digital-to-analog converter (DAC) resolution. Reducing the PAPR is therefore expected to decrease clipping probability and quantization noise thus improving the system performance in terms of bit error rate (BER). There has been a lot of research effort in PAPR reduction in wireless communications that created techniques like tone injection, block coding or selective mapping [4]. These methods can reduce the PAPR significantly but add computational complexity and require side information to be transmitted, hence decreasing the bit rate. Methods without side information include clipping and companding. While clipping is already inherent in the predistortion device a nonlinear companding transform can further reduce the PAPR without irreversibly distorting the signal, increasing transmitter complexity or decreasing the bit rate. In [5] Chung et al. have studied μ-Law companding for mitigating fiber nonlinearities, while in this work we focus on erf-companding for mitigating transmitter quantization noise and reducing nonlinear distortions induced by the IQM.

2. Companding transform and system model

In general, a companding transform is a nonlinear function $c(x)$ that attenuates large values of x and amplifies small ones. The error function $\mathrm{erf}(x) = (2/\sqrt{\pi})\int_0^x \exp(-t^2)\mathrm{d}t$ fulfills this property and has been shown to achieve the lowest BER in a dispersive channel [6]. It is therefore chosen in this work:

$$c(x) = G_\mathrm{P}\,\mathrm{erf}(G_\mathrm{C}x)\,. \tag{1}$$

The parameters G_C and G_P are positive scaling factors that have to be optimized with respect to minimal BER. The system model used for simulations is shown in Figure 1. The incoming bits are first mapped to complex M-QAM symbols where the modulation order M can be chosen as 4, 16, 64 and 256. Of the total $N_\mathrm{SC} = 256$ subcarriers 91 are left unmodulated to allow transmitter filtering and $N_\mathrm{U} = 165$ carry payload data. An inverse fast Fourier transform (IFFT) is used to obtain the time-domain OFDM signal that is cyclically extended by a guard interval (GI) of length 8 samples. The complex-valued output $w(k) = w_\mathrm{R}(k) + jw_\mathrm{I}(k)$ of the GI insertion block in Fig. 1 is a function of

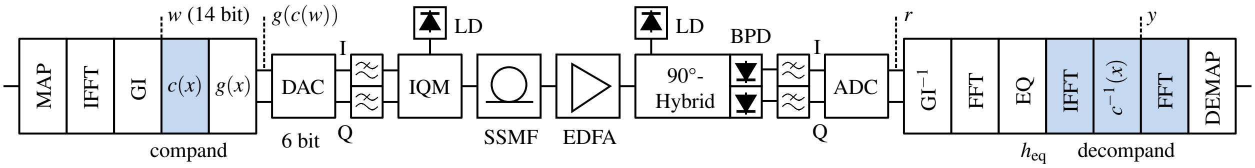


Fig. 1. System model (BPD: balanced photodiode, ADC: analog-to-digital converter).

discrete time k and its mean power $\mathrm{E}\left[w(k)w(k)^{*}\right]$ shall be normalized to $N_{\mathrm{U}}/N_{\mathrm{SC}}$. All signals used in the following are discrete-time so that we can drop k in the remainder of this paper for convenience. The companding function $c(x)$ is applied separately to the real and imaginary part of w followed by the predistortion function $g(c(w))$. The predistortion function is the inverse of the IQM characteristic and is given by:

$$g(x) = \begin{cases} A\,\mathrm{sgn}(x) & \text{for } |x| \geq 1 \\ A\frac{2}{\pi}\arcsin(x) & \text{for } |x| < 1 \end{cases}, \tag{2}$$

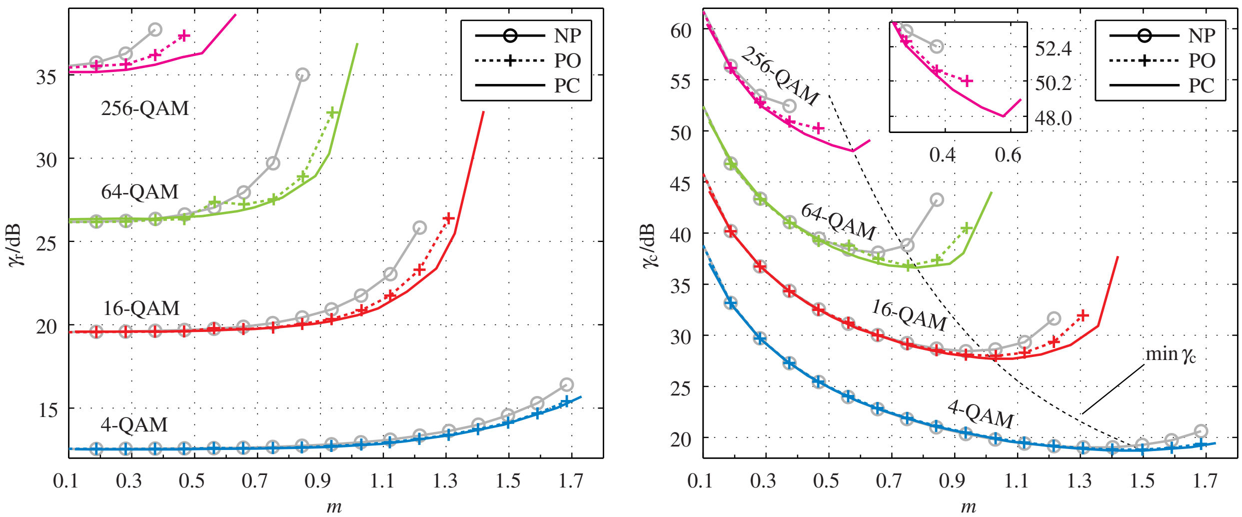
where A is the value that results in the voltage V_{π} driving the IQM to full scale. The discrete-valued implementation of companding and predistortion unit are modeled by a look-up table (LUT) with 14 bit input and 6 bit output word length, respectively. This allows a realistic evaluation of transmitter side quantization effects. The companded and predistorted signal is then converted to an analog signal by a 6 bit DAC at a speed of 32 GSa/s resulting in a bitrate of 40/80/120/160 Gbit/s for 4-/16-/64-/256-QAM (polarization multiplex is not considered). Inphase (I) and quadrature (Q) component of the analog signal modulate amplitude and phase of an optical carrier generated by a laser diode (LD), and the modulated signal is transmitted over a standard single mode fiber (SSMF) that is 80 km long. As we focus on the nonlinearity caused by the IQM, the fiber launch power is assumed to be less than 0 dBm, so that fiber nonlinearities can be neglected. At receiver side the incoming signal is amplified by an erbium doped fiber amplifier (EDFA) adding white Gaussian noise with zero mean. After coherent optical-to-electrical downconversion the sampled signal can be written as

$$r = h * \left\{\sin\left[g\left(c\left(w_{\mathrm{R}}\right)\right)\frac{\pi}{2A}\right] + j\sin\left[g\left(c\left(w_{\mathrm{I}}\right)\right)\frac{\pi}{2A}\right]\right\} + n, \tag{3}$$

where h is the channel impulse response, n is the sampled noise, and $*$ denotes the convolution operator. Eq. (3) also contains the characteristic of the two Mach-Zehnder modulators (MZMs) that constitute the IQM and are operated at the null point. It can be seen from (3) that h should first be equalized before applying $c^{-1}(x)$. For that purpose we use the well-known one-tap frequency domain equalizer (EQ) for OFDM systems: after removing the GI the received signal is transformed to the frequency domain by a fast Fourier transform (FFT) and every subcarrier is multiplied by the inverse of its channel coefficient. The channel coefficients are estimated with the help of 10 pilot symbols. To eliminate the impact of the nonlinear companding transform onto channel estimation the pilot symbols are extracted at transmitter side after the compander, transformed to frequency domain and made known to the receiver as described in [7]. This also guarantees correct scaling of the signal for decompanding. It can be done off-line and hence does not add hardware complexity. After transformation of the equalized signal to time domain the inverse of $c(x)$ is applied yielding the decompander output $y = c^{-1}(h_{\mathrm{eq}} * r)$, where h_{eq} is the equalizer impulse response. The result, y, is fed into an FFT block for demodulation and subsequently demapped to bits, which are compared to the transmit sequence for BER evaluation. Our proposed method adds no complexity to the transmitter compared to conventional predistortion. At the receiver, however, decompander and an additional FFT and IFFT operation are required. As FFT and IFFT exhibit the same length we can use one block for all three transforms in a practical receiver implementation, if the clock rate can be increased accordingly.

3. Results and discussion

For evaluation of the proposed companding technique, Monte Carlo simulations of the system presented in section 2 have been conducted. The system performance of a given configuration is assessed by the optical signal-to-noise ratio (OSNR) γ_{r} in dB, that is required to achieve a BER of 10^{-3}. Depending on the choice of G_{C} and G_{P}, the modulation index $m = \pi V_{\mathrm{rms}}/V_{\pi}$ varies, where V_{rms} is the root mean square of the IQM driving voltage. G_{C} and G_{P} have been optimized in the range $[0.1, \ldots, 5.0]$ for a given modulation index to obtain minimal γ_{r}. In Figure 2a three cases are compared for different modulation orders: neither predistortion nor companding (NP), predistortion only (PO) and joint predistortion and companding (PC). Considering the NP case it is clear that with increasing modulation index the



(a) Required OSNR for a BER of 10^{-3}.

(b) Combined OSNR penalty.

Fig. 2. Comparison of proposed joint predistortion and companding (PC) method with predistortion only (PO) and no predistortion (NP). Inset in Fig. 2b shows detail of 256-QAM.

penalty due to IQM nonlinearity increases as well. As the IQM characteristic is approximately linear for small driving voltages, its impact is negligible for small m. Furthermore, higher modulation orders are obviously more sensitive to the nonlinear IQM characteristic. Using digital predistortion (PO) the required OSNR can be reduced significantly for large m, and using the proposed companding method (PC) γ_r is reduced even further, e.g. by 0.7 dB at $m = 1.2$ and 16-QAM. It is noted that for 256-QAM the PC method improves performance for all m which is attributed to the quantization noise reduction through companding. Although there is no penalty for small m, it is undesirable to operate the modulator at such low m, as it produces a large modulation excess loss a_E, that is caused by not fully driving the MZM. We define $a_E = P_{in}/(a_I P_{out})$, where P_{in}, P_{out} and a_I are the mean optical input power, output power and insertion loss of the MZM, respectively. In our earlier work in [3] we have shown that for OFDM signals it holds $a_E = 2/\left(1-\exp(-m^2/2)\right)$. E.g. for 256-QAM, m must not exceed 0.1 if nonlinear distortion should be negligible (OSNR penalty less than 0.1 dB), which causes an excess loss of 26 dB. Our proposed method extends penalty-free operation to about $m = 0.23$, thus decreasing the excess loss by 7 dB. To study the relation between modulation index, required OSNR and excess loss in more detail we define a combined OSNR penalty $\gamma_c = \gamma_r + 10\log_{10} a_E$ that can be minimized by properly choosing m as shown in Figure 2b. For 256-QAM the proposed PC method improves the minimum of γ_c by 4.4 dB corresponding to a 2.2 dB improvement towards the PO method. The minimum is reached for $G_C = 0.4$ and $G_P = 1.2$. Except for 4-QAM, the PC method generally allows operation at higher modulation indexes.

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