Decision-Directed-Free Blind Phase Noise Estimation for CO-OFDM

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Abstract: We demonstrate an effective decision-directed-free blind phase noise compensation method for CO-OFDM transmission. By applying this technique, the common phase error can be accurately estimated using as few as three test phases.

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1. Introduction

Coherent optical orthogonal frequency division multiplexing (CO-OFDM) has been considered as a promising candidate for long-haul optical communication systems because of its high spectral efficiency (SE) and excellent tolerance towards fiber chromatic dispersion and polarization mode dispersion [1]. However, due to the long symbol duration CO-OFDM is sensitive to laser phase noise, which changes rapidly symbol-by-symbol. The impact of laser phase noise will introduce both common phase error (CPE) and intercarrier interference [2], which significantly degrades the system performance. Therefore, it is crucial that the laser phase noise is rigorously tracked, estimated and effectively compensated.

The laser phase noise impairment can be effectively mitigated by compensating the CPE using pilot subcarriers across the OFDM band [3]. However, this technique reduces the SE. To address this issue, several approaches have been proposed. In [4] a data-dependent pilot-aided technique was introduced to reduce the overhead due to pilot subcarriers by a factor of 2. In [5], a blind phase noise compensation (PNC) method has been proposed based on a decision-directed (DD) algorithm. However, this technique suffers significantly from error propagation, and thus, it cannot be applied in the presence of a large laser phase noise. In [6] the concept of blind phase search (BPS) was proposed, which can be applied effectively without suffering from error propagation. However, this technique also relies on DD (to estimate the mean-square-error) and a large number test phases (16 to 32) are required to achieve good performance. To conclude, existing blind PNC techniques based on BPS or DD are either too complex or seriously affected by error propagation, and thus are not suitable for practical implementation.

In this paper, we propose and experimentally demonstrate a hardware efficient decision-directed free (DDF) blind phase noise estimation method, which is also unaffected by error propagation. We introduce a novel cost function, which allows the CPE to be accurately estimated using just three test phases. The performance of the proposed technique is compared with pilot-aided (PA), DD and BPS compensation techniques.

2. Proposed PNC technique for CO-OFDM and simulation results

By assuming perfect FFT window synchronization, channel estimation and frequency offset compensation, the received OFDM signal $R_{m,k}$ can be expressed as [3]:

$$R_{m,k} = S_{m,k} \exp(j\Phi_m) + e_{m,k},$$  
(1)

where $S_{m,k}$ is the modulated data of the $k^{th}$ subcarrier in the $m^{th}$ symbol before transmission, $\Phi_m$ is the CPE for the $m^{th}$ symbol due to laser phase noise or phase shifts acquired during optical fiber transmission, $e_{m,k}$ represents residual inter-channel interference and random Gaussian noise.

In order to estimate the CPE we propose the following cost function, which is the mean value of the squared product of the projections of real and imaginary parts after rotation by a phase angle $\phi$:

$$F_m(\phi) = \langle \text{Re}(R_{m,k} \cdot e^{-j\phi})^2 \cdot \text{Im}(R_{m,k} \cdot e^{-j\phi})^2 \rangle,$$  
(2)

where $\langle \rangle$ stands for the averaging operation over all of the subcarriers in the $m^{th}$ symbol.

For squared QAM modulation formats with identical probabilities of constellation points, the proposed cost function reaches its maximum value at $\phi = \Phi_m$. This phenomenon can be explained by the fact that ideal squared QAMs provide a “balance” between the real and imaginary parts of constellation points, thus, maximizing the mean value of the squared product of the projections of real and imaginary parts. For simplicity we consider in the following analysis the 4-QAM (QPSK) modulation format with constellation points ($\pm 1 \pm j$). In this case, the expression (2) can be written as:

$$F_m(\phi) = 4 \langle \cos(\Phi_m - \phi + \pi/4) + \eta_{m,k} \rangle^2 \langle \sin(\Phi_m - \phi + \pi/4) + \eta_{m,k} \rangle^2,$$  
(3)
where \( \eta_{i,k}, \eta_{j,k} \) are two random Gaussian variables with zero means and variances \( \left\langle \eta_{i,k}^2 \right\rangle = \left\langle \eta_{j,k}^2 \right\rangle = 1/(2 \cdot \text{SNR}) \). Straightforward calculations show that:

\[
F_m(\phi) = 0.5 \cos(4\phi - 4\Phi_m) + (0.5 + 2 \cdot \text{SNR})
\]

(4)

Extending this result for higher order QAM modulation formats, the cost function (2) can be approximated as:

\[
F_m(\phi) = A \cos(4\phi - 4\Phi_m) + B,
\]

(5)

where \( A, B, \Phi_m \) are three variables to be determined. Knowing the form of the cost function, the CPE (\( \Phi_m \)) can be easily defined using just three test phases, for example, \( 0, \pi/4 \) and \( \pi/8 \) as follows:

\[
B = (F_m(0) + F_m(\pi/4))/2, \quad A = \sqrt{(F_m(0) - B)^2 + (F_m(\pi/8) - B)^2},
\]

\[
\Phi_m = 0.25 \cdot \text{sgn}(B - F_m(\pi/8)) \cdot \cos((F_m(0) - B)/A),
\]

(8)

where \( \text{sgn}() \) is the sign function.

To increase the phase noise tolerance, we consider here a two-stage PNC scheme, as shown in the Fig. 1(a). In the first stage, the received samples of one symbol are compensated using the estimated CPE of the previous symbol. After that the residual CPE (\( \Delta \Phi_m = \Phi_m - \Phi_{m,1} \)) is estimated using the proposed DDF blind estimation technique. If a sufficient number of samples (\( N \)) are taken into calculation, the impact of random Gaussian noise can be effectively removed, allowing us to achieve an accurate estimation of the CPE. The calculated (using (2)) and estimated (using 3 test phases) cost functions for 16-QAM with a SNR of 4 dB and \( N = 200 \) are compared in Fig. 1(b). The root-mean-square error (RMSE) of the CPE calculated by (8) is plotted in Fig. 1(c) as a function of the SNR for 16QAM, showing that a small RMSE of 0.1 rad can be achieved for \( N = 50 \) at SNRs > 7 dB. This result clearly indicates the high tolerance of the proposed CPE estimation method to Gaussian noise.

3. Experimental setup and results

To verify the effectiveness of the proposed DDF blind PNC scheme, we set up both single channel and WDM CO-OFDM transmissions using both the 16-QAM and QPSK modulation formats, as shown in Fig. 2.

The WDM transmission setup comprised a laser grid of five standard DFBs on 100 GHz grid which are substituted in turn by a 100 kHz linewidth ECL. For single channel transmission, only the ECL was used at the transmitter laser. For realizing a more realistic WDM transmission condition, 20 additional loading channels (10 GHz of bandwidth) were generated using an amplified spontaneous emission (ASE) source which is spectrally shaped using a WaveShaper wavelength selective switch (WSS) [7]. A wideband filter was used to filter out-of-band ASE noise at the transmitter. The transmission path is an acousto-optic modulator (AOM) based re-circulating loop consisting of 4 × 100 km spans of Sterlite OH-LITE (E) fiber, which have insertion loss of 18.9 to 19.5 dB/100 km. The loop switch was located in the mid-stage of the third EDFA. After propagation, the signal was filtered using a 4.2 nm flat topped filter before coherent detection with a 100 kHz linewidth local oscillator (LO). The received electrical signals were then sampled by a real-time oscilloscope at 80 Gs/s and processed offline.

The OFDM signals (400 symbols each of 20.48 ns length, 2% cyclic prefix) encoded with QPSK or 16QAM were generated offline in MATLAB using an IFFT size of 512, where 210 subcarriers were filled with data and the remainder with zeros, thus giving line rates of 20 Gb/s (18.2 Gb/s net data rate, after 7% FEC overhead.
removal) and 40 Gb/s (36.4 Gb/s net data rate), for QPSK and 16QAM respectively. For investigating the impact of laser linewidth, the effective laser linewidth was artificially enhanced by passing the received samples \( r(t) \) through a digital filter defined as: \( w(t) = r(t) \exp(\delta) \), where \( \delta \) was the phase noise enhancement and followed a Wiener-Lévy process with a variance \( \sigma^2 \) where \( \nu \) is the enhanced combined laser linewidth and \( \delta t \) is the sampling time. The offline OFDM receiver included: resampling to 25GS/s, timing synchronization, frequency offset compensation, IQ imbalance compensation, CD compensation using an overlapped frequency domain equalizer with the overlap-and-save method, channel estimation with the assistance of an initial training sequence (2 training symbols every 100 symbols) and phase noise estimation and error counting. The BER was obtained by processing 10 recorded traces \((-10^6 \text{ bits})\).

The proposed DDF blind PNC scheme is compared with two-stage blind DD, BPS and PA methods for single channel 16QAM transmission in Fig. 3(a) for a combined laser linewidth of 200 kHz. The proposed method is only matched in performance by highly complex BPS with 16 test phases. Due to error propagation, DD was unable to recover the symbol phases in the system under test, indicating that blind DD is not suitable for high order modulation formats such as 16QAM, even in systems with a typical combined laser linewidth of 200 kHz. A similar comparison result is observed in the Fig. 3(b), which shows the BER as a function of the launch power for the center channel in the WDM QPSK CO-OFDM transmission after 2400 km. The obtained result clearly indicates that the DDF blind PNC is also effective in compensating the CPE due to cross-phase modulation in WDM transmissions. In comparison with PA methods, DDF blind PNC offers better performance while avoiding the SE reduction, due to the pilot subcarriers, which is about 7.6% in this case. This result highlights the great benefit of the proposed PNC technique. The phase noise tolerances of the considered PNC methods are compared in Fig. 3(c). Without differential coding, all blind PNC methods suffer from a phase ambiguity problem if the residual unwrapped phase lies outside the range \((-\pi/4, \pi/4)\), and thus, are reliable only within a certain phase noise range. Fig. 3(c) shows that the proposed DDF blind PNC technique can be applied effectively (without differential coding) with a combined laser linewidth up to 700 kHz.

Fig. 3. Experimental results for (a) BER versus OSNR in the back-to-back case, single channel with 16QAM, the combined laser linewidth is 200 kHz; (b) BER versus the power for the center channel in QPSK WDM transmission, the distance is 2400 km, the combined laser linewidth is 200 kHz; (c) BER versus the combined laser linewidth for the center channel in QPSK WDM transmission, the distance is 2400 km, the power is – 4.8 dBm.

4. Conclusion

We have proposed and experimentally demonstrated a novel DDF blind PNC method for CO-OFDM transmission. Using just three test phases, the proposed technique offers the same performance in comparison with BPS with 16 test phases while avoiding the use of a DD algorithm.

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5. References