Efficient Blind Carrier Frequency Offset Estimation for Coherent Optical OFDM Systems

Xinwei Du[1,2], Jing Zhang[3], Yan Li[1],Zhuoran Xu[2], Mohan Gurusamy[1], Changyuan Yu[4], and Pooi Yuen Kam[1,2]

[1]Department of Electrical and Computer Engineering, National University of Singapore, 117583, Singapore
[2]National University of Singapore (Suzhou) Research Institute, China 215123, China
[3]Key Laboratory of Optical Fiber Sensing and Communications, Ministry of Education, University of Electronic Science and Technology of China, Chengdu, 611731, China
[4]Department of Electronic and Information Engineering, the Hong Kong Polytechnic University, Hung Hom, Kowloon, Hong Kong
a0123828@u.nus.edu

Abstract: We propose an efficient blind carrier frequency offset estimation algorithm for CO-OFDM systems, which utilizes only one OFDM symbol without exhaustive search computations.
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1. Introduction

Coherent optical orthogonal frequency-division multiplexing (CO-OFDM) has been an attractive technology for long-haul optical communication systems due to its high spectral efficiency and excellent tolerance to the fiber chromatic dispersion (CD) and polarization mode dispersion (PMD). However, CO-OFDM is sensitive to the carrier frequency offset (CFO) that is induced by the incoherence between the transmitter laser and the local oscillator (LO), because its symbol duration is longer than that of a single carrier system. The CFO normalized by the subcarrier spacing can be divided into an integer part and a fraction part. If the CFO is not properly dealt with, even 0.01 normalized CFO can lead to severe system performance degradation [2].

The CFO compensation algorithms can be divided into two categories: data-aided [1-3] and blind CFO estimation algorithms [4-6]. Schmidl and Cox [1] proposed to use two training symbols to estimate the integer part of CFO and the fractional part can be obtained by taking the correlation of the two identical halves of the first training symbol. To reduce the computational complexity in Schmidl's algorithm, S. Cao [3] proposed to insert a high-power pilot-tone in the middle of the spectrum, and the estimated integer part is the shifted positions of the pilot tone. For most of the data-aided algorithms, an overhead is required which reduces the effective transmission bit rate. Blind CFO estimation algorithms can not only solve this problem, but also increase the estimation accuracy. In [4], the cyclic prefix (CP) was used to estimate the CFO blindly, however, the estimation performance is highly dependent on the length of CP. H. Jeon [5] proposed a blind deterministic CFO estimation method which simplified the cost function into a cosine function and the estimated CFO can be obtained by given three test values with one OFDM symbol. This method successfully reduces the search complexity, nevertheless, the OFDM symbol needs to perform two-fold sampling before CFO estimation.

In this paper, we propose an efficient blind CFO estimation algorithm for CO-OFDM systems. Only one OFDM symbol is used for estimation and no extra operation such as oversampling is needed. A cost function is first proposed by utilizing the equal property of time shifted samples and then the cost function is simplified to a cosine function. The estimated CFO can be obtained by just solving the equations without numerical search. The normalized CFO estimation range of the proposed method is $[-0.5, 0.5)$.

2. The Principle

The n-th received time domain OFDM sample $r(n)$ with the appearance of CFO and laser phase noise (LPN) can be expressed as [7]:

$$r(n) = x(n) \otimes h(n) \cdot e^{j(2\pi f n T_s / N + \theta)} + w(n) \tag{1}$$

where $x(n)$ and $h(n)$ represent the transmitted samples and channel impulse response, respectively. Term f is the CFO, which can be normalized by the subcarrier spacing $f_0 \left(=1/T_s\right)$ and divided into the integer and fractional parts: $f = \varepsilon f_0 = \left(\varepsilon_i + \varepsilon_f\right) f_0$, with ε_i as an integer and $\varepsilon_f \in [-0.5, 0.5)$. T_s is the symbol period and N is the FFT size. Additionally, θ represents the laser phase noise and $w(n)$ is the additive white Gaussian noise (AWGN).

If the received time domain OFDM symbol is shifted by i samples, the previous received OFDM symbol $r(n-i)$ can be obtained as shown in Fig. 1(a). If we do not consider the noise term $w(n)$ and CFO ε, $r(n)$ and

$r(n-i)$ are related as: $R(k)=R_i(k)e^{j2\pi ik/N}$, where $R(k)$ and $R_i(k)$ are the corresponding expression of $r(n)$ and $r(n-i)$ in the frequency domain, separately.

However, in the presence of CFO, there will be instead a phase shift in the time domain. In this case, due to the inter-carrier interference (ICI) induced by the CFO, the equality relationship above no longer holds. If the CFO is exactly compensated before operating the FFT on $r(n)$ and $r(n-i)$, the compensated frequency domain signal will satisfy the equality relationship again. Thus, we define the least-square cost function:

$$F(\hat{\varepsilon})=\sum_{l=-N/2}^{N/2-1}\left|D(l,\hat{\varepsilon})-D_i(l,\hat{\varepsilon})e^{j2\pi il/N}\right|^2 \tag{2}$$

where $\hat{\varepsilon}$ is a trial CFO value, and $D(l,\hat{\varepsilon})$ and $D_i(l,\hat{\varepsilon})$ are the compensated frequency domain signals of $r(n)e^{-j2\pi n\hat{\varepsilon}/N}$ and $r(n-i)e^{-j2\pi(n-i)\hat{\varepsilon}/N}$. Note that when $\hat{\varepsilon}=\varepsilon$, the CFO is exactly compensated and $D(l,\hat{\varepsilon})-D_i(l,\hat{\varepsilon})e^{j2\pi il/N}=0$. Thus, the CFO can be estimated by finding the candidate that minimizes the cost function: $\hat{\varepsilon}=\arg\min_{\hat{\varepsilon}} F(\hat{\varepsilon})$. To reduce the complexity caused by the exhaustive search, here we simplify the cost function to a cosine function [5]:

$$F(\hat{\varepsilon})=A\left(1-\cos 2\pi(\varepsilon-\hat{\varepsilon})\right)+B\,. \tag{3}$$

Since there are three unknown variables in the cost function, which are A,B and ε, we can solve the equation by utilizing the property of cosine functions and calculating three cost function values, $F(0),F(0.25)$ and $F(0.5)$.

$$F(0)=A(1-\cos 2\pi\varepsilon)+B,\ F(0.25)=A(1-\sin 2\pi\varepsilon)+B,\ F(0.5)=A(1+\cos 2\pi\varepsilon)+B\,. \tag{4}$$

After calculation, we obtain:

$$A\cos(2\pi\varepsilon)=\left(F(0.5)-F(0)\right)/2,\ A\sin(2\pi\varepsilon)=\left(F(0.5)+F(0)\right)/2-F(0.25)\,. \tag{5}$$

Define a complex variable $z=A\cos(2\pi\varepsilon)+jA\sin(2\pi\varepsilon)$. Hence, the CFO ε can be finally obtained by:

$$\hat{\varepsilon}=\measuredangle z/(2\pi) \tag{6}$$

where $\measuredangle\{\cdot\}$ is the angle operation. Since the range of $\measuredangle z$ is $[-\pi,\pi)$, the resulting estimation range is $[-0.5,0.5)$.

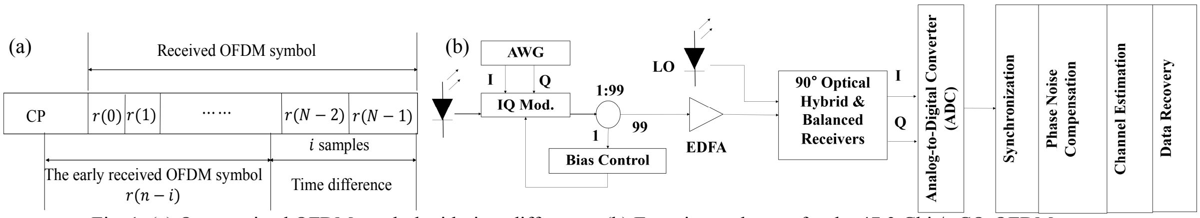

Fig. 1. (a) One received OFDM symbol with time difference; (b) Experimental setup for the 47.3 Gbit/s CO-OFDM system

3. Simulation and Experimental Results

The 47.3 Gbit/s CO-OFDM system is illustrated in Fig. 1(b). The length of transmitted Pseudo-Random Binary Sequence (PRBS) is 2^{15}-1, which is mapped onto 16-QAM. The FFT size is 256, with 128 effective subcarriers and 32-sample CP. Five pilot subcarriers are used for LPN estimation. The OFDM baseband signal is generated by an Arbitrary Waveform Generator (AWG) operating at 25 GSa/s, then fed into the optical IQ modulator to generate the 16-QAM OFDM signal. A bias control is applied to find the null point. An Erbium-doped fiber amplifier (EDFA) is employed to provide an on/off gain of 10 dB. Same with the transmitter laser, the local oscillator is operated at 1550 nm with the laser linewidth of 200 kHz. The optical signal is down-converted to the electrical field by using a balanced receiver and the electrical signal is digitized by ADCs at 50 GSa/s and stored for DSP by using MATLAB. The effective bit rate is 47.3 bit/s with the OFDM subcarrier spacing being 97.9 MHz.

Fig. 2(a) shows the MSE performance of the proposed method versus the number of shifted samples under various SNR and normalized CFO values. As shown in Fig. 2(a), the MSE curve is almost a straight line when i is larger than 8. Besides, in the presence of different CFOs, the MSEs of the proposed method under certain SNRs are quite similar, which illustrate the stable estimation performance. In Fig. 2(b), we investigate the influence of ICI caused by the laser phase noise (LPN) on the CFO estimation performance of the proposed method. The MSE of CP algorithm [4] is shown for comparison. It can be seen that the MSE of the proposed algorithm is always lower than that of the CP algorithm, which can be reduced nearly 2 orders without the appearance of LPN. Besides, the proposed algorithm can achieve the MSE of 10^{-4} with 100 kHz laser linewidth at even 2 dB SNR, which indicates its excellent noise tolerance. To verify the stability of the proposed algorithm, Fig. 3(a) and Fig. 3(b) demonstrate the MSE and BER of the proposed algorithm versus the CFO at various SNR with the laser linewidth of 100 kHz. The MSE and BER curves are all nearly straight lines in the estimation range of [−0.5, 0.5), which is consistent with the theoretical derivations. Fig. 3(c) shows the experimental BER performance versus the received optical power with about 3 GHz CFO. We use our previous algorithm in [11] to estimate the integer part of CFO. Obviously, the proposed algorithm outperforms the CP algorithm. When the received optical power is less than −16.5 dBm, the BER difference is not obvious since the dominant factor is the noise rather than the CFO. When the received optical power is above −16.5 dBm, the BER performance of the proposed algorithm is much better, which is two times lower than that of the CP algorithm at −12.5 dBm.

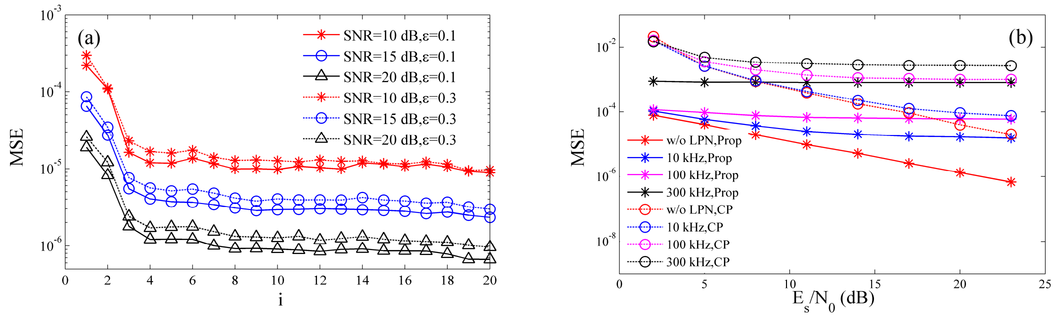

Fig. 2. (a) MSE versus the number of shift samples under various SNR and normalized CFO; (b) MSE versus SNR of the proposed and CP algorithms under different laser linewidths with the normalized CFO of 0.2 and $i = 8$

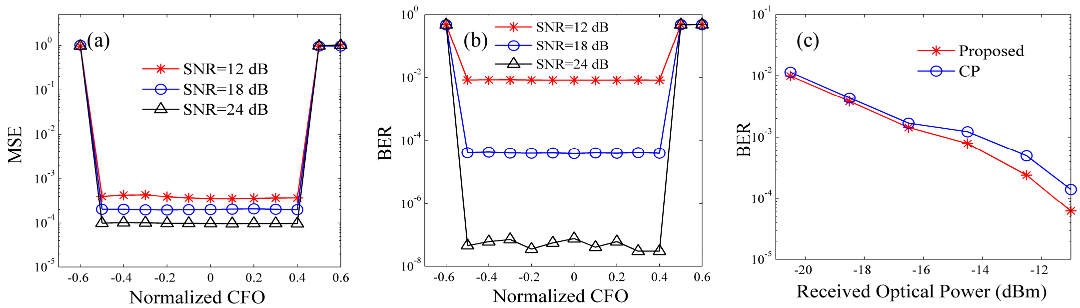

Fig. 3. (a) MSE (b) BER versus normalized CFO at different SNR values with the laser linewidth of 100 kHz; (c) Experimental BER performance of the proposed and CP algorithms versus the received optical power with the CFO of 3 GHz and laser linewidth of 200 kHz

4. Conclusion

We proposed a blind CFO estimation algorithm by using only one OFDM symbol, whose cost function is simplified to a cosine function with lower complexity. From the performance comparison between the proposed and the conventional CP algorithm, the feasibility, stability and effectiveness of the proposed algorithm are verified.

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