Bandwidth Efficient Precompensation Using a Single Digital-to-Analog Converter

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Abstract: We introduce a new electronic precompensation scheme based on the use of a single digital-to-analog converter. The proposed method allows for arbitrary amount of chromatic dispersion compensation while minimizing the occupied bandwidth of the transmitted signal.

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1. Introduction

Linear electronic precompensation relies on inverting the complex fiber optic channel transfer function. Precompensation schemes proposed so far require the use of two digital-to-analog converters (DAC): one driving the in-phase component of the modulated signal and the other driving the quadrature component of the modulated signal [1, 2]. In this paper, we show that chromatic dispersion precompensation can be accomplished using a single DAC and a single precompensation filter.

2. Reference System Model

The reference system model is depicted in Fig. 1. It consists of a Mach-Zehnder modulator, a dispersive optical fiber channel and a binary phase shift keying (BPSK) demodulator. The Mach-Zehnder modulator is biased at the transmission null point, enabling linear amplitude and phase modulation of the electrical field. The optical fiber is modeled as a dispersive channel with an impulse response of the form:

$$h(t) = (1+i)(4\pi\beta_2 L)^{-1/2} \exp\left(-i\,t^2 / 2\,\beta_2 L\right), \qquad (1)$$

with $\beta_2 = -D\lambda_0^{\,2} / (2\pi c)$; D corresponding to the dispersion parameter, c to the speed of light, λ_0 to the laser source wavelength and L to the fiber length.

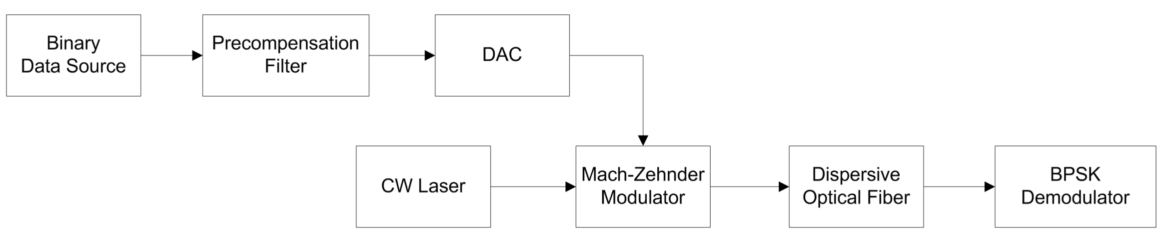

Figure 1. Reference system functional block diagram

The overall transfer function of the optical communication system represented in Fig. 1 is determined by the impulse response of the precompensation filter and the impulse response of the fiber channel. Since BPSK demodulation is used, the received pulse is calculated as follow:

$$r(t) = \mathrm{real}\{s(t) \otimes h(t)\}, \qquad (2)$$

where $s(t)$ is the precompensation filter impulse response and $\otimes$ represents the convolution operator.

3. Statement of the Optimization Problem and Optimization Results

The precompensation filter finite impulse response (FIR) is obtained through numerical optimization. The main goal of the optimization algorithm is set to maximize the energy in the center lobe of the received pulse. Doing so minimizes the system sensitivity to the receiver's sampling clock jitter [3]. To obtain a low-pass frequency response, another optimization goal is set to maximize the energy in the pass band of the precompensation filter. The optimization objective function is thus formulated as:

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$$\text{maximize} \quad f(t,f) = \mu \frac{\int_{-Ts}^{T_s} r(t)^2\,dt}{\int_{-Ts\cdot M/2}^{Ts\cdot M/2} r(t)^2\,dt} + \gamma \frac{\int_0^B S(f)^2\,df}{\int_0^{1/(2T_s/N)} S(f)^2\,df}, \tag{3}$$

where T_s is the symbol duration, N is the oversampling rate (the number of samples per symbol period) and M is the number of symbols covered by the precompensation filter impulse response. $S(f)$ is the Fourier transform of $s(t)$ and parameter B corresponds to the desired low-pass cut-off frequency. Parameters μ and γ are used to adjust the weight of each optimization goal.

For error free transmission, the received pulse should have zero intersymbol interference (ISI). A constraint is therefore added in the optimization algorithm, expressing Nyquist's first criterion [4]:

$$r(kT_s) = \begin{cases} 1, & k = 0 \\ 0, & k \neq 0 \end{cases}, \quad k = 0, \pm 1, \pm 2, \ldots . \tag{4}$$

A second constraint ensures the symmetry of the precompensation filter pulse shape:

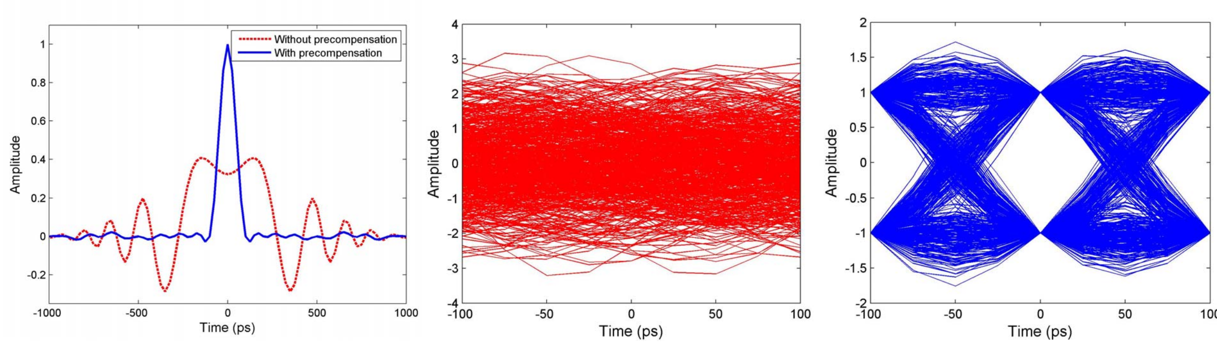
$$s(t) = s(-t). \tag{5}$$

The optimization problem is solved by the Sequential Quadratic Programming algorithm (SQP) [5] and an optimization result example is provided for the parameters listed in Table 1.

Parameter	Value
Fiber Length (L)	800 km
Bit rate (R)	10 Gbps
Symbol time period (T_s)	100 ps
Dispersion (D)	17 ps/nm/km
Emitting wavelength (λ_0)	1550 nm
OverSampling factor (N)	4
Number of symbols covered by precompensation filter (M)	256
Precompensation filter cut-off frequency (B)	10 GHz
Optimization weight for center-lobe energy maximization (μ)	1
Optimization weight for pass-band energy maximization (γ)	1000

Table 1. Optimization parameter values

Figure 2 (a) shows simulated results for the received pulses after fiber propagation. The uncompensated flat-top raised cosine pulse was selected, as well as the optimized precompensation pulse at the transmitter. It can be seen that the received raised cosine pulse spreads over -8T_s to 8T_s while most of the received precompensated pulse energy is contained between $-T_s$ and T_s. The raised cosine pulse spreading causes its associated eye diagram to be completely closed, as showed by Fig. 2 (b). In Fig. 2 (c), the received precompensated eye diagram is ISI free, since the eye is completely open.



(a) (b) (c)

Figure 2. Received uncompensated and precompensated pulses after 800 km of fiber propagation (a); eye diagrams of the received uncompensated (b) and compensated (c) BPSK sequences after 800 km of fiber propagation.

The power spectral density (PSD) of the precompensated sequence is compared to the PSD of a BPSK sequence filtered by a flat-top raised cosine pulse with a roll-off factor α = 1, for the same average output power. Fig. 3 indicates that the pre-compensated sequence spectrum is contained within 20 GHz and that its spectrum shoulders are very low, therefore providing good immunity against channel crosstalk in dense wavelength division multiplexing (DWDM) systems.

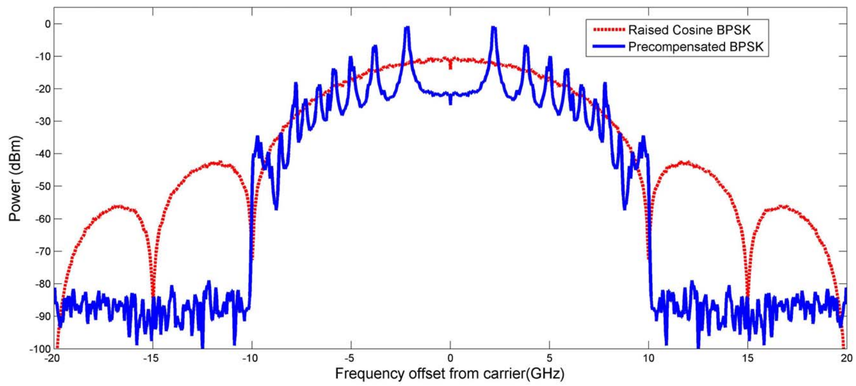


Figure 3. Spectrum density of raised cosine and precompensated BPSK transmitted sequences (sig).

4. Conclusion

In this paper, we have described the first proposal of a single DAC precompensation scheme. The main drawback associated with the proposed method is that the precompensation filter requires 5 to 10 times more coefficients than the precompensation filters of the usual two-DAC approach. This additional resource consumption could be mitigated by synthesizing the precompensation filter impulse response by an infinite impulse response (IIR) filter. Although the results were presented with BPSK demodulation, it should be noted that the proposed technique can also be applied to DPSK demodulation.

Further research work will evaluate the effects of fiber non-linearities, polarization mode dispersion (PMD) and erbium-doped fiber amplifier (EDFA) noise on the proposed scheme.

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