Novel iteration-free blind phase noise estimation for coherent optical OFDM

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We propose a novel iteration-free blind phase noise estimation scheme for coherent optical orthogonal frequency division multiplexing (CO-OFDM) systems. In the new algorithm, the cost function is selected as the similar expression with real and imaginary parts as that in the modified constant modulus algorithm, and the new cost function is derived under some assumptions, where it is infinitely approximated by the sine and cosine functions. By means of the analytical formula of the cost function, the initial coarse common phase error can be obtained with only some samples, where the algorithm avoids computational complexity of conventional blind phase noise compensation scheme. In CO-OFDM systems with high-order modulation format (32 quadrature amplitude modulation) and narrow linewidth lasers, it is proved by the simulation results that the phase noise can be effectively compensated with the proposed blind estimation method.

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Coherent optical orthogonal frequency division multiplexing (CO-OFDM) has been considered as a promising candidate for high-speed long-haul optical communications due to its high chromatic dispersion and polarization mode dispersion (PMD) tolerance[1–8]. As is known to all, the coherent system performance suffers seriously from laser phase noise, because it introduces both common phase error (CPE) and inter-carrier interference (ICI).

Many research studies have focused on the phase noise mitigation for CO-OFDM system. The digital pilot-aided scheme is a conventional solution to the problem[9], where frequent pilot symbols are required, and it leads to low bandwidth efficiency. In order to enhance the spectral efficiency, an improved algorithm with the pseudo-pilots is used to reduce the original digital pilot overhead[10]. At the same time, blind phase noise compensation method has also been proposed in CO-OFDM system[11]; however, the computational complexity is significantly increased due to the utilization of orthogonal basis expansion-based algorithms. In Refs. [12,13], a complete blind phase noise correction scheme has been presented, and the decision-directed phase equalizer (DDPE) has been proposed in the non-data-aided scheme. The DDPE has the similar effective performance with conventional equalizer (CE) for low-order modulation formats in the CO-OFDM systems, but the system performance will be degraded greatly by the decision errors for high-order modulation formats. To achieve the good phase noise correction capability, dispersion minimization (DM) algorithm has been utilized before the DDPE for the CO-OFDM systems with high-order modulation formats in Ref. [13], where the cost function used in the constant modulus algorithm (CMA) is chosen with the same expression as the OFDM wireless systems[14–16]. Although the system performance can benefit greatly from the introduction of DM algorithm before DDPE, the algorithm with iteration operations is often wrongly converged due to the severe distortion of cost function affected by the considerable noise in the CO-OFDM systems with high-order modulation formats. Next, the wrong convergence result further resists in the improper CPE compensation in the subsequent DDPE.

In this letter, we propose an effective IFB phase noise compensation method in the CO-OFDM systems with high-order modulation formats. Based on the blind phase noise suppression scheme[13], the cost function in the new DM algorithm has the similar form with real and imaginary parts as that used in the modified CMA (MCMA). The novel cost function is experimentally in close proximity to the sine and cosine functions. Then, the initial CPE can be easily calculated with only some samples instead of the iterative operation because the initial CPE is located on an extreme point of the cost function. Subsequently, the conventional DDPE scheme is used to remove the residual CPE. It is proved by the simulation results that the new phase estimation scheme has the advantages of both high bandwidth efficiency and low computational complexity for high-order modulation formats CO-OFDM systems.

We firstly assume that a complex random sequence s endures phase offset Φ in the presence of white complex Gaussian noise ω, which is drawn from a finite constellation with known statistical characteristics. The output can be written as

$$y = se^{i\Phi} + w. \tag{1}$$

A single tap de-rotator is performed to remove the phase offset, so ϕ is required for estimation. Here ϕ represents an estimate of Φ. The cost function is chosen as the same term as that in the MCMA, and it is written as

$$J = J_{\mathrm{R}} + J_{\mathrm{I}}, \tag{2}$$

where J_{R} and J_{I} are the cost functions for real and imaginary parts of the equalizer output $y=y_{\mathrm{R}}+iy_{\mathrm{I}}$, respectively, and they are defined as

$$J_{\mathrm{R}} = E\left\{\left|R(ye^{-i\phi})\right|^3\right\}, \tag{3}$$

$$J_{\mathrm{I}} = E\left\{\left|I(ye^{-i\phi})\right|^3\right\}. \tag{4}$$

Substituting Eqs. (3) and (4) into Eq. (2), the cost function can be rewritten as

$$J(\phi) = E\left\{\left|R(ye^{-i\phi})\right|^3\right\} + E\left\{\left|I(ye^{-i\phi})\right|^3\right\}, \tag{5}$$

where $R(\cdot)$ and $I(\cdot)$ denote the real and imaginary projection operators, respectively. From Eq. (2), if ϕ was exactly equal to Φ without noise and inter-symbol interference (ISI), the projection of y onto the real and imaginary axes would be composed of a collection of points at the real and imaginary parts of the symbol values defined by the constellation from which s is drawn. While in the presence of noise and ISI, the projection will be made up of a number of clusters centered at these symbol values. As ϕ is slightly different from Φ, the clusters broaden. So a reasonable criterion for evaluating Φ is to try and minimize the dispersion of the projection of the constellation onto the real and imaginary axes, and it can be realized by minimizing the abovementioned cost function J. The validity of cost function can be demonstrated by the circumstantial evidence from the cost function in Ref. [14], because both cost functions have the same form. Moreover, the imaginary part of equalizer output is considered to be introduced in the cost function, and the new algorithm results in the performance enhancement of convergence speed than those of the CMA[14–16].

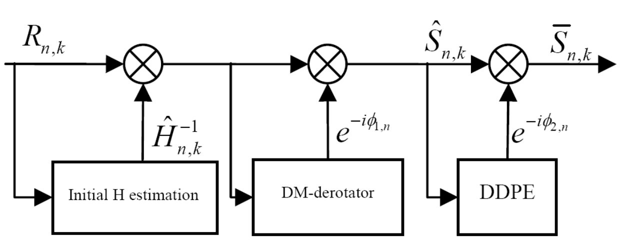

Fig. 1. Structure of the proposed non-iterative blind phase estimation algorithm.

A CO-OFDM system model is considered in the presence of phase noise, where a CO-OFDM frame is composed by N_s OFDM symbols in time domain and N_f subcarriers for each OFDM symbol in frequency domain. $S_{n,k}$ is described as the data sample in the kth data subcarrier $(k=0,\cdots,N_f-1)$ and the nth OFDM symbol $(n=0,\cdots,N_s-1)$. In the CO-OFDM systems, to focus on phase noise estimation, the fast Fourier transform (FFT) window synchronization and carrier frequency estimation can be perfectly carried out, and the fiber nonlinearity is not considered in this letter[9,12,13]. Because the optical channel in CO-OFDM changes relatively slowly, the channel distortion within an OFDM symbol $(n=0,1,\cdots,N_s-1)$ can be regarded as stationary states. Then, the phase drift within one OFDM symbol can be approximated as constant and common to all the subcarriers, and the pure phase rotation is called CPE. In this letter, the iteration-free blind (IFB) algorithm is utilized to perform the CPE mitigation[9,12].

Figure 1 shows the structure of our proposed IFB phase estimation algorithm. For OFDM signals with high-order quadrature amplitude modulation (QAM) format, the de-rotator algorithm with iteration-free operation is added before DDPE scheme to provide accurate decision-making. It is clear that the finally compensated complex data subcarriers are written as

$$\overline{S}_{n,k} = R_{n,k} \cdot \hat{H}^{-1}_{n,k} \cdot e^{-i\phi_{1,n}} \cdot e^{-i\phi_{2,n}}, \tag{6}$$

where $R_{n,k}$ denotes the received data in the kth subcarrier and the nth OFDM symbol and $\hat{H}^{-1}_{n,k}$ denotes the inverse matrix of corresponding estimated channel response. $\phi_{1,n}$ and $\phi_{2,n}$ are the phase estimation values from the samples of cost function and DDPE, respectively.

Subsequently, the IFB phase estimation algorithm is explained completely. Firstly, one pilot symbol is inserted at the beginning of OFDM window/frame only to achieve the initial channel transfer function. Then, the estimated channel transfer function needs to be updated symbol by symbol, and they can be expressed by

$$\hat{H}_{n,k} = \begin{cases} R_{n,k} \cdot S^{-1}_{n,k} & \text{when } n = 0, \\ \hat{H}_{n-1,k} \cdot e^{i\varphi_{1,n-1}} e^{i\varphi_{2,n-1}} & \text{when } n > 0, \end{cases} \tag{7}$$

where $S_{n,k}$ is the pilot data in the kth subcarrier and the nth OFDM symbol at the sender and $R_{n,k}$ is the corresponding received pilot data at the receiver.

Secondly, the rough phase offset $\phi_{1,n}$ was estimated by means of the abovementioned cost function without iterative operations, where the initial phase offset has effect on the elimination of the final false decision for high-order modulation that has been demonstrated in Ref. [13]. Here we shed new light on the initial coarse phase estimation..

Then, the initial coarse phase noise estimation is stated elaborately in combination with the received data. The corresponding received data are substituted in Eq. (5), where the expectation is obtained by calculating the average value after summation, assuming the same probability of data occurrence. It is proved that the cost function is in close proximity to the cosine function of ϕ. So a novel cost function is restructured as

$$J(\phi) = A\cos(4\cdot\phi + B) + C, \tag{8}$$

where three parameters, including the coefficients A, B, and C, can be fixed with three samples from Eq. (5). Then the coarse phase noise estimation $\phi_{1,n}$ is equal to the corresponding value of ϕ at the minimum value of the cost function from Eq. (8).

Finally, in order to remove the residual phase error, DDPE is carried out in the subsequent phase noise estimation. If $\hat{S}_{n,k}$ denotes the de-rotator phase correction version, it is expressed as

$$\hat{S}_{n,k} = R_{n,k}\cdot\hat{H}_{n,k}^{-1}\cdot e^{-i\phi_{1,n}}. \tag{9}$$

Let $D(\cdot)$ denote the decision device. According to the proposed DDPE in Ref. [6], the residual phase correction $\phi_{2,n}$ is equal to the average phase difference between $\hat{S}_{n,k}$ and $D(\hat{S}_{n,k})$ for all OFDM subcarriers, which is written as[9,12]

$$\phi_{2,n} = 1/N_f\sum_{k=0}^{N_f-1}\left(\arg\left\{\frac{\hat{S}_{n,k}}{D(\hat{S}_{n,k})}\right\}\right). \tag{10}$$

If the value of $\phi_{1,n}$ is close to the real desired phase offset, no error occurs for the decision operations in DDPE. Even though the value of $\phi_{1,n}$ is out of work and wrong decisions cannot be avoided, the case can be alleviated in the next DDPE. When the DDPE is completed, it is possible that the residual phase noise is removed and the abovementioned wrong decisions can be corrected in the final decision process.

An optical CO-OFDM transmission architecture is presented in Fig. 2. The optical CO-OFDM consists of the available software modules (commercial software: Optisystem 7.0) except the OFDM coding/decoding components. By using our own matlab codes, the OFDM coding/decoding components are implemented in the transmission system, and our proposed IFB phase noise estimation algorithm is performed in OFDM decoding portion.

The 10 Gb/s original data were firstly mapped onto 256 frequency subcarriers with high-order QAM formats (16 QAM and 32 QAM), and the length of inverse FFT processor size was 512. A cyclic prefix of length 64 was inserted in each OFDM data symbol. Then, the resulting electrical base-band OFDM signal was converted into optical signal by using an IQ Mach–Zehnder modulator. The transmission optical link consists of four uncompensated single-mode fiber spans with dispersion parameter of 17 ps/nm/km, nonlinear coefficient of 1.5 $W^{-1}km^{-1}$, PMD coefficient of 0.5ps/$\sqrt{km}$ and loss parameter of 0.2 dB/km. Spans are 50 km long and separated by erbium-doped fiber amplifiers with the noise factor of 4 dB. Moreover, the injected optical power into optical fiber is limited to about −4 dBm to avoid the fiber nonlinearity in this experiment. The receiver is based on the homodyne CO-OFDM scenario where the local oscillator (LO) wavelength is the same as the transmitter wavelength. The optical OFDM signal then beats with the LO signal in an optical 90° hybrid to obtain I and Q components of the radio frequency signal. The OFDM demodulation component with matlab programs performs frame synchronization, channel estimation, phase noise estimation and compensation, and data decision.

Next, the initial coarse phase noise estimation is carried out based on the numerical simulation results. Figure 3 shows the cost function in Eq. (5) as the function of ϕ, and the function is in close proximity to the cosine function of ϕ according to Eq. (8). To reconstruct the simple cost function, three parameters, including the coefficients A, B, and C, can be fixed with three samples from Eq. (5). For example, when $\phi = 0$, $\pi/4$, and $-\pi/8$, it is easy to obtain the definite values of A, B, and C for these special angles for Eq. (8). In this case, the novel cost function of Eq. (8) as the function of ϕ is shown in the inset of Fig. 3, and it is also demonstrated that the difference between both cost functions is very less. Then the coarse phase noise estimation $\phi_{1,n}$ is equal to the corresponding value of ϕ at the minimum value of the cost function from Eq. (8). It is clear that the rough phase offset $\phi_{1,n}$ can be estimated by means of the novel cost function

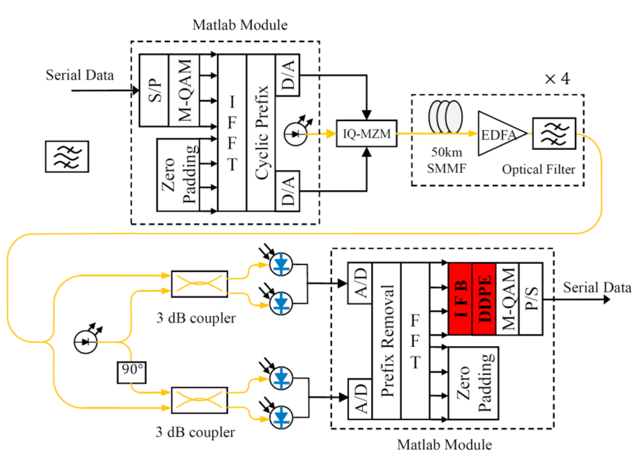

Fig. 2. Architecture of the CO-OFDM transmission system.

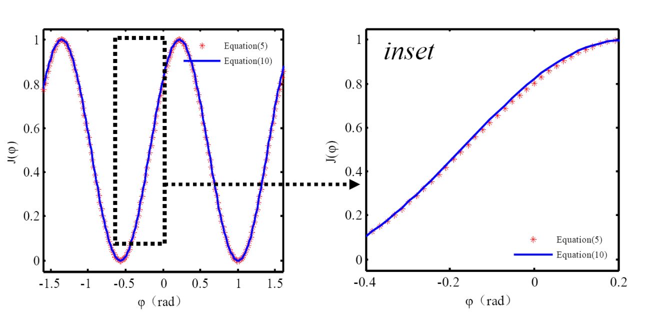

Fig. 3. Cost function of Eqs. (5) and (8).

without iterative operations for the cost function of Eq. (5).

As shown in Fig. 4, the CPE values of the first six OFDM symbols are estimated by using several methods. From Fig. 4, for two different choices of three φ values, there are very small differences between the calculated CPE and desired CPE values, where the desired CPE values are obtained by finding the corresponding value of ϕ at the minimum value of the cost function from Eq. (5) by means of traditional method of exhaustion (called min exhaustion method in Fig. 4). So, it is proved that the estimation of parameters (A, B, and C) is nearly immune to the choice of different ϕ values (at least three) due to the extreme approximation between the both cost functions curves.

In order to evaluate the effect of the IFB phase noise estimation algorithm, the CE algorithm is also proposed in the CO-OFDM systems. The CE scheme is pilot signals (subcarriers)-aided channel estimation, and the density in frequency and time of the pilots must be high enough to perform a channel estimation for all channels, where 64 pilot subcarriers are inserted in each OFDM symbol to achieve the best effect of phase noise correction, and the adjacent interval between the subcarriers is set to 4. At the received optical signal-to-noise ratio (OSNR) values of 40 dB, the linewidth of lasers at both transmitter and receiver sides is set to 30 kHz in the CO-OFDM system with 32 QAM formats. In this case, Fig. 5 shows constellation points of the equalized received symbol with no phase noise correction, CE, and IFB phase noise estimation, respectively. It is obvious that the phase noise estimation is effectively performed by using our proposed IFB scheme, and has good results similar to that of CE. Nevertheless, 64 pilot subcarriers are applied in each OFDM symbol for the CE method, whereas only one OFDM pilot symbol is inserted at the beginning of OFDM data symbols in the IFB method. There is no doubt that the spectral efficiency in the latter is greatly enhanced compared with that in the former. The algorithm with iteration operations in Ref. [13] is applied in the CO-OFDM transmission system with 32 QAM formats, and it is often wrongly converged because the cost function in Ref. [13] is severely distorted by the considerable noise in the 32 QAM formats CO-OFDM systems. Compared with that algorithm, no iterative operations is required in our proposed IFB algorithm, and the coarse phase noise estimation $\phi_{1,n}$ can be easily calculated.

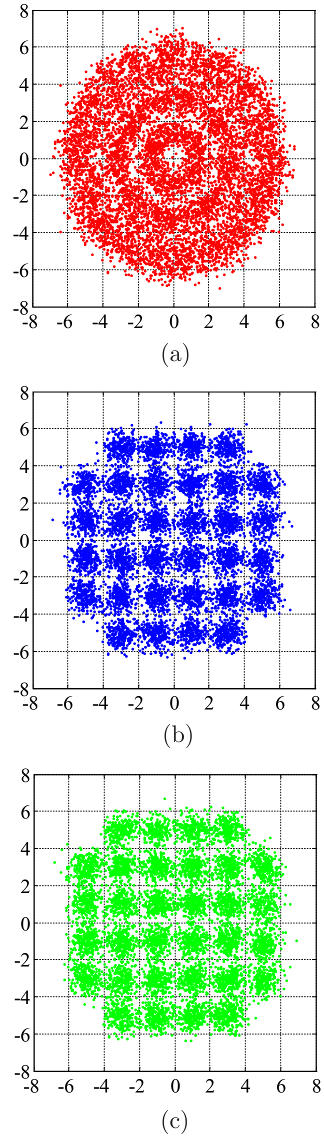

Fig. 5. Constellation points of the equalized received symbol: (a) without phase noise estimation algorithm, 32 QAM, (b) after CE, 32 QAM, and (c) after IFB, 32 QAM.

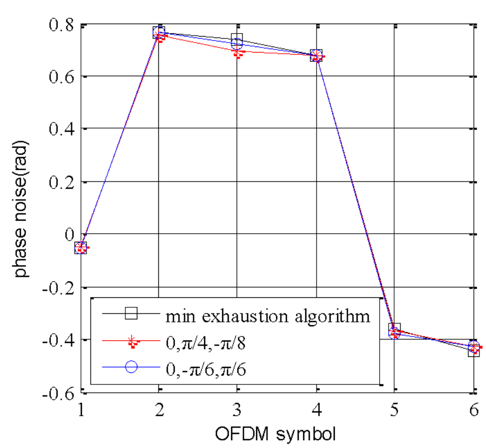

Fig. 4. CPE values of the first six OFDM symbols calculated by using several methods, including two different choices of three φ values and traditional method of exhaustion (called min exhaustion method).

Figure 6 shows the bit error rate (BER) curves versus OSNR at the receiver for the CO-OFDM systems with high-order QAM (16 QAM and 32 QAM) formats when the laser linewidths are set to 30 and 100 kHz, respectively. As shown in Fig. 6(a), the proposed IFB algorithm and CE have almost similar performance for 16 QAM with the linewidth of 30 kHz. However, for 32 QAM, the BER of two algorithms significantly deteriorates, and the proposed IFB algorithm provides a 1 dB drop with a BER of $10^{-2.5}$. As shown in Fig. 6(b), at the linewidth of 100 kHz, the BER performances for two algorithms decrease sharply, specifically for 32 QAM. The reason is that the probability of decision error increases remarkably at the low SNR, and our proposed IFB algorithm is very sensitive in this case. Then the performance drops greatly in the CO-OFDM system for high-order QAM formats with the spread of decision error. Moreover, the compensation of the CPE does not always suffice, especially for larger laser linewidth and higher order modulation format, and it is necessary to perform a blind ICI mitigation in the CO-OFDM system after the phase CPE correction using our proposed IFB scheme[17–19]. Nevertheless, our proposed algorithm can obtain the same performance as that of CE in the CO-OFDM system with the high-order QAM and the narrow linewidth, and it is proved that our proposed IFB algorithms has reached the limitation of phase noise estimation scheme for the CPE compensation rather closely.

Note that the subcarrier number has effect on the performance of the proposed algorithm. With the reduction in the subcarrier number, the duration of the OFDM symbol becomes short, and the phase noise becomes little. In this case, for the proposed data-dependent IFB algorithm, the samples in each OFDM symbol are not enough to perform the algorithm, and the performance of the algorithm becomes severely degraded. Limited to the length of the letter, the problem will be studied in the following research.

In conclusion, we propose an IFB phase noise compensation scheme to mitigate the CPE with high spectral efficiency and low computational complexity in CO-OFDM system with high-order modulation formats. The new cost function is reconstructed by obtaining only some samples, and the initial and coarse CPE can be easily obtained without iterative computation. The simulation results show that the proposed IFB scheme can perform similar to CE even in high-order modulation formats.

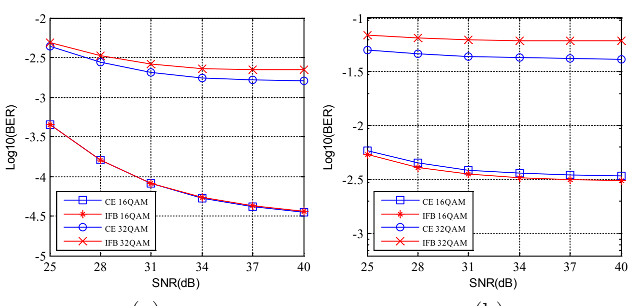

Fig. 6. BER performance of IFB and CE. Linewidth of (a) 30 and (b) 100 kHz.

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