Efficient Frequency-Domain Fractionally-Spaced Equalizer for Flexible Digital Coherent Optical Receivers

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Abstract *Training-aided frequency-domain equalization for coherent receivers based on digital-signal-processing with oversampling factor below 2 samples-per-symbol (sps) is reported. The proposed architecture shows no performance degradation by reducing the oversampling factor from 2 to 1.25 sps.*

Introduction

Coherent detection in combination with digital-signal-processing (DSP) has become a mature technology for optical transport networks. After the commercialization of the coherent 100 Gbps transponders the focus of the industry has moved to higher-capacity (i.e. 400 Gbps and beyond) and flexible-transponders which allow to trade capacity with reach[1]. Such features, unfortunately, come at the cost of higher power consumption.

Indeed, high-capacity systems require high-speed and high-bandwidth power-hungry hardware and, in addition, flexible-transponders require complex DSP coping with different modulation formats. Focusing on the DSP of the coherent receiver, it has been shown in[2] that a high portion of the processing power on the application-specific integrated circuit (ASIC) chip is consumed by the equalizer required to compensate for chromatic dispersion (CD) and polarization-mode dispersion (PMD). Power-efficient DSP which allows flexible-transponders has been recently reported[3,4]. However, a further step is needed in DSP research to relax the requirements on the hardware. In fact, the low-power variant of next generation digital-to-analog converter (DAC) and analog-to-digital converter (ADC) using 14 nm processes offer the possibility to have high electrical bandwidth at the cost of reduced sampling speed. So that DSP working at lower than 2 samples-per-symbol (sps) might be a must for future transponders.

In this paper, we present a training-aided frequency-domain equalizer which allows ADC working at sampling rate lower than the Nyquist rate. Performance evaluation is based on Nyquist coherent optical transmission systems employing polarization-division multiplexed (PDM) quadrature phase-shift keying (QPSK) and 16-ary quadrature-amplitude modulation (16QAM) formats, as well as two popular 4D modulation formats based on set-partitioning (SP)[1], 32SP-16QAM and 128SP-16QAM.

Frequency-Domain Fractionally-Spaced Equalizer Architecture

The proposed DSP architecture is shown in Fig. 1. The equalizer presented in this paper extends the work recently reported in[3] where the single-stage (SS) frequency-domain equalizer (FDE) with blind CD estimation and training-aided 2×2 multi-input multi-output (MIMO) channel estimation has been experimentally demonstrated. Respect to the architecture in[3], the novel equalizer allows a variable length of the fast-Fourier transform (FFT) and a flexible FD decimation in the data path, thanks to a different 2×2 MIMO channel estimation.

Equalization in FD is performed by transferring 50% overlapping blocks of the serial data into the FD by discrete FFT of size χM points, where χ is the oversampling factor of the digitized received data and M is the length of the FFT assuming symbol spaced sampling. After applying the compensating function to each block, the signal is χ-fold decimated and then transferred back into the time-domain (TD) by discrete inverse-FFT (IFFT)

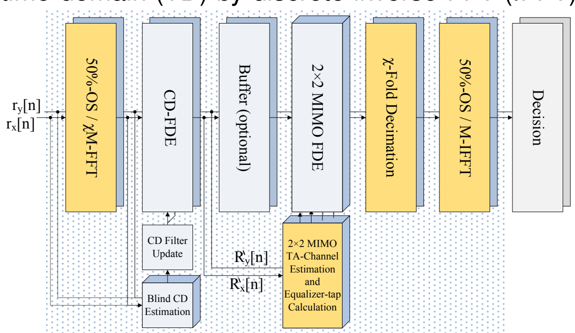


Fig. 1: Single-stage FDE.

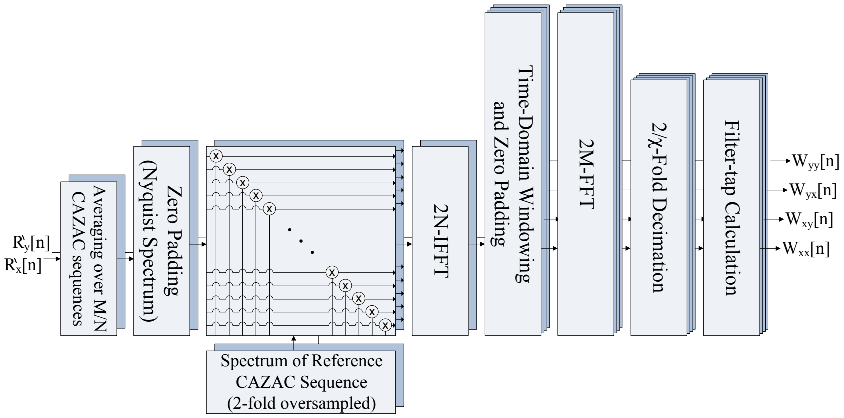


Fig. 2: Detailed structure of the 2×2 MIMO channel estimation and equalizer-tap calculation.

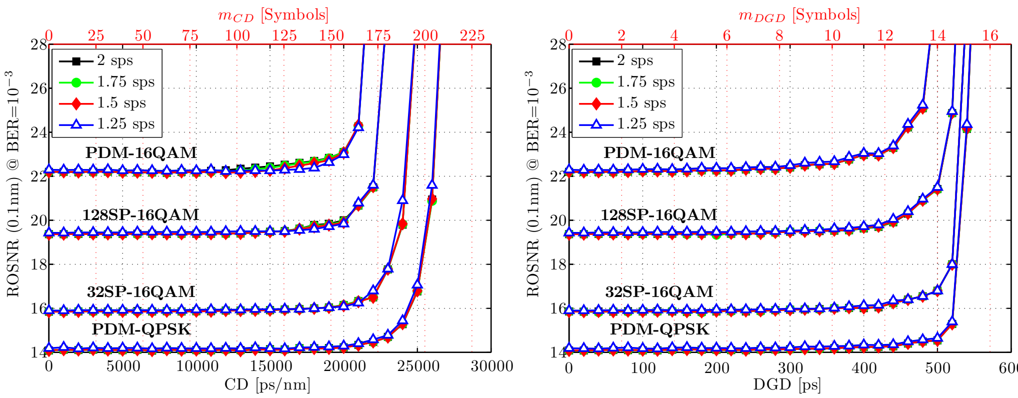
of length M points. The overlap is finally cut off to restore the serial data stream (overlap-save (OS) method). The compensation is applied with a two-tap per sub-band equalizer, where the first tap is dedicated for CD compensation and the second tap for PMD compensation. The length M of the equalizer is chosen accordingly with the maximum CD expected at the point of estimation. Note that a tap of the CD compensation refers to one complex coefficient while a tap of the PMD compensation consists of four complex coefficients. The buffer is an optional module which can be used to compensate for the 2×2 MIMO channel estimation and filter-update processing-delay. CD estimation can be performed either in TD or in FD as in[5,6]. The last DSP module makes bit decision and bit-error counting.

The novel 2×2 MIMO channel estimation is illustrated in Fig. 2. The channel is acquired based on M/N consecutive constant-amplitude zero auto-correlation (CAZAC) sequences, where $N \leq M$ is the length of the CAZAC code in symbols[4]. Based on the spectral properties of a repeated sequence, averaging of the M/N CAZAC codes is performed by retaining every M/N spectral component and discarding the other frequency components containing only noise. The obtained spectra (x- and y-polarization) are then extended to a length of $2N$ spectral components by zero padding. At this point, it should be noted that the reference CAZAC sequences used for 2×2 MIMO channel estimation at the receiver are 2-fold oversampled and consist of RZ50 pulses independently of the pulse shaping used at the transmitter. This choice has been made to preserve the constant amplitude and zero-circular auto-correlation properties of the original CAZAC code. The channel is acquired by a multiplication per sub-band between the received spectra and the conjugate spectra of the reference CAZAC sequences[4]. Such operation is not needed at the frequencies where the spectra contain zeros. The obtained channel elements are then transferred into the TD by discrete IFFT such to extract the four channel impulse responses (CIR) of the 2×2 MIMO channel. This operation is performed by a rectangular time domain windowing function of length N_{TDW} symbols. The four CIR are zero padded such that the length of each vector is $2M$ points. Following, the four channel elements are transferred into the FD by discrete FFT and χ-fold downsampled in order to finally match the length of the equalizer χM taps. The taps of the equalizer can be calculated by using the minimum-mean-square-error, the zero-forcing or matched filter criteria. Depending on the criterion employed a different equalizer-tap solution might be obtained for every χ value.

Performance Evaluation

The performance evaluation is based on simulated 28-GBaud PDM-QPSK, 32SP-16QAM, 128SP-16QAM and PDM-16QAM coherent optical systems transmitting Nyquist shaped pulses (roll-off factor $\alpha = 0.1$). Simulations of the linear channel include CD, first-order PMD referred also as differential group delay (DGD), polarization rotation angle and polarization phase. Additive white-Gaussian noise (AWGN) is loaded onto the signal before an optical Gaussian band-pass filter (2nd-order, double-sided 35-GHz) and an electrical Bessel filter (5th-order, 19-GHz) which defines the opto-electronic front-end of the coherent receiver. An ADC digitizes the received signal at χ sps with $\chi \in \{1.25, 1.5, 1.75, 2\}$ sps. The χ-fold oversampled signal is then processed by the equalizer architecture illustrated in Fig. 1.

The parameter M is set to 256 symbols. CD estimation is performed blindly in FD as proposed in[5]. The 2×2 MIMO channel is estimated by using



(a) CD Tolerance **(b)** DGD Tolerance

Fig. 3: SS-FDE: performance evaluation.

4 consecutive CAZAC sequences of length $N = 64$ symbols with guard interval of length $N_{GI} = 8$ symbols. The TDW is set to be $N_{TDW} = 16$ symbols. The $k = 1, 2, ..., 256\chi$ taps dedicated for CD compensation are calculated as $W_{CD}(\omega_k) = H^*_{CD}(\omega_k)$, where $H_{CD}(\omega_k)$ is the transfer function of the CD and $(\cdot)^*$ denotes the complex-conjugate operation. The $k = 1, 2, ..., 256\chi$ taps of 2×2 MIMO equalizer part are calculated by a matched filter defined as $\mathbf{W}(\omega_k) = \frac{\mathbf{H}^H(\omega_k)}{\Theta(\omega_k)}$, where $\mathbf{H}(\omega_k)$ is the estimated 2×2 MIMO channel, $\Theta(\omega_k)$ serves as a noise-whitening filter (NWF) and $(\cdot)^H$ denotes the complex-conjugate transpose operation. The NWF is calculated as a Frobenius norm of the 2×2 channel matrix $\mathbf{H}(\omega_k)$. These matched filter solutions are valid for any χ[4].

Fig. 3 present the SS-FDE tolerance with regard to CD (Fig. 3a) and DGD (Fig. 3b), respectively. The required optical-to-signal-noise-ratio (ROSNR) at bit-error-rate (BER) $= 10^{-3}$ is plotted as function of the impairment under consideration and as function of the channel memory m_{Ch} defined in terms of adjacent symbols over which a transmitted pulse spreads due to the influence of the channel. Assuming that CD and DGD are the major players for inter-symbol interference (ISI) in an optical transmission system: $m_{Ch} = m_{CD} + m_{DGD}$, where $\mathrm{m}_{CD} = D \cdot \Delta_\omega \cdot B_s \cdot \frac{{\lambda_c}^2}{c}$ and $\mathrm{m}_{DGD} = \tau \cdot B_s$, where D is the accumulated CD, Δ_ω is the signal bandwidth, B_s is the signal baudrate, $\lambda_c = 1550$ nm is the carrier wavelength, c is the speed of light in vacuum and τ is the DGD.

Results show that for the PDM-QPSK system, 20000 ps/nm of CD and 440 ps of DGD can be compensated without OSNR penalty with respect to the back-to-back (B2B) case. Such tolerances are progressively reduced for 32SP-16QAM, 128SP-16QAM and PDM-16QAM, respectively, due to the higher noise sensitivity. As observed from the channel memory tolerance, these curves are strictly dependent on the parameter M, N, N_{GI}, N_{TDW} and on the pulse shaping[4]. However, here it is important to note that for Nyquist systems with $\alpha = 0.1$ no OSNR penalty is observed when the oversampling factor χ is reduced from 2 sps to 1.25 sps. For $\alpha > 0.1$, due to aliasing, the oversampling factor χ needs to be increased to values larger than 1.25 sps.

Conclusions

Single-stage frequency-domain equalization for data oversampled below 2 sps has been demonstrated in PDM-QPSK, 32SP-16QAM, 128SP-16QAM and PDM-16QAM Nyquist systems. For roll-off factor equal to 0.1, no BER performance degradation has been observed for oversampling down to 1.25 sps. The proposed equalizer architecture is suitable for next generation high-capacity and flexible-transponders.

References

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