A novel DSP algorithm for improving the performance of digital coherent receiver using single-ended photo detection

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Abstract A novel digital signal processing algorithm is proposed to improve the performance of digital coherent receivers using single-ended photo detection. The effectiveness of the proposed algorithm is experimentally verified by using a PolMux-RZ-8PSK modulation format.

Introduction

Recently, digital signal processing (DSP) based intra-dyne coherent detection technology has received significant attention for high-speed optical transmission. This technology allows both the amplitude and phase information of the signal to be extracted by coherent mixing of the signal with CW light from the local oscillator (LO) before photo detection. Note that the coherent-mixing term is linearly proportional to the optical field of the original signal, but the direct square-law detection of the modulated signal will cause distortion of the extracted phase and amplitude information, which may severely degrade the performance of DSP-based dispersion compensation, polarization recovery, and PMD compensation. Traditionally, this distortion is mitigated by using either balanced detection [1] or single-ended photo detection with very high local-oscillator-to-signal power ratio (LOSPR) [2].

In this paper we propose a novel DSP algorithm to address this problem. The proposed method allows single-ended photo detection to be used with significantly lower LOSPR. A lower LOSPR will relax the requirement on LO power and RIN spec, and therefore may ease the design of coherent receiver.

Algorithm

For a polarization- and phase-diversity coherent receiver using single-ended photo detection, the signal powers received by the photo detector (PD) in the in-phase (0°) and quadrature branch (90°) at one of the two polarizations are given by

$$P_I(t) = P_S(t) + P_L(t) + 2\sqrt{P_S(t)P_L(t)}\cos(\theta(t)) \quad (1)$$

$$P_Q(t) = P_S(t) + P_L(t) + 2\sqrt{P_S(t)P_L(t)}\sin(\theta(t)) \quad (2)$$

Here $P_S(t)$ and $P_L(t)$ denote the signal and the LO power, respectively. $\theta(t)$ represents the relative phase between the received optical signal and the reference optical signal (i.e. the LO). After photo detection and A/D conversion, the digitized electrical signal with AC coupling can be given by

$$I_I(n) \cong \tilde{I}_S(n) + 2\sqrt{I_S(n)I_L(n)}\cos(\theta(t_n)) \quad (3)$$

$$I_Q(n) \cong \tilde{I}_S(n) + 2\sqrt{I_S(n)I_L(n)}\sin(\theta(t_n)) \quad (4)$$

where $I_S(n)$ and $I_L(n)$ denote the photo-detected signal and LO, respectively. $\tilde{I}_S(n)$ is the AC component of $I_S(n)$. The purpose of the proposed DSP algorithm is to find the approximate value of $\tilde{I}_S(n)$. Assume that $I_L(n) >> I_S(n)$, and to a first-order approximation we can have

$$I_I(n) \approx 2\sqrt{I_S(n)I_L(n)}\cos(\theta(t_n)) \quad (5)$$

$$I_Q(n) \approx 2\sqrt{I_S(n)I_L(n)}\sin(\theta(t_n)) \quad (6)$$

From (5) and (6) we can have

$$I_I^2(n) + I_Q^2(n) \approx 4\tilde{I}_S(n)\bar{I}_L(n) + 4\bar{I}_S\bar{I}_L \quad (7)$$

$$\overline{I_I^2 + I_Q^2} = 4\bar{I}_S\bar{I}_L \quad (8)$$

where the bars over the symbols denote the corresponding DC components (i.e. time-averaged portion). As a result of (7) and (8), we have the first-order approximation for $\tilde{I}_S(n)$ as

$$\tilde{I}_S^{(1)}(n) = \frac{I_I^2(n) + I_Q^2(n) - \overline{I_I^2 + I_Q^2}}{4\bar{I}_L} \quad (9)$$

Note that $\bar{I}_L$ is a constant which only depends on the LO power and the receiver configuration. Its value easily can be determined by doing an initial calibration. For a typical coherent communication system with sampling rate equal to twice of the baud/symbol rate, the known LOSPR (defined as $\bar{P}_L/\bar{P}_S$) can be approximated as $\bar{I}_L/\bar{I}_S$. Thus $\bar{I}_L$ can be found by $\bar{I}_L \approx 0.5\sqrt{\overline{I_I^2 + I_Q^2} \cdot \mathrm{LOSPR}}$.

To get the second-order approximation for $\tilde{I}_S(n)$, we can simply replace $I_I(n)$ and $I_Q(n)$ in (5) and (6), and thus in (9) and (10) by $I_I(n) - \tilde{I}_S^{(1)}(n)$ and $I_Q(n) - \tilde{I}_S^{(1)}(n)$. A similar method can be used to obtain a higher-order approximation for $\tilde{I}_S(n)$.

Experiment

The proposed algorithm has been tested in an 8×114 Gb/s, 25GHz-spaced DWDM transmission experiment using PolMux-RZ-8PSK modulation format. The experimental setup (see Fig. 1) is very similar to the one we described in [3]. The 8PSK signal is generated by using one Mach-Zehnder modulator (MZM1) plus two phase modulators (PM1 and PM2), each driven by a 19 Gb/s data signal. MZM2 is driven by a 19 GHz clock to carve out 50%-duty-cycle RZ pulses. The 19 Gb/s data is generated by time-multiplexing four 4.75 Gb/s 2^{11}-1 PRBS signals. The polarization-multiplexing is achieved by dividing and recombining the signal with a 322 symbol delay before a polarization beam combiner. We have two transmitters (for the odd and the even channels), each modulating four 50-GHz spaced wavelengths (from 1538.7nm to 1540.1nm). The odd and even channels are combined using a 25 GHz interleaver. The line system consists of 8 spans of 80 km SSMF and EDFA-only amplification (with -2.5dBm/ch launch power). No optical dispersion compensation is used in this experiment. We added 10 CW loading channels to

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ensure optimal operation of the in-line EDFAs. At the receiver, the polarization- and phase-diverse coherent detection uses a polarization-diversity 90-degree hybrid, a tunable ECL local oscillator (linewidth~100 kHz) and four single-ended PDs (13dBm input power maximum). The received signal and the LO power can be varied using the gain of $EDFA_R$ and/or $EDFA_L$ to achieve different values of LOSPR. The sampling and digitization (A/D) function is achieved by using a 4-channel Tektronix real-time sampling scope (DSA72004) with 50 Gs/s sample rate and 16GHz electrical bandwidth. The captured data is then post-processed using a desktop PC.

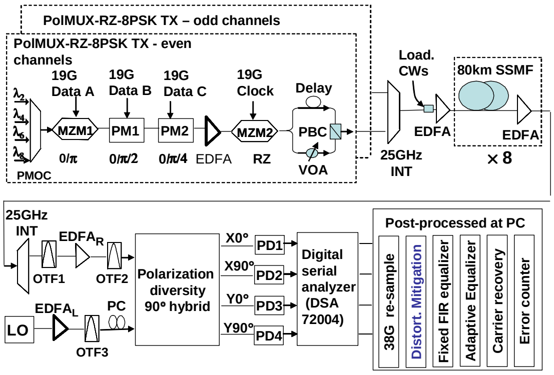

Fig. 1. Experimental setup. PD: photo detector

The DSP algorithm consists of six steps. First the 50 Gs/s signal is down-sampled to 38 Gs/s; second the distortion due to $\widetilde{I}_S(n)$ is removed (approximately) by using the algorithm described in the previous section. Third, the chromatic dispersion is compensated by using a fixed FIR filter with 156 complex-valued taps. In the fourth step, we perform polarization recovery and PMD/residual CD compensation by using 4 adaptive FIR filters (each with 13 taps) optimized by the CMA algorithm [4]. Carrier recovery is performed in the fifth step using the Viterbi-Viterbi algorithm [5]. Finally we carry out differential decoding, Gray-code mapping and BER counting. For this experiment, errors were counted based on 2.4×10^6 bits.

Fig. 2 shows the measured BER of channel 4 (OSNR=25dB) versus various values of LOSPR after 640km of transmission with three different post-processing scenarios: (1) without using the proposed DSP algorithm and using the proposed algorithm with (1) first-order and (3) second-order approximation. Different values of LOSPR were obtained by changing both the signal and LO power in a way that the effective power from the coherent-mixing term remains constant (see Fig.3). The purpose is to make sure that the penalty from the sampling scope due to its noise floor and limited A/D resolution (8 bits) does not change with LOSPR. With the signal and LO powers shown in Fig. 3, the estimated shot noise and thermal noise is small compared to the LO-ASE beating noise. Thus the received electrical SNR before the sampling scope can be approximated as

$$SNR \approx \{\overline{P}_L \cdot OSNR \cdot B_0\} / \{(\overline{P}_s + \overline{P}_L) B_e\} \qquad (10)$$

where Be denotes the electrical bandwidth, and B_0 denotes the optical noise bandwidth used for OSNR measurement. Note that with a constant OSNR, and $\overline{P}_L >> \overline{P}_S$, SNR does not change with LOSPR. Thus the performance difference between different values of LOSPR shown in Fig. 2 mainly comes from the distortion caused by $\widetilde{I}_S(n)$. With LOSPR varying from 9.5dB to 19.5dB, the achieved best BER using the traditional method is 1.6e-3 at LOSPR of 19.5 dB. By using a second-order approximation, however, the proposed algorithm allows a similar performance to be achieved with LOSPR of 11.5 dB. At the same LOSPR of 19.5 dB, the proposed algorithm improved the BER from 1.6e-3 to 9e-4. For the proposed algorithm, the performance with higher order approximation was also investigated, as shown in Fig. 4. The performance difference between order 2 and higher orders is small when LOSPR>10.5 dB. For LOSPR of 9.5 dB, the performance improves for order 1 and 2, but becomes worse for third- and higher-order approximations.

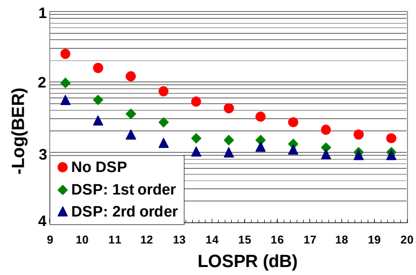

Fig. 2. BER versus LOSPR with different process scenarios.

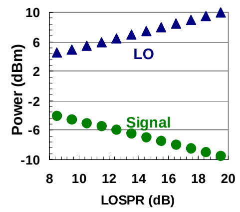

Fig. 3. Signal and LO power incident on the PDs

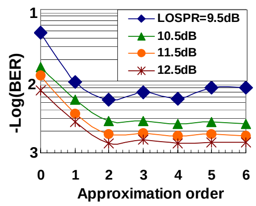

Fig.4. Performance with higher-order approximation

Conclusions

A novel DSP algorithm is proposed and experimentally verified for improving the performance of coherent receivers using single-ended photo detection. More than 7dB reduction in local-oscillator-to-signal power ratio was shown to be possible with the proposed algorithm. This largely relaxes the requirement on LO power and relative-intensity noise.

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