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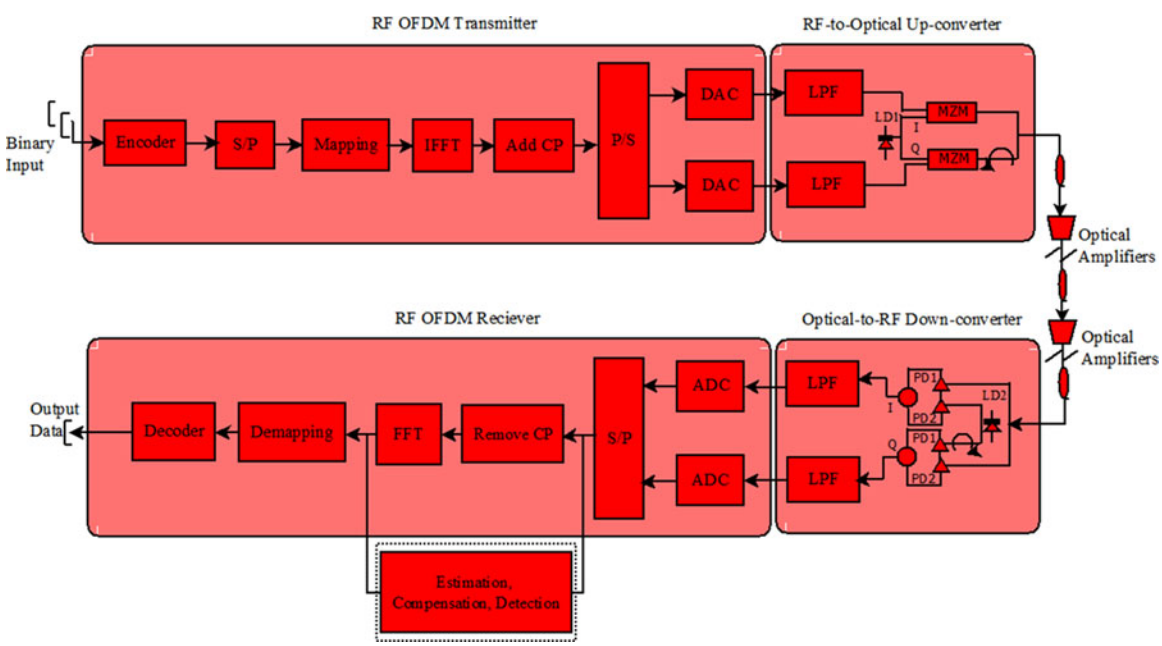

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Efficient Constant Modulus Based Carrier Frequency Offset Estimation for CO-OFDM Systems

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Abstract: This paper proposes a blind carrier frequency offset (CFO) estimation method for coherent optical orthogonal frequency division multiplexing (CO-OFDM) systems, using constant modulus signaling. The proposed scheme is based on a robust cost-function, which deviates from the common assumption that the channel frequency response slowly varies either in time or frequency domain. The proposed method adopts a cost-function similar to the Godard's method for blind channel equalization. Using Monte Carlo simulations, the proposed method is shown to offer a superior performance compared to prominent existing methods, in a practical optical link scenario. Also, we show that the proposed cost-function can be approximated and expressed in a closed-form in such a way that the CFO estimate is obtained using only three trial values.

Index Terms: CO-OFDM, carrier frequency offset, blind estimation, constant modulus.

1. Introduction

The deployment of optical networks has become inevitable due to the phenomenal advancement in the communications industry and the associated extraordinary demand for high data throughput. The orthogonal frequency division multiplexing (OFDM) technique has been incorporated so as to enhance the overall output of the existing optical transport networks. The coherent optical orthogonal frequency division multiplexing (CO-OFDM) has received considerable attention in recent years due to its robustness, efficiency and flexibility [1]–[7]. However, the CO-OFDM is highly susceptible to carrier frequency offset (CFO), which degrades the overall performance and system efficiency [8], [9].

An increasing number of research studies have presented various CFO estimation methods, in both the optical and wireless domain [8]–[13]. Also, blind constant modulus based CFO estimation schemes have been proposed in the wireless domain [14]–[18]. A modest kurtosis-type criterion is used for CFO estimation in [19]. The approach exploits the variance of interfering subcarriers and the Gaussianity of a random sequence, to derive the proposed cost function. The kurtosis-type estimator is built on the idea that if the CFO has not been totally compensated, the distribution of the post-DFT sequence is closer to Gaussian than when the CFO has been perfectly compensated. Another blind CFO estimation method is presented in [20], which exploits the smoothing power spectrum. In the approach, the cost function is based on the similarity of the frequency response between two adjacent subcarriers. The method can be utilized for both constant modulus (CM) and non-CM signaling. In [21], an improved method is proposed to address the drawbacks of the kurtosis-based estimators. In this method, CFO estimation is achieved by minimizing an objective function based on the power difference between received constellations using CM signaling. The performance of this method is shown to be superior to the method in [19], but with similar computational complexity. However, these schemes suffer from gross degradation under severe channel conditions [22]. The power difference estimator (PDE-T), which is a hybrid time-frequency domain estimator, is proposed in [23]. The cost function minimizes the power difference between subcarriers in two consecutive OFDM symbols and assumes a slowly varying channel in the time domain. Recently, the amplitude difference estimator (ADE-T) method [24], similar to the PDE-T, was proposed. In the ADE-T method, the CFO estimate is obtained by deriving a cost function based on the magnitude of the received sequence with two similar subcarriers having the same indexes. The cost function, which is derived by exploiting the channel coherence in time, has quasi-regular shape and can be further approximated to achieve a closed-form CFO estimation [24]. The ADE-T method is less sensitive to noise and instabilities along the channel. In [25], a CFO estimator is presented, whose cost function is derived based on the powers of non-diagonal elements of covariance matrix. The information on the CFO is embedded in the covariance matrix of the received sequence. The method achieves CFO estimation by minimizing the total off-diagonal power induced by the inter-channel interference (ICI) in the frequency domain. However, the implementation of this scheme requires large number of OFDM symbols, which causes processing delay. Also in [22], a circularly shifted covariance (CSC) matrix method was proposed, where estimation is achieved by forcing the out-of-band elements of the matrix to zero. The covariance matrix obtained has a banded structure and the CFO is estimated by minimizing the power of the elements that are outside the band. The cost function of this method is dependent on the property of the channel matrix, which makes it susceptible to impairments along the channel. This approach gives an enhanced performance and requires a reduced number of OFDM symbols compared to the method in [25]. But it becomes unstable and the efficiency degrades as the out-of-band elements of the covariance matrix decreases under severe channel conditions.

Thus, a completely blind low-complexity CFO estimation approach for constant modulus constellations CO-OFDM systems is proposed, with a cost function similar to one utilized for blind channel equalization in [26]. The performance of the proposed approach is analyzed and compared with prominent existing methods in a practical optical system with an uncompensated fiber link in terms of the mean square error (MSE), the bit-error-rate (BER) and the convergence speed.

The main contributions in this paper therefore include:

1. The investigation of the performance of prominent constant modulus based blind estimation schemes, which have hitherto not been implemented and analyzed in the optical domain. This paper therefore investigates how these existing constant modulus schemes perform in the optical scenario, with fiber dispersion.
2. A blind low-complexity CFO estimator is proposed and compared with existing constant modulus schemes. In the existing methods, the cost functions are totally dependent on the channel characteristics, where it is usually assumed that the channel slowly varies over consecutive symbols. However, the proposed estimator is independent of this general assumption. This in fact makes the proposed method robust against channel impairments.
3. In order to achieve low-complexity, the proposed cost function is derived and approximated as a cosine function, so that the CFO estimate is obtained using the curve fitting method, where only three trial values are required.
4. The derived closed-form expression ensures a low complexity similar to the existing schemes while offering a superior overall performance.

The rest of this paper is organized as follows. Section 2 presents the CO-OFDM system model. The CO-OFDM system employed is modeled in the presence of CFO with dispersion along the fiber link. Section 3 discusses the existing estimation schemes in the literature. In Section 4, the proposed blind CFO estimation is derived and discussed. Section 5 presents the simulation results for the proposed estimator. The performance of the proposed method in comparison with prominent existing methods is also discussed. Finally, Section 6 gives the conclusion.

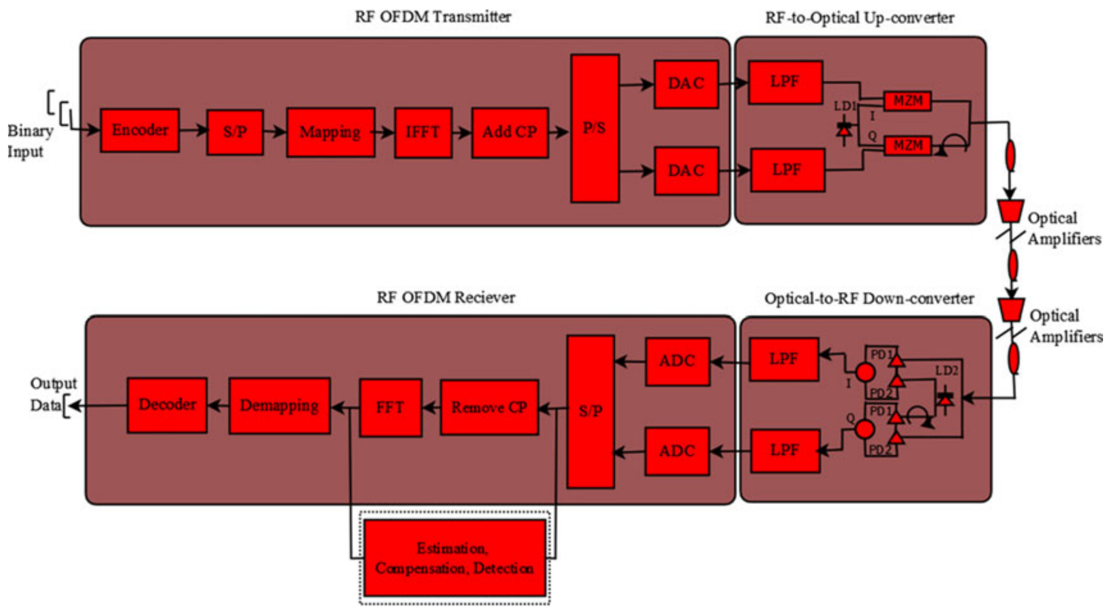

Fig. 1. The block diagram of a typical CO-OFDM transceiver.

2. System Model

Considering a CO-OFDM system as described in Fig. 1, with N subcarrier, the transmitted nth OFDM signal is given by $\mathbf{X}_n = [X_n(0), X_n(1) \ldots X_n(N-1)]^T$, where data symbols $\mathbf{X}_n$ are uniformly drawn from a CM constellation. The time domain nth OFDM $\mathbf{d}_n = [d_n(0), d_n(1) \ldots d_n(N-1)]^T$ is given as:

$$d_n(k) = \frac{1}{\sqrt{N}} \sum_{m=0}^{N-1} X_n(m) e^{\frac{j2\pi mk}{N}}, \tag{1}$$

which can be expressed as $\mathbf{d}_n = W\mathbf{X}_n$, where $\mathbf{W}$ is the normalized N $\times$ N IDFT matrix. The elements of $\mathbf{W}$ are defined by $W_{\text{mk}} = \frac{1}{\sqrt{N}} e^{\frac{j2\pi mk}{N}}$, where m and k denote the row and column indices respectively and vary from 0 to N $-$ 1.

The signal then passes through the optical channel and the received sequence is given as:

$$\mathbf{y}_n = e^{j\frac{2\pi\varepsilon n}{N}(N+N_g)} \mathbf{C}(\varepsilon) \mathbf{W}\mathbf{Z}_n\mathbf{X}_n + \mathbf{g}_n \tag{2}$$

where $\varepsilon \in (-0.5,\ 0.5)$ is the CFO, and the accumulated phase shift caused by the CFO on the time domain samples is given by $\boldsymbol{C}(\varepsilon) = diag([e^{j\frac{2\pi\varepsilon}{N}\times 0}, e^{j\frac{2\pi\varepsilon}{N}\times 1} \ldots . e^{j\frac{2\pi\varepsilon}{N}\times(N-1)}]^T)$, whose nth OFDM symbol is given as $e^{j\frac{2\pi\varepsilon n}{N}(N+N_g)}$. Also, N_g represents the guard interval, while $\boldsymbol{Z_n}$ is the holistic channel impulse response of the fiber link encompassing the polarization mode dispersion, group velocity dispersion and other polarization dependent losses [2], [27].

The received sequence y_n is multiplied by $\boldsymbol{C}^*(\bar{\varepsilon})$ for CFO compensation using a trial value of CFO $\bar{\varepsilon}$, then fed to the DFT to obtain

$$\mathbf{Y}_n = \mathbf{W}^{\text{H}}\mathbf{C}^*(\bar{\varepsilon})\,\mathbf{y}_{\text{n}}, \tag{3}$$

where $\mathbf{W}^{-1} = \mathbf{W}^{\text{H}}$ is a unitary matrix. Also, the mth element of $\mathbf{Y}_n$ can be expressed as:

$$Y_n(m) = \frac{1}{\sqrt{N}} \sum_{k=0}^{N-1} y_n(k) e^{-j\frac{2\pi k}{N}(m-\bar{\varepsilon})}, \tag{4}$$

where

$$y_n(k) = \frac{e^{j\frac{2\pi\varepsilon n}{N}(N+N_g)}}{\sqrt{N}} \sum_{m=0}^{N-1} X_n(m) Z_n(m) e^{j\frac{2\pi k}{N}(m-\varepsilon)} + g_n(k). \tag{5}$$

Also, as stated earlier, $Z_n(m)$ is the comprehensive channel impulse response of the fiber link, which includes the group velocity dispersion (GVD). The GVD $\varnothing_n(m)$, which is primarily a phase shift due to distortion in the fiber-link, can be expressed mathematically as:

$$\varnothing_n(m) = \pi.c.q_f.\frac{f_m^2}{f_o^2}, \tag{6}$$

where q_f denotes the chromatic dispersion in the link, f_m is the frequency for the mth subcarrier while f_o is the center optical frequency.

From the above expressions, the received signal can be analyzed and the CFO can be evaluated.

3. Existing Blind CFO Estimation Methods

Various schemes have been proposed and implemented in the literature to combat the degrading impact of CFO on the performance of CO-OFDM systems. However, most of the existing schemes are pilot-based or data-aided schemes, which results in an increased overhead in the optical system [8]–[11]. Maximum likelihood based methods have been utilized but these also come with the associated computational burden [12], [13]. In contrast to the methods mentioned above, the blind estimation schemes offer a more efficient approach. Various blind estimation schemes have been proposed in the field of wireless communications. The prominent among the existing blind estimation methods are therefore examined and derived for CFO estimation in the optical scenario.

3.1 The Power Difference Estimator (PDE)

In [21], a blind CFO estimation scheme for OFDM systems with CM signaling is proposed. The scheme, which is referred to as PDE-F, is based on the assumption that the channel frequency response is approximately the same for two neighboring subcarriers. Based on this assumption, a cost function is derived and utilized for blind CFO estimation in a closed-form. Also, it is assumed that the CFO has been perfectly compensated before the DFT stage, thus the DFT output is considered to be without ICI. In [23], a similar method, which is referred to as the PDE-T scheme, is implemented for CFO estimation in OFDM systems. The method, which is a hybrid time-frequency-domain estimator, assumes the channel frequency response varies slowly in the time domain. Therefore the resulting sequence after DFT in a noise-free case is described as:

$$\mathbf{Y}_n|_{\bar{\varepsilon}=\varepsilon} = \mathbf{Z}_n\mathbf{X}_n. \tag{7}$$

Considering the case of CM signaling, the squared amplitude of the resulting sequence after DFT is taken as the squared amplitude of the optical channel frequency response, which is expressed as:

$$\left| Y_n(m)|_{\bar{\varepsilon}=\varepsilon} \right|^2 = |Z_n(m)|^2. \tag{8}$$

Also assuming that the channel response slowly changes in the frequency domain, then consecutive subcarriers have equal power. Thus,

$$\left| Y_n(m)|_{\bar{\varepsilon}=\varepsilon} \right|^2 \approx \left| Y_n(m-1)|_{\bar{\varepsilon}=\varepsilon} \right|^2. \tag{9}$$

Based on the above expressions, a cost function is formulated, and the CFO estimation is achieved by the expression

$$\hat{\varepsilon} = arg \min_{\bar{\varepsilon}\in(-0.5,\ 0.5)} J_{P_f}(\bar{\varepsilon}), \tag{10}$$

where

$$J_{P_f}(\bar{\varepsilon}) = \sum_{n=0}^{M-1} \sum_{m=0}^{N-1} \left(|Y_n(m)|^2 - |Y_n(m-1)|^2 \right)^2, \tag{11}$$

where the assumption in (9) is valid for all subcarriers. The CFO is assumed constant over M consecutive OFDM symbols and (11) can be expanded and further approximated as [21], [23]:

$$J_{P_f}(\bar{\varepsilon}) \approx A \cos[2\pi(\varepsilon - \bar{\varepsilon})] + C, \tag{12}$$

where A and C are constants with real values, independent of $\bar{\varepsilon}$ but dependent on the optical channel and symbol realization as detailed in [23]. Also, the cost function achieves its minimum at $\bar{\varepsilon} = \varepsilon$.

The curve-fitting method as described in [19], can be utilized for the minimization process since the cost-function is sinusoidal.

3.2 The Amplitude Difference Estimator (ADE-T)

In [24], another cost function was proposed for blind CFO estimation. The method is achieved by using a cost-function based on the sum of the products of the signal amplitudes on each pair of equivalent subcarriers from consecutive OFDM symbols. Considering the case of CM signaling while assuming a slow time-varying channel response, then

$$|Y_n(m)| \approx |Y_n(m-1)|, \tag{13}$$

and the cost-function is described as

$$J_{A_f}(\bar{\varepsilon}) = \sum_{n=0}^{M-1} \sum_{m=0}^{N-1} (|Y_n(m)| - |Y_n(m-1)|)^2. \tag{14}$$

Thus CFO can be estimated by minimizing the cost-function as

$$\hat{\varepsilon} = \arg \min_{\bar{\varepsilon} \in (-0.5,\ 0.5)} J_{A_f}(\bar{\varepsilon}). \tag{15}$$

3.3 The Circularly Shifted Covariance Method

A circularly shifted covariance method is proposed in [22], with the cost-function based on the covariance matrix obtained through circularly shifted OFDM blocks in the time-domain. This method is achieved by obtaining covariance matrix using few numbers of OFDM symbols while forcing out-of-band elements of the matrix to zero. Thus the sample covariance matrix is generated using the expression

$$R_n(\bar{\varepsilon}) = \frac{1}{N} \mathbf{F}_n \mathbf{F}_n{}^H, \tag{16}$$

where $\mathbf{F}_n$ is obtained with circularly shifted received sequence $Y_n(m)$ constituting its columns. This method is highly dependent on the fiber channel length L as detailed in [22].

4. Proposed Constant Modulus Based Blind CFO Estimation

As summarized above, the various methods are based on a common assumption that the channel varies slowly over consecutive symbols. However, this assumption may not hold in the case where the CFO is not perfectly estimated. The approach in [22] deviates from this assumption, but is highly dependent on the channel characteristics due to the out-of-band elements of the covariance matrix. Therefore, we present a cost-function, which is independent of the above assumptions and drawbacks. The proposed approach is achieved based on the approximation of the constant modulus cost-function similar to the one utilized for blind channel equalization in [25]. Hence, the following cost-function is proposed for the blind CFO estimation

$$J_{\mathrm{GA}}(\bar{\varepsilon}) = E\left\{\left(|Y_n(m)|^2 - R\right)^2\right\}, \tag{17}$$

where R is a constant chosen to guarantee the minimization of $J_{\mathrm{GA}}(\bar{\varepsilon})$. Thus, $J_{\mathrm{GA}}(\bar{\varepsilon})$ should be minimized with respect to the trial value of ε, denoted as $\bar{\varepsilon}$, and the CFO estimate is obtained by

$$\hat{\varepsilon} = \arg \min_{\bar{\varepsilon} \in (-0.5,\ 0.5)} J_{\mathrm{GA}}(\bar{\varepsilon}). \tag{18}$$

In order to achieve a closed-form CFO estimation, the cost-function in (17) is therefore approximated as (see Appendix I)

$$J_{\mathrm{GA}}(\bar{\varepsilon}) \approx \mathring{A} \cos(2\pi\partial) + B, \tag{19}$$

where $\mathring{A}$ and B are real values and independent of $\bar{\varepsilon}$ and ε, while $\partial = \varepsilon - \bar{\varepsilon}$.

Therefore, approximating the cost-function in (17) eliminates the need for the exhaustive search and the CFO estimate can be obtained using the curve-fitting method as described in [19]. The minimum of the approximated cost-function in (19) is evaluated at three distinct points, $\bar{\varepsilon} = 0,\ 0.25,\ \text{and}\ -0.25$. Thus, the estimate of the CFO is obtained by

$$\hat{\varepsilon} = \begin{cases} \frac{1}{2\pi}\tan^{-1}\left(b/a\right) & \text{for } a \geq 0 \\ \frac{1}{2\pi}\tan^{-1}\left(b/a\right) + \frac{1}{2} & \text{for } a < 0 \text{ and } b \geq 0, \\ \frac{1}{2\pi}\tan^{-1}\left(b/a\right) - \frac{1}{2} & \text{for } a < 0 \text{ and } b \leq 0 \end{cases} \tag{20}$$

where $a = \{(1/2)(J_{\mathrm{GA}}(\bar{\varepsilon} = 0.25) + J_{\mathrm{GA}}(\bar{\varepsilon} = -0.25)) - J_{\mathrm{GA}}(\bar{\varepsilon} = 0)\}$, and $b = \{(1/2)(J_{\mathrm{GA}}(\bar{\varepsilon} = 0.25) + J_{\mathrm{GA}}(\bar{\varepsilon} = -0.25))\}$.

The above method thus enables the closed-form estimation of the CFO and reduces the computational complexity of the proposed cost-function.

5. Simulation and Discussion

The quadrature phase-shift keying CO-OFDM system is utilized in the computer simulations to investigate and analyze the performance of the various estimation methods. The CO-OFDM system is based on a central wavelength of 1550 nm, with $N = 64$ subcarriers and cyclic prefix of length 16. The sampling rate of 10 GS/s is utilized and the OFDM duration is 7.2 ns with fiber link loss coefficient of 0.2 dB/ km. Existing schemes are compared with the proposed method in terms of mean square error and bit-error rate using Monte Carlo simulations while the normalized CFO ε, is assumed to be uniformly distributed in the range $(-0.5,\ 0.5)$. The optical system model is fully implemented mimicking a practical scenario with prevailing fiber-link dispersions, whose effects on optical links are detailed in [2], [27]. The fiber dispersion is 17 ps/km/nm while the erbium-doped fiber amplifier is of 16 dB gain with noise figure of 4 dB and the non-linear coefficient of the fiber is 1.32/W/ km.

First an independent analysis is carried out on the performance and the behavior of the proposed cost function. The general attribute and performance is a function of the constant R. The connotation R can be taken as a dispersion coefficient as described in [26], [28], where it becomes dependent on the transmitted data. However, to achieve a blind CFO estimation without the priori knowledge of the transmitted signal, R can be taken as a positive real constant [28]–[30]. To ascertain the value of R at which our proposed method achieve the best performance, we vary the value of R and the MSE plot is shown in Fig. 2. From the plot, the proposed method achieves the best performance at $R = 1.0$ while $R = 0.5$ gives a degraded output. Also in Fig. 3, the MSE is plotted as a function of R at different OSNRs. It can be seen from the plot that $R = 0.9$ offers the best performance at OSNR = 15 dB. However at OSNR = 22 dB, $R = 1.0$ gives the best performance.

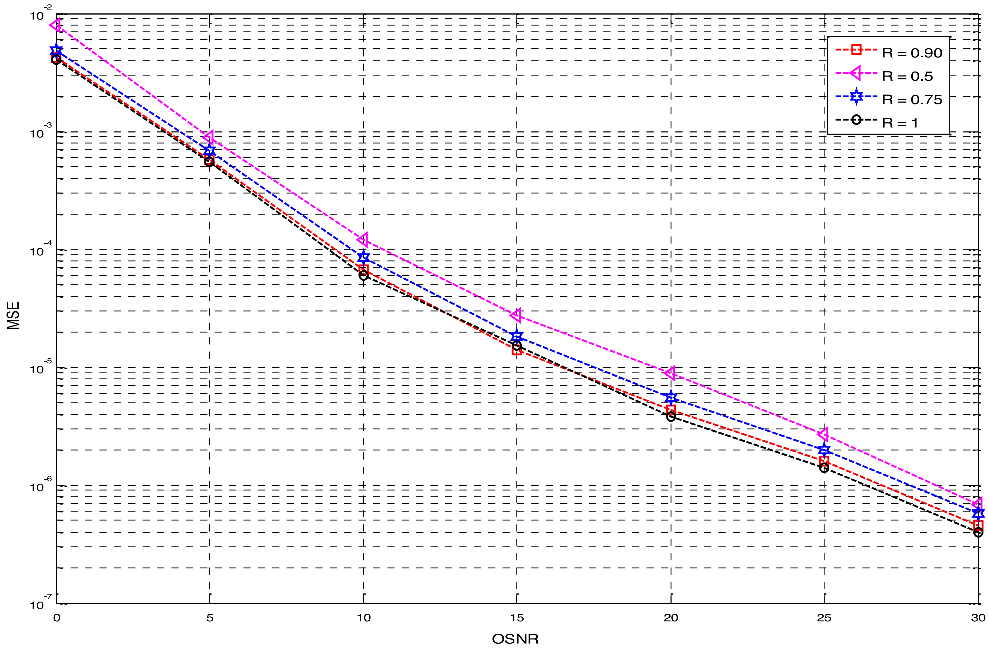

Fig. 2. The MSE versus OSNR plot for different values of R.

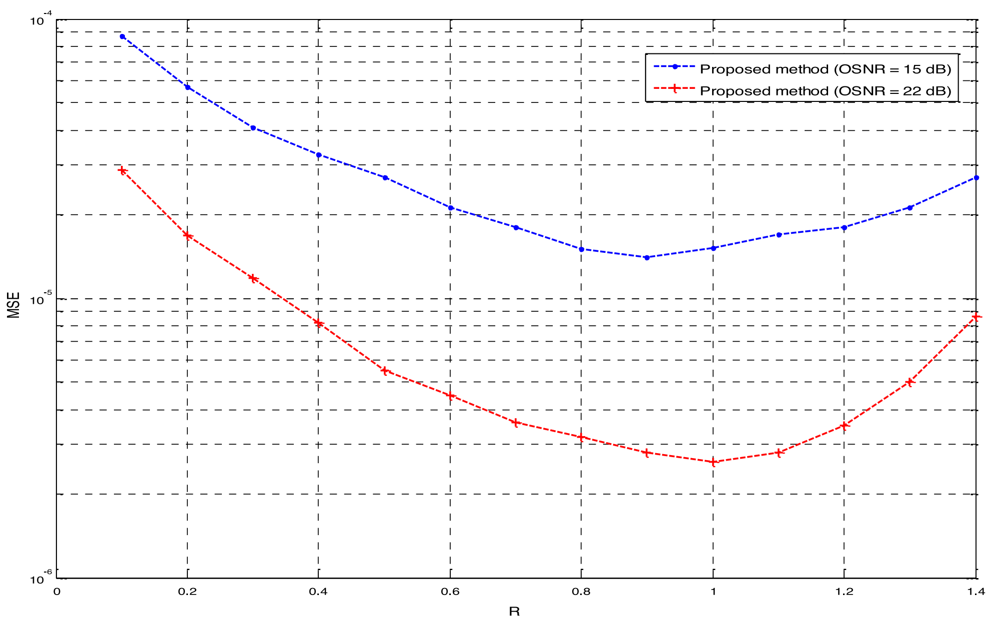

Fig. 3. MSE plot of the proposed method as a function of R (OSNR = 15 dB, 22 dB).

Furthermore, the steady-state characteristics and the convergence speed of the proposed algorithm is investigated at OSNR = 12 dB as shown in Fig. 4. From the MSE plot, it can be seen vividly that the proposed algorithm achieves the fastest convergence at $R = 1.0$. At the value $R = 1.0$, the proposed algorithm already attains steady-state with the number of OFDM symbols at 600. The other values of R achieves steady-state with the number of OFDM symbol at 700 or greater. Hence $R = 1$ is selected as the preferred value for the proposed method since it gives the fastest convergence and an efficient overall performance.

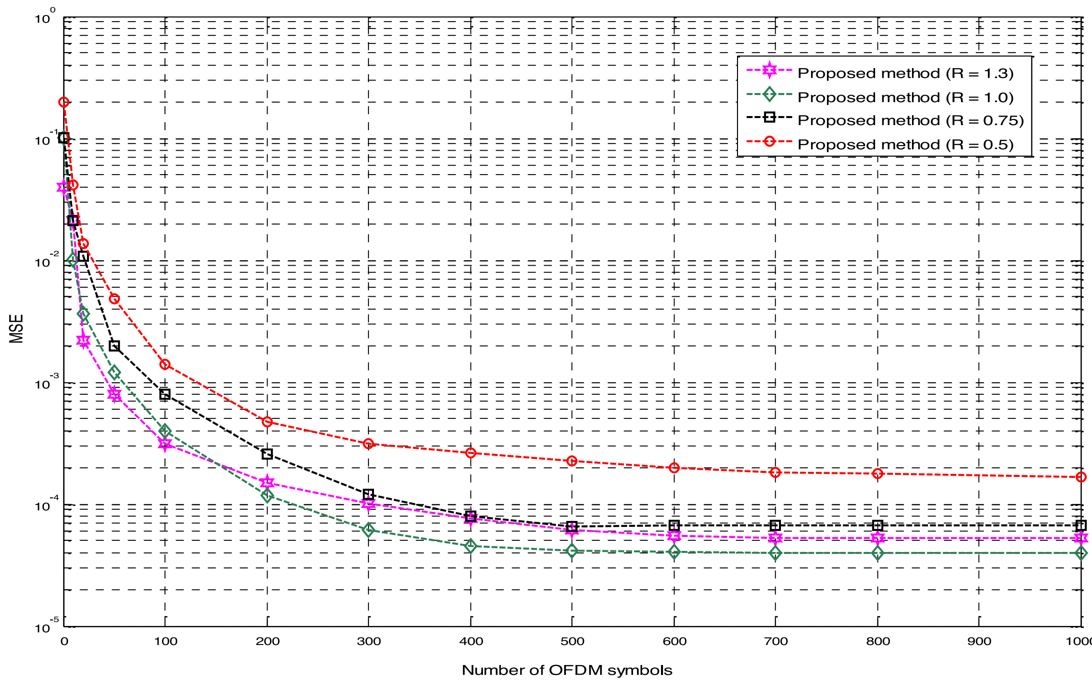

Fig. 4. Convergence of the proposed algorithm for different values of R at OSNR = 12 dB.

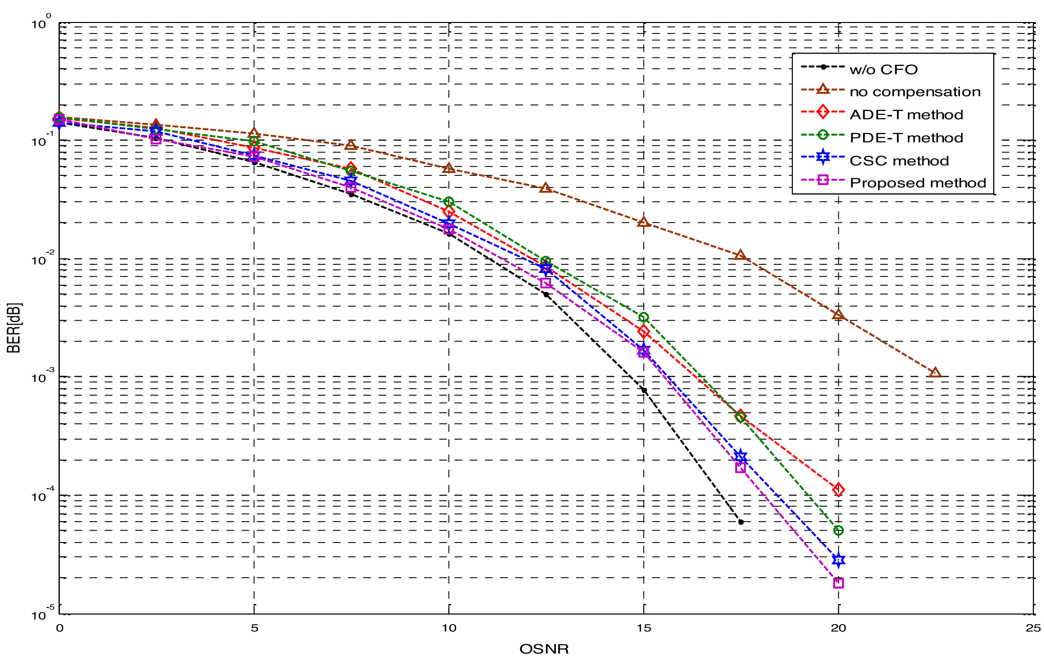

Fig. 5. BER sensitivity for the proposed estimator in comparison with existing methods using a compensated fiber link.

In Fig. 5, the BER performance of existing methods in the literature is compared with the proposed method as implemented in the optical scenario with CFO ε, uniformly distributed in the range $(-0.5,\ 0.5)$. The effectiveness of the algorithms is demonstrated by showing the scenario where there is no CFO in the CO-OFDM system, with the CFO at 0.15. The plot shows the performance of the PDE-T method in [23]. The PDE-T method has been shown to outperform the PDE-F method

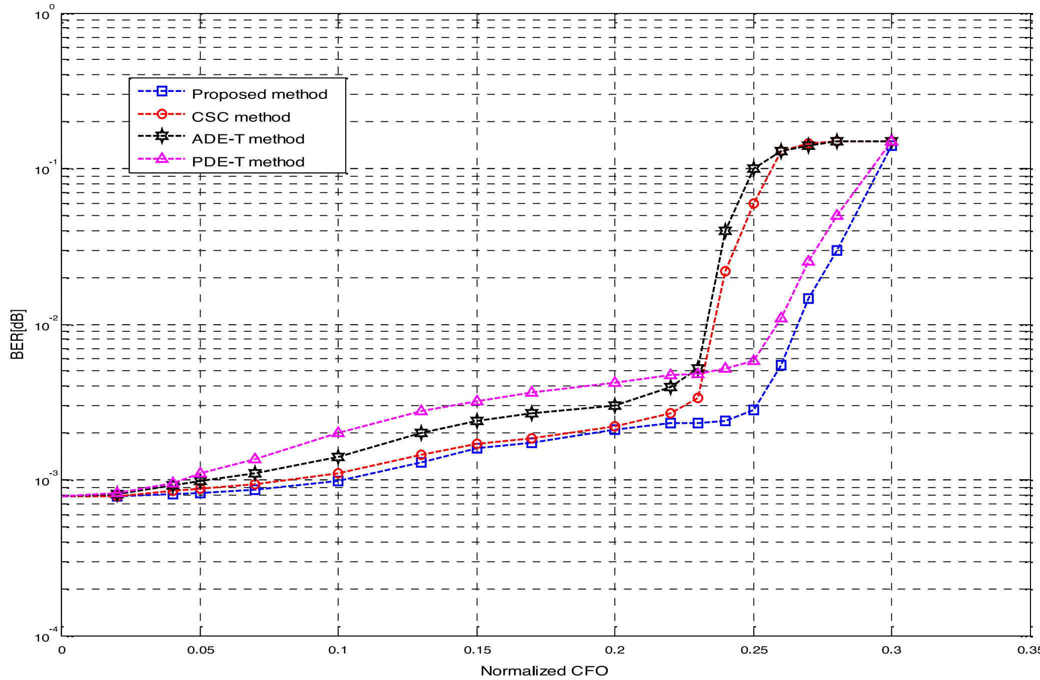

Fig. 6. BER plot for varying values of CFO at OSNR = 15 dB.

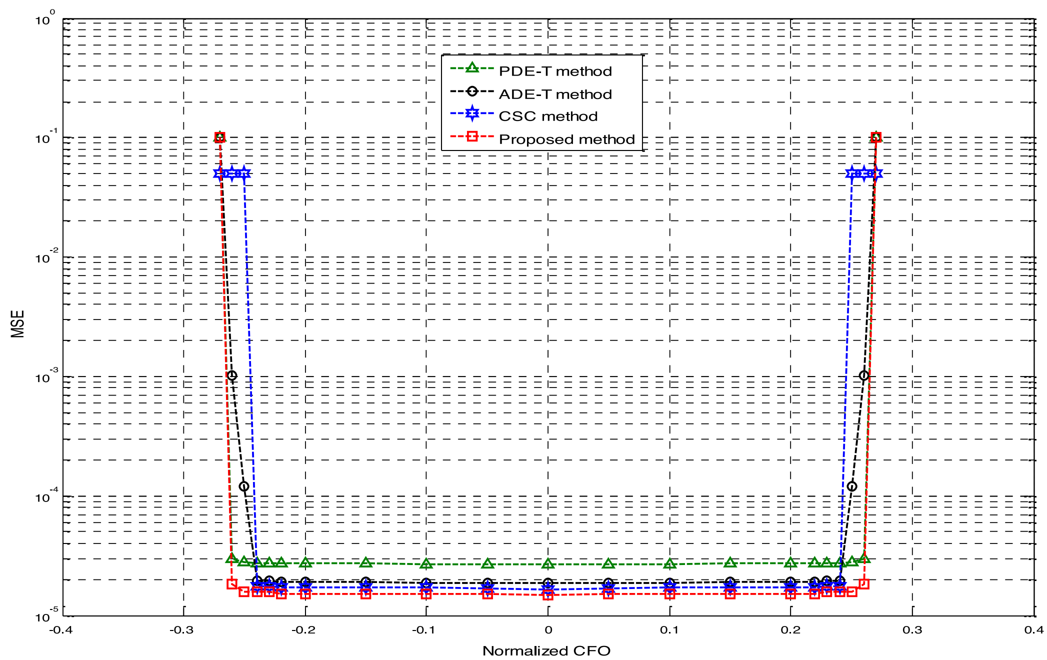

Fig. 7. MSE plot of the various methods showing range limits (OSNR = 15 dB).

[21], therefore it is chosen as one of the methods used for performance comparison. Also, the performance of the CSC method [22] and the ADE-T method are shown in the plot. The BER and the MSE plots with varying values of CFO are shown in Figs. 6 and 7 respectively, at OSNR = 15 dB. It can be seen that the range of the proposed system, like the other methods considered, is limited by the constant modulus constraint. The methods give acceptable performance up to CFO $\leq$ 0.25,

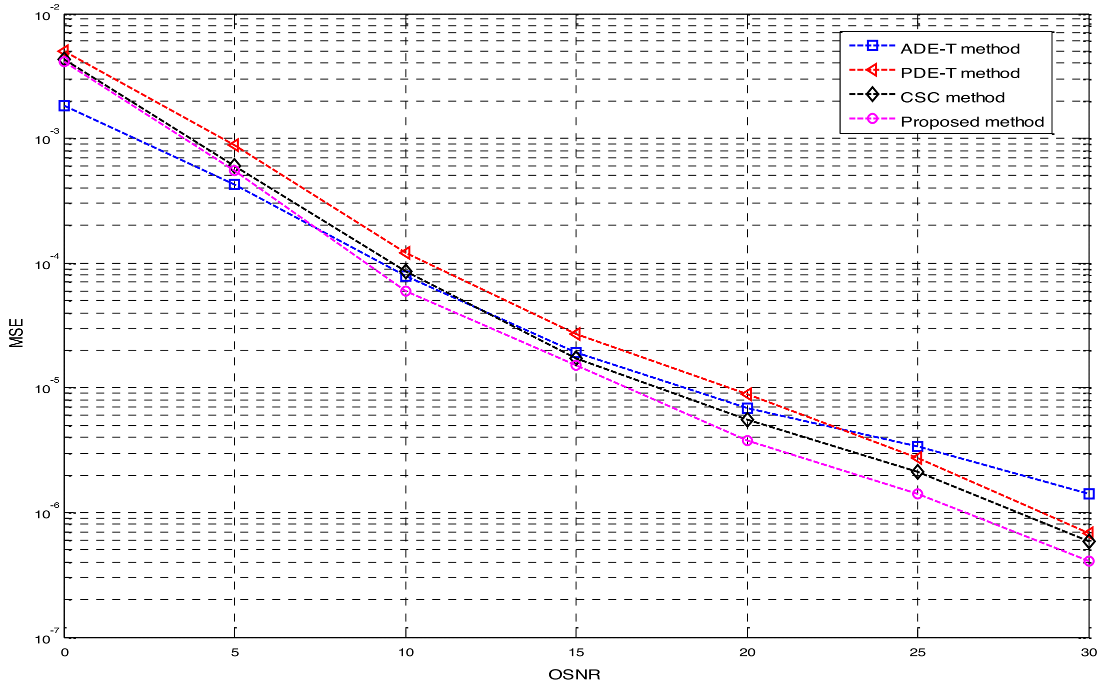

Fig. 8. MSE performance of the various estimation methods using a fiber link of 8 × 80 km spans.

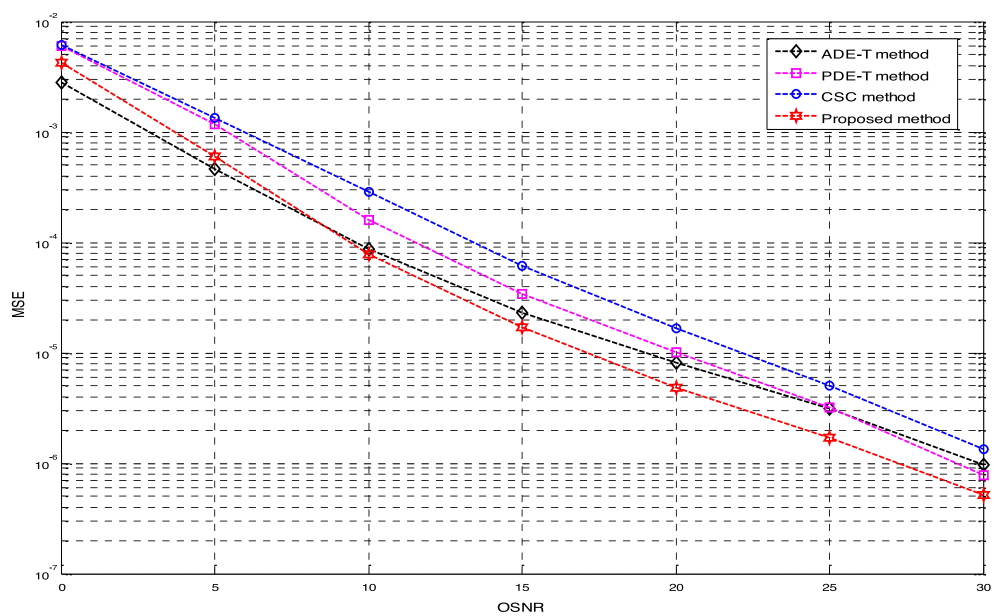

Fig. 9. MSE performance of the various estimation schemes under fiber impairments with total haul 1.8×10^3 km.

beyond which there is fast system degradation. The estimation range can be increased considerably by employing a second stage iterative estimator [31], although this comes with an increased system complexity. The MSE comparison of the various existing methods is compared with the proposed method as shown in Figs. 8 and 9. In Fig. 8, the plot is carried out with the fiber link of eight optically amplified 80 km fiber spans. From the plot, the ADE-T method achieves a superior performance as compared to the other methods including the proposed method at low OSNRs. However, the

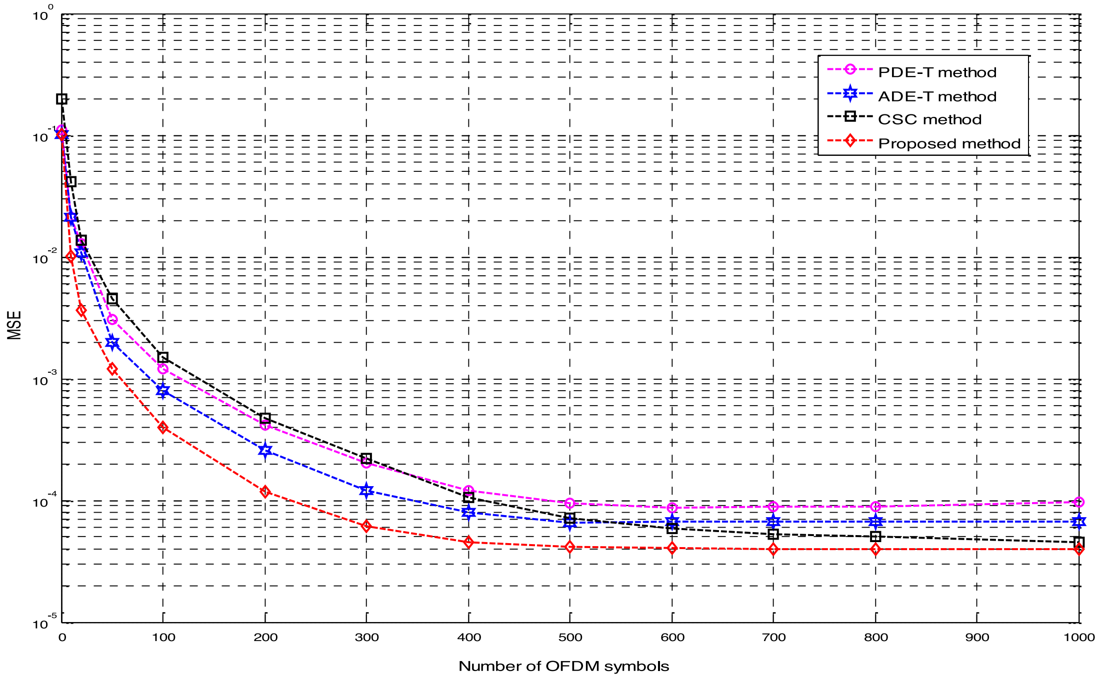

Fig. 10. Steady-state dynamics of the various estimation schemes at OSNR = 12 dB.

performance degrades at high OSNR values, where it is outperformed by the PDE-T method. The CSC method outperforms both the ADE-T and the PDE-T methods at high OSNR values although the ADE-T method offers a superior performance at low OSNR value. The proposed method is outperformed by the ADE-T method at OSNR $<$ 8 dB. Also, the proposed scheme offers a superior performance to the PDE-T and the CSC method for all OSNR values. However, at OSNR values greater than 8 dB, the proposed method effectively gives a better performance. The effectiveness of the various methods is further investigated under an uncompensated fiber-link with a total fiber haul of 1.8×10^3 km. Under this fiber condition, the general performance of the various algorithms degrades as shown in Fig. 9. The plot shows that the CSC method suffers greatly under fiber impairments and its performance becomes inferior to the other blind estimation schemes. The ADE-T out-performs all the other methods at low OSNR values but at increased OSNRs it is outperformed by the PDE-T method and the proposed method.

In Fig. 10, the steady-state dynamics of the estimation schemes under consideration is investigated. As seen from the plot, the proposed method offers a superior performance and achieves the fastest convergence as compared to the other schemes. Also in Fig. 11, the plot shows the behavior of various schemes at OSNR = 15 dB, as the fiber length is increased up to 2000 km. From the graph, the performance of the proposed method suffers slight degradation with increased dispersion and fiber length. However, it offers a stable and efficient performance across the fiber length, compared to the ADE-T method. The plot shows the CSC method as very unstable as compared to the other methods under consideration. The performance of the CSC method is greatly influenced by dispersion and impairments along the fiber-link of length L. In the proposed method, the estimation is performed using the demodulated OFDM symbols from the FFT operations and the total number of multiplications required is of the order $O(N)$ as with the other existing methods (Table 1). Thus, the proposed method achieves a low-complexity, stable and efficient performance, which is shown to be robust against fiber dispersions.

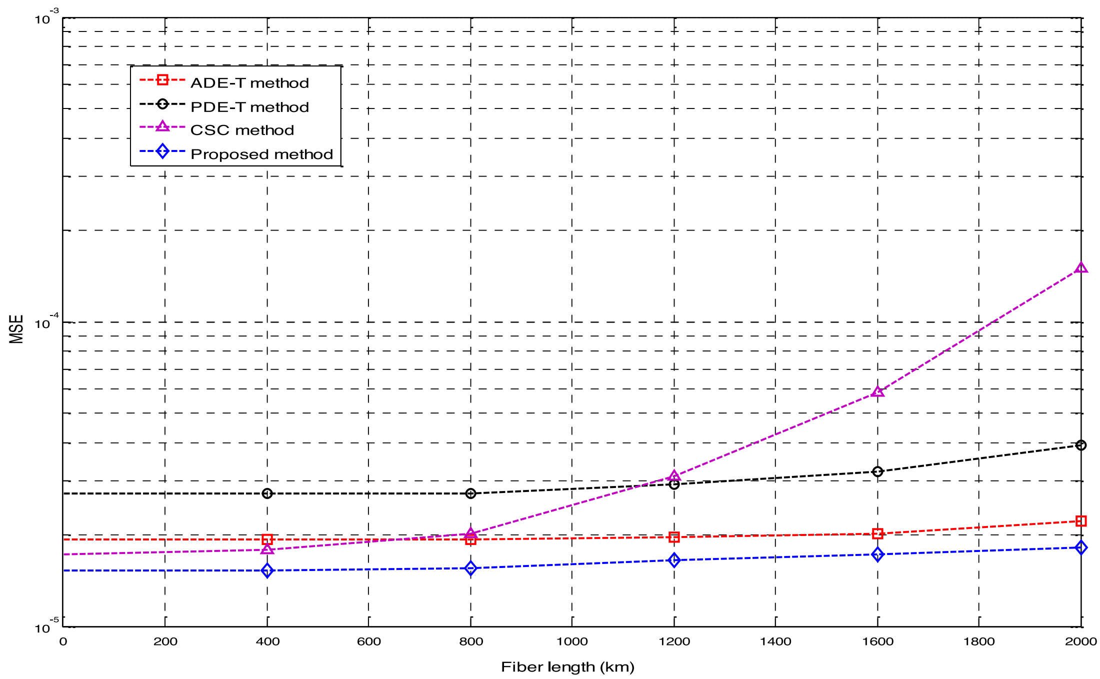

Fig. 11. MSE performance of the various estimation methods as function of the fiber length at OSNR = 15 dB.

TABLE 1
COMPUTATIONAL COMPLEXITY

Algorithm	Multiplications	Addition	Complex multiplications
PDE-T	21N+3	12N	2N
PDE-F	21N+3	12N	2N
CSC	8(N+L)	L+3	2N-4L+2
ADE-T	15N+3	8N	2N
Proposed	18N+3	14N	2N

6. Conclusion

A highly efficient totally blind CFO estimation method has been proposed and implemented for CM-signaling based CO-OFDM systems. First, existing blind estimation schemes in the literature have been adapted, derived and implemented in the optical scenario. The performances of these existing schemes are further compared with the proposed blind CFO estimator. The blind CFO estimation method has been proven through analysis and simulation to achieve a superior performance as compared to the prominent existing blind estimation schemes. The proposed estimator shows high stability and performance in the presence of fiber impairments and dispersion. The proposed cost-function is also approximated as a cosine function, thus the CFO is estimated in close-form using only three trial values. This ensures that the proposed method achieves a reduced computational complexity similar to the existing methods. Therefore, the proposed method offers a balanced and efficient overall performance as compared to existing blind estimation methods.

Appendix I

Expanding (17) gives

$$J_{GA}(\bar{\varepsilon}) = \mathrm{E}\left\{|Y_n(m)|^4\right\} - \mathrm{E}\left\{2R.|Y_n(m)|^2\right\} + R^2, \tag{21}$$

Substituting (5) into (4), while assuming the noise due to the ASE of the optical amplifiers is minimal throughout the fiber-link, gives:

$$Y_n(m) = \frac{e^{j\frac{2\pi\varepsilon n}{N}(N+N_g)}}{N}\sum_{m=0}^{N-1}\hat{B}_n(m)\sum_{k=0}^{N-1}e^{j\frac{2\pi k}{N}(m+\partial-n)}, \tag{22}$$

where $\partial = \varepsilon - \bar{\varepsilon}$, $\hat{B}_n(m) = X_n(m)Z_n(m)$

Substituting (22) into (21) gives,

$$\begin{aligned} J_{GA}(\bar{\varepsilon}) = & E\left\{\left|\frac{e^{j\frac{2\pi\varepsilon n}{N}(N+N_g)}}{N}\sum_{m=0}^{N-1}\hat{B}_n(m)\sum_{k=0}^{N-1}e^{j\frac{2\pi k}{N}(m+\partial-n)}\right|^4\right\} \\ & - E\left\{2R\left|\frac{e^{j\frac{2\pi\varepsilon n}{N}(N+N_g)}}{N}\sum_{m=0}^{N-1}\hat{B}_n(m)\sum_{k=0}^{N-1}e^{j\frac{2\pi k}{N}(m+\partial-n)}\right|^2\right\} + R^2, \end{aligned} \tag{23}$$

Thus,

$$\begin{aligned} |Y_n(m)|^4 = & \frac{1}{N^4}\sum_{m_1,m_2,m_3,m_4=0}^{N-1}\hat{B}_n(m_1)\hat{B}_n^*(m_2)\hat{B}_n(m_3)\hat{B}_n^*(m_4) \\ & \times \sum_{k_1,k_2,k_3,k_4=0}^{N-1}e^{\frac{j2\pi\partial\omega}{N}}e^{j\frac{2\pi}{N}(k_1m_1-k_2m_2+k_3m_3-k_4m_4)} \times \sum_{n=0}^{N-1}e^{\frac{-j2\pi n\omega}{N}}, \end{aligned} \tag{24}$$

where $\omega = k_1 - k_2 + k_3 - k_4$.

Also,

$$\sum_{n=0}^{N-1}e^{\frac{-j2\pi n\omega}{N}} = \begin{cases} N, & \omega = N, 0, -N \\ 0, & \text{otherwise.} \end{cases} \tag{25}$$

Hence (24) can be simplified as shown below where S_{n_1} is a real constant independent of ∂, which is obtained by substituting $\omega = 0$ in (24)

$$\begin{aligned} |Y_n(m)|^4 = & \frac{2}{N^3}\mathcal{R}e\left\{e^{-j2\pi\partial}\sum_{m_1,m_2,m_3,m_4=0}^{N-1}\hat{B}_n(m_1)\hat{B}_n^*(m_2)\hat{B}_n(m_3)\hat{B}_n^*(m_4)\right. \\ & \left.\times\sum_{k_1=0}^{N-1}\sum_{k_2=k_1+1}^{N-1}\sum_{k_3=0}^{k_2-k_1-1}e^{j\frac{2\pi}{N}k_1(m_1-m_4)}e^{-j\frac{2\pi}{N}k_2(m_2-m_4)}e^{j\frac{2\pi}{N}k_3(m_3-m_4)}\right\} + S_{n_1}. \end{aligned} \tag{26}$$

Now define

$$\Phi \triangleq \{m_1, m_2, m_3, m_4\}$$

$$\Phi_1 \triangleq \{\Phi | m_1 = m_2, or\ m_3 = m_4\}$$

$$\Phi_2 \triangleq \{\Phi | m_1 \neq m_2, \text{and } m_3 \neq m_4\},$$

where the indices $m_1, m_2, m_3, m_4 \in \{0, 1, \ldots, N-1\}$, $\Phi = \Phi_1 \cup \Phi_2$, and $\Phi_1 \cup \Phi_2 = \varnothing$. Therefore (26) can be expressed as:

$$|Y_n(m)|^4 = \frac{2}{N^3}\mathcal{R}e\left\{e^{-j2\pi\partial} D_n(\Phi_1)\right\} + \frac{2}{N^3}\mathcal{R}e\left\{e^{-j2\pi\partial} D_n(\Phi_2)\right\} + S_{n_1}, \tag{27}$$

where

$$D_n(\Omega) = \sum_{\substack{m_1, m_2, m_3, m_4 = 0 \\ m_1, m_2, m_3, m_4 \in \Omega}}^{N-1} \hat{B}_n(m_1)\hat{B}_n^*(m_2)\hat{B}_n(m_3)\hat{B}_n^*(m_4)$$
$$\times \sum_{k_1=0}^{N-1}\sum_{k_2=k_1+1}^{N-1}\sum_{k_3=0}^{k_2-k_1-1} e^{j\frac{2\pi}{N}k_1(m_1-m_4)} e^{-j\frac{2\pi}{N}k_2(m_2-m_4)} e^{j\frac{2\pi}{N}k_3(m_3-m_4)},$$

and $\Omega \in \{\Phi_1, \Phi_2\}$.

Also,

$$|Y_n(m)|^2 = \frac{1}{N^2}\sum_{m_1,m_2=0}^{N-1}\hat{B}_n(m_1)\hat{B}_n^*(m_2)$$
$$\times \sum_{k_1,k_2=0}^{N-1} e^{\frac{j2\pi\partial\omega}{N}} e^{j\frac{2\pi}{N}(k_1m_1-k_2m_2)} \times \sum_{n=0}^{N-1} e^{\frac{-j2\pi n\omega}{N}}. \tag{28}$$

If $\omega = 0$, and using the condition in (25), then (28) becomes

$$|Y_n(m)|^2 = \frac{1}{N^2}\sum_{m_1,m_2=0}^{N-1}\hat{B}_n(m_1)\hat{B}_n^*(m_2)\sum_{k=0}^{N-1} e^{\frac{j2\pi kq}{N}}, \tag{29}$$

where $q = m_1 - m_2$ and $k_1 = k_2 = k$. The expression in (29) is therefore independent of $\bar{\varepsilon}$, as well as the CFO ε, and is denoted as S_{n_2}.

Therefore substituting (29), (27) into the cost function in (17) gives:

$$\begin{aligned} J_{GA}(\bar{\varepsilon}) &= \frac{2}{N^3}\mathcal{R}e\left\{e^{-j2\pi\partial} D_n(\Phi_1)\right\} + \frac{2}{N^3}\mathcal{R}e\left\{e^{-j2\pi\partial} D_n(\Phi_2)\right\} + S_{n_1} - \frac{2R}{N}.S_{n_2} + R^2, \\ &= \frac{2}{N^3}\mathcal{R}e\left\{e^{-j2\pi\partial}((D_n(\Phi_1) + D_n(\Phi_2))\right\} + S_{n_1} - \frac{2R}{N}.S_{n_2} + R^2, \end{aligned} \tag{30}$$

and (30) can be reduced to $J_{GA}(\bar{\varepsilon}) \approx \mathring{A}\cos(2\pi\partial) + B$, where $\mathring{A} = -2(D_n(\Phi_1) + D_n(\Phi_2))/N^3$, while $B = S_{n_1} - \frac{2R}{N}.S_{n_2} + R^2$.

References

[1] A. Barbieri, G. Colavolpe, T. Foggi, E. Forestieri, and G. Prati, "OFDM vs. single-carrier transmission for 100 Gbps optical communication," *J. Lightw. Technol.*, vol. 28, no. 17, pp. 2537–2551, Sep. 2010.

[2] W. Shieh and I. Djordjevic, *OFDM for Optical Communications*. New York, NY, USA: Academic, 2010.

[3] I. B. Djordjevic and B. Vasic, "Orthogonal frequency division multiplexing for high-speed optical transmission," *Opt. Exp.*, vol. 14, pp. 3767–3775, 2006.

[4] G. Shen and M. Zukerman, "Spectrum-efficient and agile CO-OFDM optical transport network: Architecture, design, and operation," *IEEE Commun. Mag.*, vol. 50, no. 5, pp. 82–89, May 2012.

[5] W. Shieh and C. Athaudage, "Coherent optical orthogonal frequency division multiplexing," *Electron. Lett.*, vol. 42, pp. 587–589, 2006.

[6] M. Nazarathy *et al.*, "Phased-array cancellation of nonlinear FWM in coherent OFDM dispersive multi-span links," *Opt. Exp.*, vol. 16, pp. 15777–15810, 2008.

[7] M. G. Taylor, "Coherent detection method using DSP for demodulation of signal and subsequent equalization of propagation impairments," *IEEE Photon. Technol. Lett.*, vol. 16, no. 2, pp. 674–676, Feb. 2004.

[8] Y. Liu, Y. Peng, S. Wang, and Z. Chen, "Improved FFT-based frequency offset estimation algorithm for coherent optical systems," *IEEE Photon. Technol. Lett.*, vol. 26, no. 6, pp. 613–616, Mar. 2014.
[9] D. Huang, T. H. Cheng, and C. Yu, "Accurate two-stage frequency offset estimation for coherent optical systems," *IEEE Photon. Technol. Lett.*, vol. 25, no. 2, pp. 179–182, Jan. 2013.
[10] X. Zhou, X. Chen, and K. Long, "Wide-range frequency offset estimation algorithm for optical coherent systems using training sequence," *IEEE Photon. Technol. Lett.*, vol. 24, no. 1, pp. 82–84, Jan. 2012.
[11] Y. Chen, S. Adhikari, N. Hanik, and S. L. Jansen, "Pilot-aided sampling frequency offset compensation coherent optical OFDM," in *Proc. Opt. Fiber Commun. Conf. Expo. Nat. Fiber Opt. Eng. Conf.*, 2012, pp. 1–3.
[12] W. Shieh, "Maximum-likelihood phase and channel estimation for coherent optical OFDM," *IEEE Photon. Technol. Lett.*, vol. 20, no. 8, pp. 605–607, Apr. 2008.
[13] Y.-H. Kim and J.-H. Lee, "Joint maximum likelihood estimation of carrier and sampling frequency offsets for OFDM systems," *IEEE Trans. Broadcast.*, vol. 57, no. 2, pp. 277–283, Jun. 2011.
[14] M. Luise, M. Marselli, and R. Reggiannini, "Low-complexity blind carrier frequency recovery for OFDM signals over frequency-selective radio channels," *IEEE Trans. Commun.*, vol. 50, no. 7, pp. 1182–1188, Jul. 2002.
[15] W. Chin and S. Chen, "A blind synchronizer for OFDM systems based on SINR maximization in multipath fading channels," *IEEE Trans. Veh. Technol.*, vol. 58, no. 2, pp. 625–635, Feb. 2009.
[16] X. Zeng and A. Ghrayeb, "CFO estimation schemes for differential OFDM systems," *IEEE Trans. Wireless Commun.*, vol. 8, no. 1, pp. 124–129, Jan. 2009.
[17] M. Ghogho and A. Swami, "Blind frequency-offset estimator for OFDM systems transmitting constant-modulus symbols," *IEEE Commun. Lett.*, vol. 6, no. 8, pp. 343–345, Aug. 2002.
[18] A. Al-Dweik, A. Hazmi, S. Younis, B. Sharif, and C. Tsimenidis, "Blind iterative frequency offset estimator for orthogonal frequency division multiplexing systems," *IET Commun.*, vol. 4, no. 16, pp. 2008–2019, Nov. 2010.
[19] Y. Yao and G. B. Giannakis, "Blind carrier frequency offset estimation in SISO, MIMO, and multiuser OFDM systems," *IEEE Trans. Commun.*, vol. 53, no. 1, pp. 173–183, Jan. 2005.
[20] L. Wu, X. D Zhang, P. S. Li, and Y. T. Su, "A blind CFO estimator based on smoothing power spectrum for OFDM systems," *IEEE Trans. Commun.*, vol. 57, no. 7, pp. 1924–1927, Jul. 2009.
[21] X. N. Zeng and A. Ghrayeb, "A blind carrier frequency offset estimation scheme for OFDM systems with constant modulus signaling," *IEEE Trans. Commun.*, vol. 56, no. 7, pp. 1032–1037, Jul. 2008.
[22] J.-H. Oh, J.-G. Kim, and J.-T. Lim, "Blind carrier frequency offset estimation for OFDM systems with constant modulus constellations," *IEEE Commun. Lett.*, vol. 15, no. 9, pp. 971–973, Sep. 2011.
[23] A. Al-Dweik, A. Hazmi, S. Younis, B. Sharif, and C. Tsimenidis, "Carrier frequency offset estimation for OFDM systems over mobile radio channels," *IEEE Trans. Veh. Commun.*, vol. 59, no. 2, pp. 974–979, Feb. 2010.
[24] S. Lmai, A. Bourre, C. Laot, and S. Houcke, "An efficient blind estimation of carrier frequency offset in OFDM systems," *IEEE Trans. Veh. Commun.*, vol. 63, no. 4, pp. 1945–1950, May 2014.
[25] T. Roman, S. Visuri, and V. Koivunen, "Blind frequency synchronization in OFDM via diagonality criterion," *IEEE Trans. Signal Process.*, vol. 54, no. 8, pp. 3125–3135, Aug. 2006.
[26] D. N. Godard, "Self-recovering equalization and carrier tracking in two dimensional data communication systems," *IEEE Trans. Commun.*, vol. COM-28, no. 11, pp. 1867–1875, Nov. 1980.
[27] N. Rossi, P. Serena, and A. Bononi, "Polarization-dependent loss impact on coherent optical systems in the presence of fiber nonlinearity," *IEEE Photon. Technol. Lett.*, vol. 26, no. 4, pp. 334–337, Feb. 2014.
[28] J. Whitehead and F. Takawira, "Performance analysis of the linearly constrained constant modulus algorithm-based multiuser detector," *IEEE Trans. Signal Process.*, vol. 53, no. 2, pp. 643–653, Feb. 2005.
[29] J. Du, Q. Peng, and H. Zhang, "Adaptive blind channel identification and equalization for OFDM-MIMO wireless communication systems," in *Proc. 14th IEEE Pers., Indoor Mobile Radio Commun.*, 2003, pp. 2078–2083.
[30] Y. Chen, T. Le-Ngoc, B. Champagne, and C. Xu, "Recursive least squares constant modulus algorithm for blind adaptive array," *IEEE Trans. Signal Process.*, vol. 52, no. 5, pp. 1452–1456, May 2004.
[31] A. Al-Dweik, A. Hazmi, S. Younis, B. Sharif, and C. Tsimenidis, "Blind iterative frequency offset estimator for orthogonal frequency division multiplexing systems," *IET Commun.*, vol. 4, no. 16, pp. 2008–2019, 2010.