Blind Adaptive Low-Complexity Time-domain Equalizer for 100 Gb/s Direct-Detection Optical OFDM Systems Over Long-Reach SSMF

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***Abstract*-In this paper, we report a low complexity blind adaptive time-domain equalizer (TEQ) to reduce the cyclic prefix (CP) length for 100 Gb/s direct detection optical orthogonal frequency division multiplexing (DDO-OFDM) system combined with wavelength division multiplexing (WDM) system over 400 km standard single mode fiber (SSMF) length. We show that a significant complexity reduction can be achieved by designing a modified cost function of multicarrier equalization based on orthogonality restoration (MERO) algorithm on one hand, and on deploying the symmetrical TEQ property on the other hand. Our study shows that, the reduction of the system adaptation complexity can be realized up to 50% when designing and optimizing the stated system parameters. Our low-complexity MERO (LMERO) simulation model exhibits better performance, for a shorter CP length equal to 0.39 % of the OFDM symbol duration, than the existing conventional algorithms in term of bit error rate (BER) versus optical signal to noise ratio (OSNR) with a much lower complexity.**

***Index Terms*—cyclic prefix (CP), DDO-OFDM, LMERO, SSMF, time-domain equalizer (TEQ), WDM.**

I. Introduction

Orthogonal frequency-division multiplexing (OFDM) is a multicarrier modulation (MCM) that has been successfully used in several standards including wireless local area network (WLAN), asymmetric digital subcarrier line (ADSL) and digital video/audio broadcasting (DVB/DAB) [1, 2]. In recent years, OFDM is receiving significant interest from the fiber communication research community for its durability against inter-symbol interference (ISI), when the symbol period for each subcarrier is longer than the delay spread caused by group velocity dispersion (GVD) [3]. In the literature, two fundamentals optical OFDM techniques have been proposed, direct detection optical OFDM (DDO-OFDM) [4] and coherent optical OFDM (CO-OFDM) [5]. The first one has been suggested for the future long-reach passive optical network (LR-PON) [6-8], as it is simpler to be implemented and cost-effective. To avoid the effect of ISI and to maintain the orthogonality of subcarriers, OFDM system inserts a cyclic prefix (CP) at the beginning of each frame in which the length of the later must be longer than the channel impulse response (CIR) [1, 9]. The use of a long CP length increases energy wastage and reduces the system throughput [10]. In order to mitigate these drawbacks, a short CP length is requested with respect to the length of CIR and a time-domain equalizer (TEQ), also known as a channel shortening equalizer (CSE), is used at the beginning of the receiver to shorten the effective channel (i.e. channel-TEQ combined), which makes the delay spread of the effective channel less than or equal to the length of CP [9,11-16].

Recently, researchers have proposed the use of TEQ in optical OFDM systems. A low complexity decision feedback time-domain equalizer (LCDF-TEQ) has been proposed in Ref. [13], while the normalized least mean squares (NLMS-TEQ) has been studied in Ref. [14], and the recursive least square (RLS-TEQ) in Ref. [15]. These methods require a long training sequence to estimate the channel and a large number of taps to make the CP shorter, which reduce the throughput and increase the complexity of the equalizer, respectively. The complexity in this case increases exponentially with the number of the taps. Blind methods, based on the use of the property of MCM, do not require the knowledge of the exact values of the transmitted signal, in contrast to the trained methods [17]. In the blind approach literature, some limited numbers of TEQ design have been explored, such as sum-squared autocorrelation (SAM) [17], single lag-autocorrelation minimization (SLAM) [18], multicarrier equalizer by restoration of redundancy (MERRY) [9], and the multicarrier equalizer by restoration of orthogonality (MERO) [9]. SAM and SLAM are the most well-known and performing algorithms. However, their complexities

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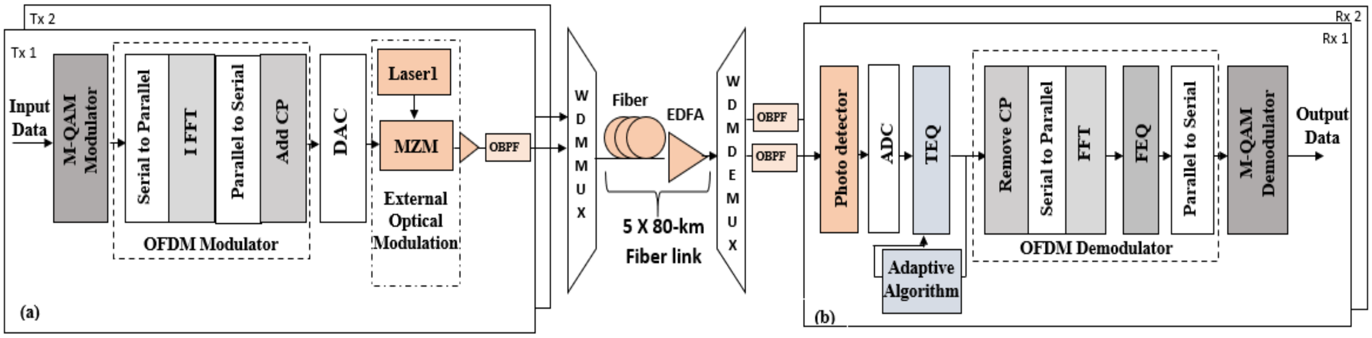

Fig. 1. Block diagram of the proposed WDM-DDO-OFDM system: (a) Transmitter, (b) Receiver with time domain equalizer (TEQ) and frequency domain equalizer (FEQ)

are higher due to the minimization of the squared autocorrelation and moving average (MA) method used to estimate the expectation operation. MERRY has a low complexity, but a very slow convergence as it updates TEQ coefficients per symbol. However, MERO has a fast convergence and a low complexity compared to SAM, where it exploits the orthogonality between subcarriers and the CP of the MCM system to shorten the effective channel [9].

According to the demand of reducing the energy wastage in the CP, we propose a MERO- TEQ with a low complexity (LMERO-TEQ) by designing a modified cost function of MERO algorithm. The complexity reduction is achieved by using the instantaneous method to estimate the expectation without the square, and further utilizing symmetrical TEQ property to reduce the complexity of the adaptation of TEQ to 50%. This method reduces the CP length by shortening the effective channel and furthermore makes the system less complex.

The system wavelength division multiplexing (WDM) allows multiple wavelength channels to be transmitted over the same fiber providing high transmission rates [19]. In this paper, the WDM-DDO-OFDM system [20] is used in order to achieve data rate of $100\,\mathrm{Gb/s}$. Moreover, to avoid the frequency-selective power fading caused by the chromatic dispersion, a single sideband (SSB) is used.

This paper is organized as follows. Section II presents a description of the transmitter and the receiver of the WDM-DDO-OFDM system. In Section III, some discussions on MERO-TEQ design, the modified cost function and the simplified MA method for updating the TEQ coefficients are presented. In addition, the exploitation of the symmetrical propriety of TEQ to reduce the computational complexity of the LMERO algorithm by fifty percent is reported as well in this section. Section IV provides simulation results of the LMERO algorithm in the WDM-DDO-OFDM system for a long-reach of SSMF with a comparison to others existing algorithms. Finally, some conclusions are drawn in Section V.

II. System Modeling

In Fig.1, the block diagram of the proposed WDM-DDO-OFDM system is illustrated.

In each optical channel WDM, the input data is transmitted through an M-QAM modulator to produce M-array sequences. Then, the data is converted into parallel data to form a vector suitable for the input of the inverse fast Fourier transform (IFFT). The digital time domain signal is obtained by using N-point IFFT then the symbol OFDM can be expressed [13] as

$$x(n) = \frac{1}{N}\sum_{m=0}^{N-1} X(m)e^{\frac{j2\pi mn}{N}}, \qquad m = (0, \dots, N-1) \tag{1}$$

where, n and m are the time domain index of the OFDM sample, and subcarrier index, respectively.

A CP of length ν is added to avoid ISI caused by GVD in which the extended OFDM symbol $x_\nu(n)$ contains $N+\nu$ number of sample (bins). The bins in $x_\nu(n)$ are uncorrelated (orthogonal) to each other and we can write [13]

$$x_\nu(n) = \begin{cases} x(n-\nu+N) & 0 < n < \nu - 1 \\ x(n-\nu) & \nu \le n \le N+\nu-1 \end{cases} \tag{2}$$

In the optical channel, the CP duration T_ν condition in the absence of differential group delay (DGD) is given by

$$\frac{c}{{f_c}^2}|D_t|N_{sc}\Delta f \le T_\nu \tag{3}$$

where c is the speed of light, f_c is the optical carrier frequency, D_t is the total amount of chromatic dispersion in units of $\mathrm{ps/nm}$, N_{sc} is the number of subcarriers used and Δf is subcarrier channel spacing [21].

The time domain signals will then be serialized and converted into an analog signal using a digital to analog converter (DAC). After that, a radio frequency (RF) up-converter is used to ensure the required spectral gap between the optical carrier and the lowest OFDM subcarrier, who's role is to avoid the effect of subcarrier-subcarrier mixing signal (SSBM) generated during square-law detection via the photodiode (PD) in the receiver [4]. The optical signal of each channel is achieved by using a continuous wave (CW) laser and a Mach-Zehnder modulator (MZM), then it is amplified to compensate the MZM loss and filtered by an optical bandpass filter (OBPF) to suppress the lower side of the optical spectrum.

All the optical channels are recombined using a WDM multiplexer (MUX) and sent through a 5-spans transmission line, each has an 80 km SSMF (ITU-G652) and an Erbium-doped fiber amplifier (EDFA).

At the receiver, another WDM demultiplexer (DEMUX) is used to separate individual optical tones from the desired receiver. Then, an OBPF is utilized to reduce the amplified spontaneous emission (ASE) noise induced by EDFA and what gives subsequently a low relative intensity noise (RIN) of the laser.

After detecting the received optical signal by a photodiode, the electrical signal is down-mixed and converted from an analog to a digital signal using respectively a RF-down-converter and analog to digital converter (ADC). We can write the received signal in the time domain as

$$r(n) = x_v(n) * h(n) + w(n) = \sum_{l=0}^{L_f} h(l) x_v(n-l) + w(n) \quad (4)$$

where $h(n)$ is the discrete impulse response of SSMF and $w(n)$ is a zero-mean additive white Gaussian noise (AWGN) sequence uncorrelated with the source sequence of variance σ_w^2 and $*$ stands for convolution operation.

Then, the electrical signal propagates through the blind TEQ with a length of $L_g + 1$. The effective channel is a discrete time convolution between the channel vector $\mathbf{h}$ and the equalizer tap vector $\mathbf{g}$, i.e.,$\mathbf{c} = \mathbf{h} * \mathbf{g}$, where the length of the effective channel is $L_c + 1$, and $L_c = L_f + L_g$.
The output sequence $y(n)$ of the TEQ is given by

$$y(n) = r(n) * g(n) = \sum_{i=0}^{L_g} g(i) r(n-i) = \mathbf{g}^{\boldsymbol{T}} \mathbf{r}_{\boldsymbol{n}} \quad (5)$$

where $\mathbf{g}^{\boldsymbol{T}}$ is the transposed vector of $\mathbf{g}$.

After removing the CP and applying the N-point FFT, one-tap frequency domain equalizer (FEQ) or zero-forcing equalizer is used to compensate the phase and amplitude distortions. The coefficients of FEQ are calculated by inverting the fiber frequency response and they can be expressed as

$$F(m) = \frac{1}{\widehat{H}(m)}, \qquad m = 1, \dots, N_{sc} \quad (6)$$

where $\widehat{H}(m)$ is the estimated channel frequency response.

According to expression (6), the estimation of the channel frequency response is requested, which can be obtained by transmitting N_{sc} symbols at the beginning of the data information equivalent to one symbol OFDM;

$$\widehat{H}(m) = \frac{R(m)}{T(m)} + \frac{V(m)}{T(m)}, \qquad m = 1, \dots, N_{sc} \quad (7)$$

where $T(m)$, $R(m)$, and $V(m)$ are the transmitted training symbols, received training symbols and noise in the frequency domain, respectively. Furthermore, the training symbol is also used previously for synchronization.

III. LOW-COMPLEXITY MERO TEQ DESIGN

In this section, we describe the low complex TEQ design in three parts: (i) First we define the MERO algorithm using MA method. (ii) Then we design a modified MERO-TEQ cost function with the instantaneous method to estimate the expectation and without the square. This simplified version of the MA method is similar to the work reported in [1] and [17], (iii) Then we exploit the symmetrical TEQ design to further reduce the complexity of the implementation of LMERO-TEQ by fifty percent.

A. MERO-MA Algorithm

The MERO algorithm proposed in [9] is a combination of two algorithms: minimization of correlation of adjacent samples (MCAS) algorithm and multicarrier equalizer by restoration of redundancy (MERRY) algorithm.

It is known that the added samples in the CP are identical to the last samples of the OFDM symbol, which we can write as

$$x_v(MK + n) = x_v(MK + n + N), \qquad n \in \{1,2, \dots, v\} \quad (8)$$

where $x_v(MK + n)$ is the n^{th} sample of the K^{th} transmitted OFDM symbol, $M = N + v$ is the total OFDM symbol duration.

The main idea of MERO algorithm is to minimize the difference between the correlation of the adjacent samples and their counterparts in the CP by minimizing the following cost function

$$J_{MERO} = \left| \mathbf{E}\big[\mathbf{E}[y(n+\Delta) y(n+1+\Delta)] - \mathbf{E}[y(n+N+\Delta)\, y(n+1+N+\Delta)]\big] \right|^2, \qquad n \in \{1, \dots, v\} \quad (9)$$

where $\Delta \in \{0, \dots, M-1\}$ is the symbol synchronization parameter, which represents the desired delay channel-TEQ combination. The coefficient update equation for MERO algorithm, utilizing the steepest gradient-descent algorithm, is given by

$$\mathbf{g}^{\mathrm{k}+1} = \mathbf{g}^{\mathrm{k}} - \mu \nabla_g J_{MERO} \quad (10)$$

where $\mathrm{k} = 0, \dots, L_g - 1$ is the index TEQ coefficients, μ denotes the step size ($\mu = 0.1$), $\nabla_g J_{MERO}$ is the gradient of the MERO cost function with respect to $\mathbf{g}$ that is given by

$$\begin{aligned} \nabla_g J_{MERO}(n) = \mathbf{E}[\{&\mathbf{E}[y(n+\Delta)\, \mathbf{r}_{n+1+\Delta}] \\ &- \mathbf{E}[y(n+N+\Delta)\, \mathbf{r}_{n+1+N+\Delta}] \\ &+ \mathbf{E}[\mathbf{r}_{n+\Delta}\, y(n+1+\Delta)] \\ &- \mathbf{E}[\mathbf{r}_{n+N+\Delta}\, y(n+1+N+\Delta)]\} \\ &\times \{\mathbf{E}[y(n+\Delta) y(n+1+\Delta)] \\ &- \mathbf{E}[y(n+N+\Delta) y(n+1+N+\Delta]\}] \end{aligned} \quad (11)$$

Generally, the implementation of the existing blind adaptive algorithms is performed by using the MA or autoregressive (AR) methods [1,16,17] in which the MA method is defined by

$$\mathbf{E}[y(n) y(n+1)] = \sum_{n=1}^{S} \frac{y(n) y(n+1)}{S} \quad (12)$$

TABLE I

COMPLEXITY COMPARISONS IN TERMS OF NUMBER OF MULTIPLICATIONS AND ADDITIONS PER TEQ UPDATE

Algorithms	Moving average estimate	Instantaneous estimate
SAM [17]	$4SL_g(L_f - v)$	$4L_g(L_f - v)$
MERO [9]	$8SL_g$	$8L_g$
LMERO	$6SL_g$	$6L_g$

where S is a number of samples per averaging windows size. A better estimation can be achieved by using a large number of samples S, but this leads to a higher complexity. This latest is proportional to S.

B. Low Complexity MERO (LMERO) Algorithm

As per the central limit theorem, we can assume that the output of IFFT is a stationary process due to the wide-sense stationary Gaussian process [1,16], a simplified low complexity instantaneous estimate is used, which is approached as

$$\mathbf{E}[y(n)y(n+1)] \approx y(n)y(n+1) \quad (13)$$

On the other hands, a modified cost function design is done to reduce the complexity of the MERO by removing the global expectation and the square of the eq. (9). This change does not bring any loss in performance as will be investigated in Section IV. The modified cost function is defined as

$$J_{LMERO} = \mathbf{E}[y(n+\Delta)y(n+1+\Delta)] - \mathbf{E}[y(n+N+\Delta)\, y(n+N+1+\Delta)] \quad (14)$$

where $n \in \{1, \dots, v\}$ and $\Delta \in \{0, \dots, M-1\}$.

Based on (13) and (14), the coefficient update equation for the LMERO algorithm with the low-complexity method is given by

$$\mathbf{g}^{k+1} = \mathbf{g}^{k} - \mu\{\{[y(n+\Delta)\mathbf{r}_{n+1+\Delta}] - [y(n+N+\Delta)\mathbf{r}_{n+1+N+\Delta}]\} + \{[\mathbf{r}_{n+\Delta}y(n+1+\Delta)] - [\mathbf{r}_{n+N+\Delta}y(n+1+N+\Delta)]\}\} \quad (15)$$

To compare the complexity of LMERO-TEQ, the MERO and SAM algorithms are considered as a reference. In [9], SAM used MA method and instantaneous implementations method with a complexity of $4SL_g(L_f - v)$ and $4L_g(L_f - v)$ multiplications\additions per each update, respectively. Thus, the algorithm with the instantaneous method is lower in complexity by S times compared with the MA method. On the other hand, MERO has a complexity equal to $8SL_g$ multiplications\additions per updates using the MA method. In the LMERO algorithm, the computational complexity of the eq. (15) is equal to $6SL_g$ and $6L_g$ multiplications\ additions per each update using the MA method and the instantaneous method, respectively. Therefore, the complexity of LMERO-TEQ is reduced by 25% (i.e. $(8-6)/8$) compared to MERO-TEQ. Table I summarizes the complexity of algorithms using MA estimate and the low-complexity instantaneous estimate.

C. LMERO-TEQ With Symmetrical Property

It has been shown in numerous papers [22, 23] that in the case of an infinite or finite length of TEQ, the optimum TEQ coefficients become respectively symmetrical or they have been forced to be perfectly symmetrical in order to reduce the complexity implementation by 50% compared to the non-symmetrical TEQ design.

The symmetrical implementation of the proposed algorithm is named as LMERO-Sym-TEQ in which their coefficient update equation is given by

$$\mathbf{f}^{k+1} = \mathbf{f}^{k} - \mu\nabla_g J_{LMERO} \quad (16)$$

Where $\mathbf{f}$ is a real $(\lfloor L_g/2 \rfloor \times 1)$ vector with a length equal to half of $\mathbf{g}$. The adaptation of LMERO-Sym-TEQ is performed only for $\lfloor L_g/2 \rfloor$ and symmetry is enforced to form **g.**

Hereafter, the TEQ coefficients can be put in the form as in expression (17) for an odd TEQ length or as in expression (18) for an even case.

$$\mathbf{g}^T = [\mathbf{f}^T (I_{cd}\mathbf{f})^T] = \begin{bmatrix} g_0 \\ g_1 \\ \vdots \\ g_{\lfloor L_g/2 \rfloor} \end{bmatrix}^T \left(\begin{bmatrix} 0 & 0 & 0 & \dots & 1 \\ 0 & 0 & \dots & 1 & 0 \\ \vdots & \vdots & \vdots & \vdots & \vdots \\ 0 & 1 & \cdots & 0 & 0 \\ 1 & 0 & \cdots & 0 & 0 \end{bmatrix} \begin{bmatrix} g_0 \\ g_1 \\ \vdots \\ g_{\lfloor L_g/2 \rfloor} \end{bmatrix}^T \right)^T \quad 17)$$

$$\mathbf{g}^T = [\mathbf{f}^T \lambda (I_{cd}\mathbf{f})^T] = \begin{bmatrix} g_0 \\ g_1 \\ \vdots \\ g_{\lfloor L_g/2 \rfloor} \end{bmatrix}^T \lambda \left(\begin{bmatrix} 0 & 0 & 0 & \dots & 1 \\ 0 & 0 & \dots & 1 & 0 \\ \vdots & \vdots & \vdots & \vdots & \vdots \\ 0 & 1 & \cdots & 0 & 0 \\ 1 & 0 & \cdots & 0 & 0 \end{bmatrix} \begin{bmatrix} g_0 \\ g_1 \\ \vdots \\ g_{\lfloor L_g/2 \rfloor} \end{bmatrix}^T \right)^T \quad (18)$$

Where λ is a scalar initializes as 1 and $\lfloor . \rfloor$ is the *floor* function.

In this subsection, it can be observed that the LMERO-Sym-TEQ algorithm can be initialized by only computing half of coefficients TEQ, without degrading performance. This leads to a significant reduction in implementation complexity by 50% compared to the LMERO-TEQ defined in subsection (B) that use all the TEQ length L_g. However, the LMERO-TEQ algorithm, which use the instantaneous method to estimate the expectation to replace either the MA method used in MERO algorithm (subsection (A)), updates every sample due to its dependence on the current settings (i.e. current received signals) and reduces the algorithm complexity by a factor of S multiplication/addition per update. In contrast, the MA method performs the updating of each S samples and imposed a large enough S samples to give a reliable estimate of the expectation which increases the complexity of the algorithm by a factor of S. Furthermore, the LMERO-TEQ was obtained by modifying the MERO-TEQ cost function defined in subsection (A), which reduces the algorithm complexity by 25% without any loss in

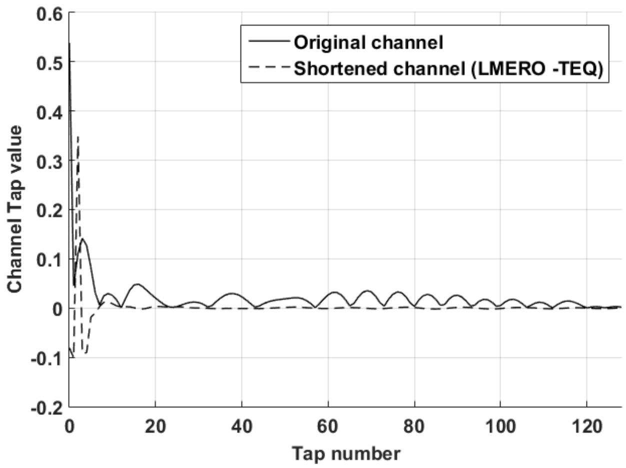

Fig. 2. Original and effective (shortened) CIR.

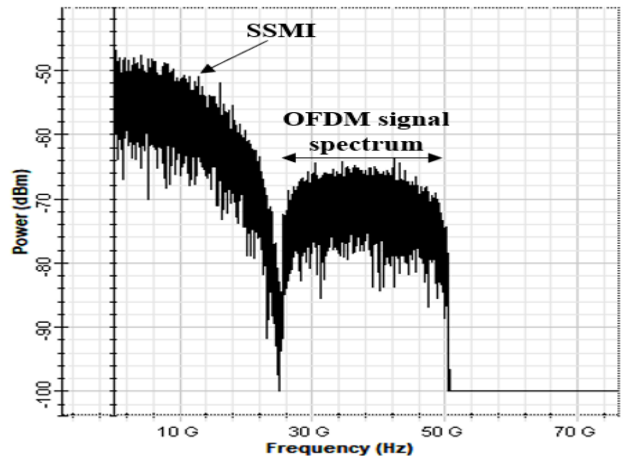

Fig. 3. Electrical spectrum of the DDO-OFDM received signal after 400 km of SSMF with an $\mathrm{OSNR} = 25$ dB.

performance as will be illustrated in the simulation studies in Section IV.

IV. Simulation Results And Discussion

As illustrated in Fig.1, numerical simulations have been carried out in the 100 Gb/s WDM-DDO-OFDM transmission system with a SSB optical signal based on co-simulation of the Optisystem and Matlab platforms [24]. The pseudo random bit sequence (PRBS) data with a pattern length of $2^{15} - 1$ bits are mapped with 4-QAM into 512 data subcarriers. Furthermore, another 512 nulled subcarriers, of 1024 IFFT points, are used to mitigate the *subcarrier×subcarrier* beating noise. The data rate of 100 Gb/s is obtained by combining two 50 Gb/s which is equivalent to two 25 GBaud using the WDM system. The OFDM symbol duration of each channel is $\mathrm{T} = 20.48$ ns. The sampling rate of DAC/ADC is equal to 50 Gsamples/s. The electrical signal was up-converted to an RF frequency $\mathrm{F_{RF}} = 38$ GHz and modulated by an MZM with an extinction ratio of 20 dB and a CW laser source with a linewidth of 100 kHz and is operating at wavelength $\lambda_1 = 1552.52$ nm for the first channel and at wavelength $\lambda_2 = 1550$ nm for the second channel. The output power of each CW laser equal to 0 dBm. By assuming that the nonlinearity effect is negligible or has been compensated, the fiber is modelled as a linear channel. The transmission line is 5 spans of 80 km SSMF (ITU-G652) and after each span, an optical amplifier EDFA (Erbium-doped fiber amplifiers) with a flat gain of 20 dB and noise figure of 5 dB is inserted. The proprieties of the optical fiber are as follows; an operating wavelength of 1550 nm, an attenuation coefficient of 0.2 dB/km, a dispersion coefficient 17 ps/(km. nm) and a fiber length of 400 km.

A 50 GHz photodiode for each channel is used at the receiver to detect the optical signal. The total PD noise is influenced (subjected) by adding the contribution of shot noise and thermal noise. These two noise factors are independent random processes with approximately Gaussian statistics.

The aim of TEQ is to shorten the CIR. In Fig. 2, we show the impulse response of the original channel and of the combined channel-equalizer (effective channel) after executing LMERO algorithm. We can see that we amended to achieve our goal and furthermore, the LMERO algorithm can shorten the CIR in a window equal to 8 taps less than the CP length which is 128 taps.

Fig.3 represents the electrical spectrum of the received DDO-OFDM signal of one channel detected by a photodetector after transmission over 400 km of SSMF with an $\mathrm{OSNR} = 25$ dB. It is clearly shown that the use of the SSB optical spectrum and the spectral gap before the transmission respectively can reduce the frequency selective power fading caused by the chromatic dispersion and the subcarrier-subcarrier mixing interference (SSMI) effect generated by the beating between OFDM

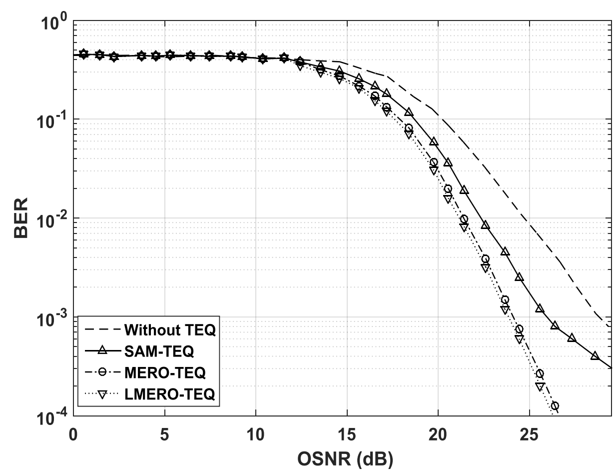

Fig. 4. BER as a function of OSNR for DDO-OFDM without and with different TEQ for an inadequate CP length equal to $\mathrm{T}/256$.

subcarriers when the signal is detected by a square-law photodetector. Nevertheless, there remain some residues of the latter effect due to the use of the short length of CP which will be compensated by the LMERO-TEQ.

To evaluate the performance of the system, the bit error rate (BER) against optical signal to noise ratio (OSNR) is plotted. The BER is measured by counting the number of different bits between the transmitter and receiver. The OSNR was defined

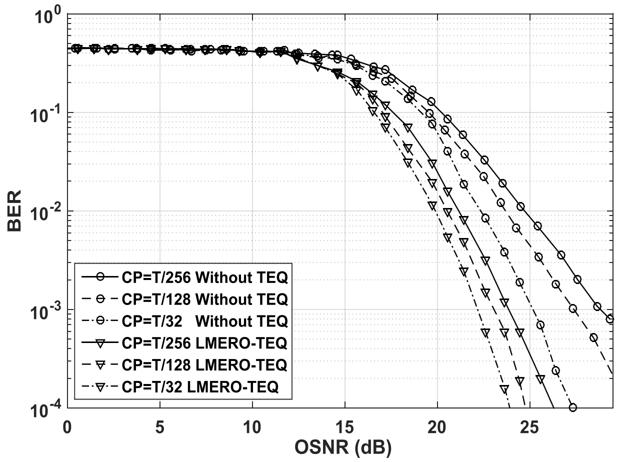

Fig. 5. BER as a function of OSNR for DDO-OFDM with and without LMERO-TEQ for different sizes of CP.

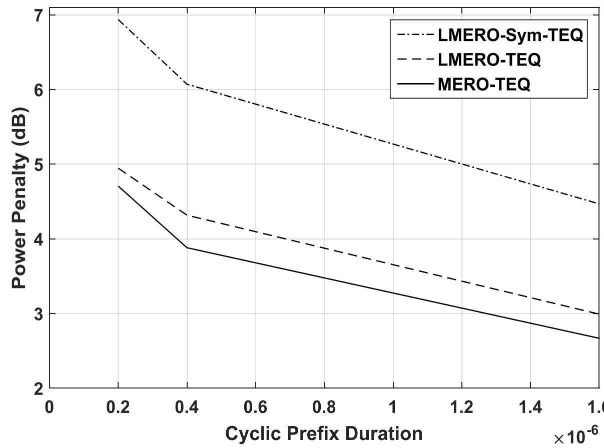

Fig. 7. Power penalty as a function of the CP duration for MERO-TEQ, LMERO-TEQ and LMERO-Sym-TEQ while considering BER $= 10^{-3}$ as a reference.

as the total optical signal power divided by the amplified spontaneous emission (ASE) noise power. The latter two evaluation parameters are measured over a 12.5 GHz noise bandwidth, which is equivalent to 0.1 nm at 1550 nm.

For an inadequate CP length, Fig.4 illustrates the BER performance for four cases: (i) LMERO-TEQ, (ii) MERO-TEQ, (iii) SAM-TEQ, and without TEQ. It can be observed that the BER of LMERO-TEQ and MERO-TEQ are slightly inferior to the others for low OSNR due to noise enhancement, while for high OSNR, its BER are very lower to the others. On the other hand, the LMERO-TEQ with its low-complexity has a gain of 0.255 dB and of 2.1 dB relative to MERO-TEQ and SAM-TEQ, respectively, by taking BER 10^{-3} as a reference. This is

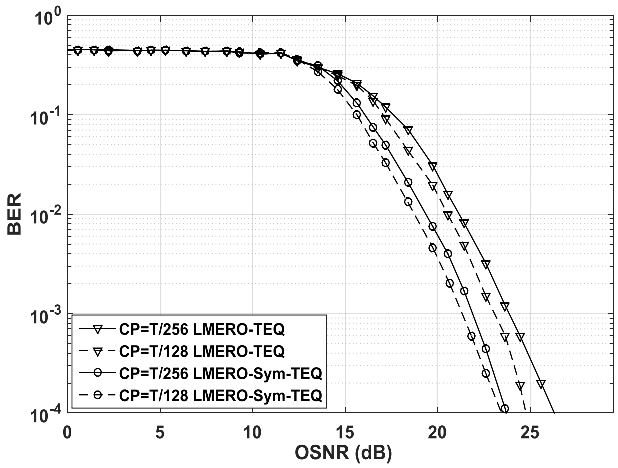

Fig. 6. BER as a function of OSNR for DDO-OFDM for LMERO-TEQ with and without using symmetrical property with different sizes of CP.

due to its effectiveness in minimizing the correlation between the adjacent samples.

With a view to reducing the CP of optical OFDM system, in Fig. 5 we show the performance of LMERO-TEQ for different size of CP. It is clear from this figure that for a very weak CP length of $\mathrm{T}/256$, the system without using the TEQ is very much affected by the ISI and ICI interferences. However, when the LMERO-TEQ is used with the CP length of $\mathrm{T}/256$, the performance of the system becomes much better than for a long CP length without using the TEQ, which demonstrates that our proposed algorithm can reduce the CP length to 0.39 %.

Furthermore, to reduce the implementation complexity of LMERO-TEQ by a percentage of 50%, the symmetrical property is used, in which the adaptation of TEQ can be performed by only computing half of the TEQ coefficients (i.e. $\lfloor L_g/2 \rfloor$) and symmetry is enforced to form $\mathbf{g}$. As can be seen in Fig. 6, our simulation model shows that LMERO-Sym-TEQ gives better performance than LMERO-TEQ in term of OSNR. By taking BER of 10^{-3} as a reference, the LMERO-Sym-TEQ improves the OSNR from 23.87 dB to 21.88 dB and from 23.06 dB to 21.31 dB compared to LMERO-TEQ for a $\mathrm{CP} = \mathrm{T}/256$ and $\mathrm{CP} = \mathrm{T}/128$, respectively.

Next, in Fig. 7 we represent the obtained power penalties as a function of CP duration. Note that, the power penalty indicates how much OSNR should be wasted on the CP samples. By taking BER of 10^{-3} as a reference, the maximum power penalties that have to be paid are 4.7 dB, 4.95 dB and 6.93 dB if MERO-TEQ, LMERO-TEQ and LMERO-Sym-TEQ are not used, respectively, when CP equal to $\mathrm{T}/256$.

According to Figure 7, when the CP is much shorter than the CIR, a severe effect of ISI and ICI occurs in the DDO-OFDM receiver without using the TEQ, which distorts the original signal. Under this worst case, the efficiency of the use of TEQ at the DDO-OFDM receiver is very clear for a smaller CP length (i.e. from $\mathrm{CP} = \mathrm{T}/256$ to $\mathrm{CP} = \mathrm{T}/128$). On the other hand, an important CP length can reduce the effect of ISI and ICI, which is explained in the Fig. 7 through the progressive decreasing of the gradient by increasing the CP length, but a large part of the transmitted energy is wasted on the CP samples.

V. Conclusion

A blind adaptive time-domain equalizer with low-computational complexity has been proposed and demonstrated for WDM-DDO-OFDM system over 400 km SSMF length. The validation of the modified MERO cost function algorithm, known as LMERO-TEQ, has been verified and compared with the existing SAM and MERO algorithms. Significant complexity reduction of 50% has been achieved and demonstrated by using an instantaneous estimate to calculate the expectation without the square and exploiting the symmetrical TEQ property. Our simulations results demonstrate that with a shorten CP length, the system performance has been improved by 6.93 dB when LMERO-Sym-TEQ with a low complexity is employed. It has been demonstrated that the LMERO-TEQ can reduce the CP up to 0.39%.

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