Compressed Sensing for Clipping Noise Cancellation in DCO-OFDM Systems based on Observation Interference Mitigation

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***Abstract*—In this paper, we propose a modified clipping noise cancellation scheme using compressed sensing (CS) technique with observation interference mitigation for direct current biased optical (DCO) orthogonal frequency division multiplexing (OFDM) systems. The interference components in potential observations are theoretically analyzed and approximately estimated by exploiting the statistical model of the clipped DCO-OFDM signal. Then, the modified CS scheme eliminates the estimated interference from the potential observations, which can minimize the contamination influence of channel noise on CS recovery algorithm. In addition, a strategy jointly considering the compressed ratio and the decision noise is presented to generate the measurement matrix, which can sample the reliable data tones as final observations. With this scheme, the clipping noise in time domain can be effectively corrected and the bit error rate (BER) performance of the system is significantly improved. Simulation results demonstrate that the proposed scheme can perform well even under severe clipping conditions.**

***Index Terms*—Peak-to-average power ratio; clipping noise; compressed sensing; intensity modulation with direct detection; OFDM**

I. Introduction

Orthogonal frequency division multiplexing (OFDM) has been regarded as a promising modulation candidate for use in many modern communication systems due to its robustness against multi-path fading [1]–[3]. With the low cost advantage, applying OFDM to optical fiber communication and visible light communication based on intensity modulation and direct detection (IM/DD) structure has gained a lot of research interest [4], [5]. Direct current biased optical OFDM (DCO-OFDM) is an alternative for the optical OFDM and is simple for implementation [6]. However, DCO-OFDM suffers from high peak-to-average power ratio (PAPR) problem and is more susceptible to nonlinear distortions. Since the light-emitting diode (LED) has a limited linear dynamic range, the DCO-OFDM signal with high PAPR would drive the LED at the transmitter to saturation, which will produce more nonlinear distortions thus degrade the overall system performance. Therefore, the PAPR of DCO-OFDM should be reduced for the linear and power efficient operation of LED.

A number of PAPR reduction schemes have been proposed for OFDM systems in the literature [7], [8]. Among them, clipping is the simple and economical deployment technique which directly limits the peak envelope of the transmitting signal to a given threshold level and can be employed in the real-valued IM/DD channel without any modification. However, clipping is a nonlinear operation that leads to in-band and out-of-band distortions, which can deteriorate the bit error rate (BER) performance and cause spectral spreading. To improve the system performance, adopting appropriate method to compensate the clipping distortions is necessary. According to the compressed sensing (CS) theory [9], the sparse signal can be reconstructed with high probability from its compressed observations. It is known that the clipping noise of the clipped DCO-OFDM signal is sparse in time domain. Recently, several significant findings based on CS algorithms for clipping noise correction have been reported [10]–[14]. In [10], Al-Safadi proposes a CS scheme where several tones are reserved at the transmitter before clipping and the clipping noise is estimated by exploiting the specified observations of these reserved tones. However, the data rate is reduced and the BER performance is not satisfactory. In [11], there is no data rate loss since the compressed observations of the pilot tones are exploited; however, the reconstruction accuracy is vulnerable to the channel noise and it shows unsatisfactory BER performance. In [12], Kim proposes a clipping noise cancellation scheme using CS, which selectively exploits the reliable observations of the clipping noise instead of using the whole data tones. Although the weakness of CS reconstruction to channel noise has been overcome, the number of observations is very large and the scheme cannot perform well under severe clipping. Motivated by the results in [12], a priori aided CS scheme and an iterative clipping scheme for clipping noise cancellation are presented in [13] and [14], respectively. Similarly, they will encounter the same problems mentioned in [12]. In particular, these methods cannot be directly used in DCO-OFDM system due to the Hermitian symmetry requirement in IM/DD channel.

In this paper, a modified clipping noise cancellation scheme using CS technique with the observation interference mitigation is proposed for DCO-OFDM system. A strategy for measurement matrix generation is adopted to obtain the reliable observations with a fixed compressed ratio which is less contaminated by decision noise. We corrected the clipping noise by exploiting its compressed observations underlying in the data tones, which has eliminated the estimated observation interference and thus is less contaminated by channel noise. The proposed scheme performs well and the clipping distortions can be corrected effectively. To the best of our knowledge, this paper is the first work to apply modified CS scheme to the DCO-OFDM systems in IM/DD optical communications.

The rest of this paper is organized as follows. In Section II, we briefly present the PAPR characteristic of DCO-OFDM and analyze the statistical model of clipping noise. In Section III, a modified CS scheme is proposed to overcome the sensitivity to channel noise and reduce the clipping distortion. Section IV provides the simulation results and discussions. Section V concludes this paper.

The notations are defined as follows. The time domain and the frequency domain vectors are represented by the lower and upper boldface italic letters (eg., $\boldsymbol{x}$ and $\boldsymbol{X}$), respectively. The $(n+1)^{th}$ element of the column vector $\boldsymbol{x}$ is denoted as $x(n)$. Upper case boldface letters (e.g., $\mathbf{F}$) denote the matrices. $\mathcal{X}_n$ denotes the n^{th} constellation point of signal constellation. $\mathbb{E}\{\cdot\}$, $(\cdot)^T$, $(\cdot)^*$, $|\cdot|$ and $\mathcal{Z}(\cdot)$ are the expectation, the transpose, the complex conjugate, the absolute value and symmetric conjugate operators. $\mathcal{N}\left(\mu,\sigma^2\right)$ denotes the Gaussian distribution with mean μ and variance σ^2. $\widehat{a}$ denotes an estimate of parameter of interest a.

II. System Model

A. DCO-OFDM and PAPR

Considering an OFDM symbol with N sub-carriers, the bit stream is mapped onto $\boldsymbol{X}$ based on the M-ary quadrature amplitude modulation (M-QAM) constellation. We assume that each constellation point is modulated with equal probability as $1/M$. The time-domain DCO-OFDM signal $\boldsymbol{x}$ is generated by 2N-point inverse discrete Fourier transform (IDFT), illustrated by:

$$x(n)=\frac{1}{\sqrt{2N}}\sum_{k=0}^{2N-1}X_H(k)\,e^{\frac{j2\pi nk}{2N}},n=0,1,\cdots,2N-1 \tag{1}$$

where $\mathbf{X}_H=\left[\mathbf{X}^T,\left(\mathcal{Z}(\mathbf{X})\right)^T\right]^T$ follows Hermitian symmetry property [15] to produce real-valued time-domain signal $\boldsymbol{x}$. According to the central limit theorem [8], for $N\to\infty$, $\boldsymbol{x}$ can be modeled as Gaussian distribution with $\mathcal{N}\left(0,\sigma_x^2\right)$.

The PAPR of $\boldsymbol{x}$ is defined as the ratio between the maximum peak power and the average power in one symbol period, which can be written as [8]

$$\Upsilon_x=\frac{\max\limits_{0\le n\le 2N-1}\left\{|x(n)|^2\right\}}{\mathbb{E}\left\{|x(n)|^2\right\}}=\frac{\max\limits_{0\le n\le 2N-1}\left\{|x(n)|^2\right\}}{\sigma_x^2}. \tag{2}$$

Therefore, the complementary cumulative distribution function (CCDF) can be calculated by

$$\text{CCDF}=\text{Prob}\left\{\Upsilon_x>r_0\right\}=1-\left[\text{erf}\left(\sqrt{r_0/2}\right)\right]^{2N} \tag{3}$$

where $\text{erf}(\psi)=2/\sqrt{\pi}\int_0^{\psi}e^{-t^2}dt$ denotes the error function.

B. Clipping Noise

Due to the high PAPR of OFDM signal, clipping is usually performed digitally before digital-to-analog converter (DAC) at given threshold. The clipped signal is expressed as [7]

$$\bar{x}(n)=\begin{cases}x(n), & A_l\le x(n)\le A_u,\\ A_u, & x(n)>A_u,\\ A_l, & x(n)<A_l.\end{cases} \tag{4}$$

For DCO-OFDM, we consider symmetric clipping $(A_u=-A_l=A_{th})$ here. The clipping ratio (CR) is defined as

$$\gamma=\frac{A_{th}}{\sqrt{\mathbb{E}\left\{|x(n)|^2\right\}}}=\frac{A_{th}}{\sigma_x}. \tag{5}$$

The output of clipping can be written as

$$\bar{x}(n)=x(n)+c(n) \tag{6}$$

where $c(n)$ is the clipping noise and sparse in the time domain. Since the envelope of $x(n)$ is Rayleigh distributed and N is sufficiently large, the average power of clipped $\bar{x}(n)$ can be calculated as

$$\sigma_{\bar{x}}^2=\sigma_x^2\left\{1-\sqrt{\frac{2}{\pi}}\gamma\exp\left(-\frac{\gamma^2}{2}\right)-\left(1-\gamma^2\right)\text{erfc}\left(\frac{\gamma}{\sqrt{2}}\right)\right\} \tag{7}$$

where $\text{erfc}(\cdot)=1-\text{erf}(\cdot)$. Nevertheless, based on the Bussgang theorem [6] for Gaussian inputs $\boldsymbol{x}\sim\mathcal{N}\left(0,\sigma_x^2\right)$, the clipped $\bar{x}(n)$ can be also statistically decomposed into two uncorrelated parts as

$$\bar{x}(n)=\alpha x(n)+d(n) \tag{8}$$

where $d(n)$ is the clipping distortion which is statistically uncorrelated to $x(n)$, e.g., $\mathbb{E}\left[x^*(n)\,d(n)\right]=0$ and $\alpha\le 1$ is a linear attenuation factor which can be calculated by

$$\alpha=\frac{\mathbb{E}\left[x^*(n)\,\bar{x}(n)\right]}{\mathbb{E}\left[|x(n)|^2\right]}=1-\text{erfc}\left(\frac{\gamma}{\sqrt{2}}\right) \tag{9}$$

We assume $d(n)$ to be a Gaussian random variable with zero mean and variance being σ_d^2, where σ_d^2 can be obtained by

$$\begin{aligned}\sigma_d^2&=\mathbb{E}\left[|\bar{x}(n)|^2\right]-\alpha^2\mathbb{E}\left[|x(n)|^2\right]\\&=\sigma_x^2\Big\{1-\gamma\sqrt{\frac{2}{\pi}}\exp\left(\frac{-\gamma^2}{2}\right)-\left(1-\gamma^2\right)\text{erfc}\left(\frac{\gamma}{\sqrt{2}}\right)\\&\quad-\alpha^2\Big\}.\end{aligned} \tag{10}$$

And the average power of $c(n)$ can be obtained by

$$\sigma_c^2=\left\{\left(1+\gamma^2\right)\text{erfc}\left(\frac{\gamma}{\sqrt{2}}\right)-\sqrt{\frac{2}{\pi}}\gamma\exp\left(-\frac{\gamma^2}{2}\right)\right\}\sigma_x^2. \tag{11}$$

In addition, the $\bar{x}(n)$ in frequency domain can be expressed as

$$\bar{X}(k)=X(k)+C(k)=\alpha X(k)+D(k) \tag{12}$$

where $C(k)$ and $D(k)$ are the $c(n)$ and $d(n)$ in frequency domain, respectively.

III. CS Scheme Based On Observation Interference Mitigation

A. Analysis of Interference in Observations

At the receiver, the digital received symbol in the frequency domain can be expressed as

$$Y(k) = H(k)\bar{X}(k) + W(k) \tag{13}$$

where $H(k)$ denotes the frequency domain channel response and $W(k)$ is the additive white Gaussian noise (AWGN) with variance of σ_w^2. For every received frame, the perfectly known channel response and the perfect synchronization in the demodulation process are assumed. Then, the equalized symbol can be obtain by a zero-forcing channel equalization, expressed by

$$\begin{aligned} Y_{eq}(k) &= H^{-1}(k)Y(k) \\ &= X(k) + C(k) + H^{-1}(k)W(k) \\ &= \alpha X(k) + D(k) + H^{-1}(k)W(k). \end{aligned} \tag{14}$$

After that, a maximum likelihood (ML) estimator for $X(k)$ is adopted as follows

$$\hat{X}(k) = \lfloor Y_{eq}(k) \rfloor = \arg \min_{\substack{\mathcal{X}_i \\ i=1,2,\cdots,M}} \left| \alpha^{-1} Y_{eq}(k) - \mathcal{X}_i \right| \tag{15}$$

where $\lfloor \cdot \rfloor$ denotes the ML estimator. If the $C(k)$ was estimated and subtracted from $Y_{eq}(k)$, the decision accuracy in (15) will be improved accordingly.

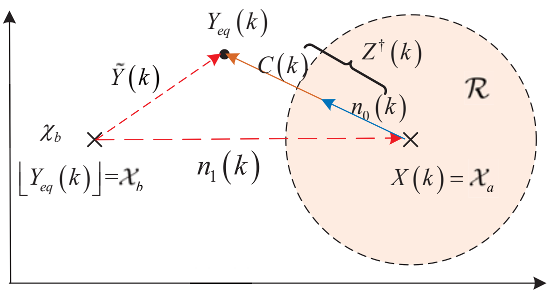

Fig. 1. Geometrical representation of the noise influence on symbol decision.

Then (14) can be rewritten in matrix form as

$$\mathbf{Y}_{eq} = \mathbf{H}^{-1}\mathbf{Y} = \mathbf{X} + \mathbf{C} + \mathbf{H}^{-1}\mathbf{W} \tag{16}$$

where $\mathbf{H} \triangleq \text{diag}(H)$, $\mathbf{Y}$, $\mathbf{X}$, $\mathbf{C}$ and $\mathbf{W}$ are the $N \times 1$ column vectors, $\mathbf{n}_0 = \mathbf{H}^{-1}\mathbf{W}$ is the channel noise which is introduced by a zero-forcing equalizer, $\mathbf{Z}^{\dagger} = \mathbf{C} + \mathbf{n}_0$ is the combined additive noise and clipping distortion, $\mathbf{C} = \mathbf{Fc}$ is the clipping noise in frequency domain, and $\mathbf{F}$ is an $N \times N$ unitary DFT matrix. To recover the sparse $\mathbf{c}$ by CS technique at the receiver, a measurement matrix $\mathbf{\Phi}$ should be adopted to construct the potential observations as

$$\tilde{\mathbf{Y}} = \mathbf{\Phi}(\mathbf{Y}_{eq} - \lfloor \mathbf{Y}_{eq} \rfloor) = \mathbf{\Phi}\mathbf{F}\mathbf{c} + \mathbf{\Phi}(\mathbf{X} - \lfloor \mathbf{Y}_{eq} \rfloor + \mathbf{n}_0). \tag{17}$$

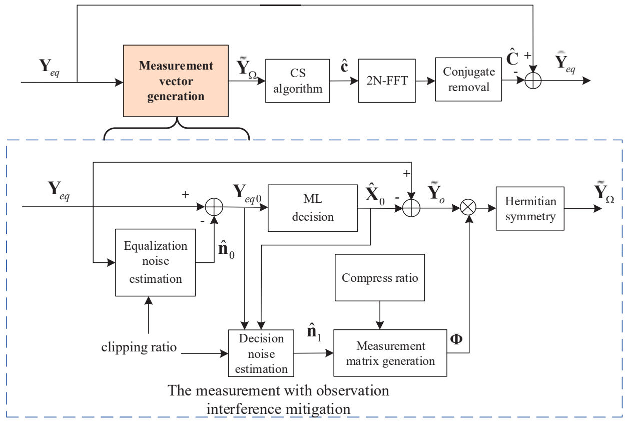

Fig. 2. Block diagram of the proposed CS scheme for clipping noise cancelation with observation interference mitigation..

We denote $\mathbf{Z} = \mathbf{X} - \lfloor \mathbf{Y}_{eq} \rfloor + \mathbf{n}_0$ as the interference component. As seen in (17), if the matrix $\mathbf{\Phi F}$ obeys a good restricted isometry property (RIP), the reconstruction of $\mathbf{c}$ from $\tilde{\mathbf{Y}}$ can be considered as a typical CS problem [9]. However, as illustrated by Fig. 1, $\mathbf{Y}_{eq}$ may be severely contaminated by $\mathbf{Z}^{\dagger}$, resulting in some data samples apart from the decision region $\mathcal{R}$. In this case, the estimated symbol is

$$\hat{X}(k) = \lfloor Y_{eq}(k) \rfloor = \mathcal{X}_b \neq X(k) \tag{18}$$

and the noise $\mathbf{n}_1 = \mathbf{X} - \lfloor \mathbf{Y}_{eq} \rfloor$ would be produced by (17) accordingly. The reconstruction performance of $\mathbf{c}$ seems mainly affected by $\mathbf{n}_1$ due to the fact that its amplitude is greater than $\mathbf{n}_0$. In fact, it is mainly determined by $\mathbf{n}_0$ when γ is fixed with a constant. As the perturbation $\mathbf{Z}^{\dagger}$ takes $\mathbf{Y}_{eq}$ out of its correct decision region, $\mathbf{n}_1$ is produced. Therefore, $\mathbf{n}_0$ plays a dominant role and determines the magnitude level of $\mathbf{n}_1$.

B. Interference Mitigation and Measurement Matrix Generation

As illustrated in Fig. 2, the block diagram describes the proposed CS scheme for clipping noise cancellation based on the observation interference mitigation. In general, the time domain $\hat{\mathbf{c}}$ can be recovered using CS recovery algorithm from the well-chosen observations $\tilde{\mathbf{Y}}_{\Omega}$, where Ω denotes the subset index of the reliable data tones in $\mathbf{Y}_{eq}$. After FFT, $\hat{\mathbf{C}}$ is obtained and subtracted from $\mathbf{Y}_{eq}$. The final symbol decision can be made from $\hat{\mathbf{Y}}_{eq}$ by (15). The detailed procedure of the proposed module is also demonstrated in the subfigure. Compared with the conventional CS scheme, the interference is mitigated from the potential observations and a strategy for measurement matrix generation is proposed here. The main objective is to find out the appropriate Ω and to sample the reliable data as effective final observations.

Firstly, the channel noise $n_0(k)$ will be estimated from $Y_{eq}(k)$. From (10) to (14), we know that $n_0(k)$ is linearly correlated to $Y_{eq}(k)$ and follows Gaussian distribution with

$\mathcal{N}\left(0, \sigma_w^2 / |H(k)|^2\right)$. Based on Bayes formula, the minimum mean square error estimator of $n_0(k)$ can be calculated by

$$\begin{aligned}\hat{n}_0(k) &= \mathbb{E}\{Y_{eq}(k) - \alpha X(k) - D(k)| \mathrm{Y}_{eq}(k)\} \\ &= Y_{eq}(k) - \alpha \sum_{i=1}^{M} \frac{P_r(Y_{eq}(k)| \mathcal{X}_i)\mathcal{X}_i}{\sum_{i=1}^{M} P_r(Y_{eq}(k)| \mathcal{X}_i)}\end{aligned} \tag{19}$$

where $P_r(Y_{eq}(k)| \mathcal{X}_i)$ is the conditional probability density function (PDF). According to (14), we know that the variable $\mathcal{V}(k) = Y_{eq}(k) - \alpha X(k)$ also follows Gaussian distribution as $\mathcal{N}(0, \sigma_{\mathcal{V}}^2)$, where

$$\sigma_{\mathcal{V}}^2 = \sigma_d^2 + \left|H^{-1}(k)\right|^2 \sigma_w^2. \tag{20}$$

Therefore, the conditional PDF $P_r(Y_{eq}(k)| \mathcal{X}_i)$ can be express as

$$P_r(Y_{eq}(k)| \mathcal{X}_i) = \frac{1}{\sqrt{2\pi\sigma_{\mathcal{V}}^2}} \exp\left(-\frac{\mathcal{V}^2(k)}{2\sigma_{\mathcal{V}}^2}\right). \tag{21}$$

Then $\hat{n}_0(k)$ can be obtained by combining (19) and (21).

Secondly, the new symbol $Y_{eq0}(k)$ is obtained by subtracting $\hat{n}_0(k)$ from $Y_{eq}(k)$, which would pull some symbols close to $\mathcal{R}$. After that, $Y_{eq0}(k)$ is fed into the ML estimator and we have the new estimated symbol as $\hat{X}_0(k) = \lfloor Y_{eq}(k) - \hat{n}_0(k) \rfloor$. Therefore, based on Bayes formula, the decision noise $n_1(k) = X(k) - \hat{X}_0(k)$ can be estimated by

$$\begin{aligned}\hat{n}_1(k) &= \mathbb{E}\{\hat{n}_1(k)| \mathrm{Y}_{eq0}(k)\} \\ &= \sum_{i=1}^{M} \left(\mathcal{X}_i - \hat{X}_0(k)\right) P_r(\mathcal{X}_i| Y_{eq0}(k)) \\ &= \sum_{i=1}^{M} \left(\mathcal{X}_i - \hat{X}_0(k)\right) \frac{P_r(Y_{eq0}(k)|\mathcal{X}_i)}{\sum_{i=1}^{M} P_r(Y_{eq0}(k)|\mathcal{X}_i)}\end{aligned} \tag{22}$$

where $P_r(Y_{eq0}(k)|\mathcal{X}_i)$ can be approximately represented as

$$P_r(Y_{eq0}(k)|\mathcal{X}_i) \simeq \frac{1}{\sqrt{2\pi\sigma_d^2}} \exp\left(-\frac{(Y_{eq0}(k) - a\chi_i)^2}{2\sigma_d^2}\right). \tag{23}$$

Thirdly, in a matrix form, the new potential observation vectors can be constructed by

$$\tilde{\mathbf{Y}}_0 = \mathbf{Y}_{eq} - \hat{\mathbf{X}}_0. \tag{24}$$

Since the main objective is to find out the well-chosen observation $\tilde{\mathbf{Y}}_\Omega$ from $\tilde{\mathbf{Y}}_0$, we can sample the reliable data components according to the measurement matrix $\mathbf{\Phi}$ based on the selection strategy, which can be described as follows

a) The decision noise vector $\hat{\mathbf{n}}_1$ is rearranged in ascending order in magnitude. Then, the corresponding index set is obtained as Ω_n.
b) Based on the compressed ratio ρ, we choose a set Ω from the top floor (ρN) coordinates of Ω_n, where floor $(\cdot)$ denotes the standard operation that floors to the nearest integer.
c) As the cardinality of Ω is M, an $M \times N$ measurement matrix $\mathbf{\Phi}$ is then generated by selecting the corresponding M rows from an $N \times N$ identity matrix $\mathbf{I}_N$. This means that $\mathbf{\Phi} = \mathbf{I}_N(\Omega)$.

Therefore, the final well-chosen observation $\tilde{\mathbf{Y}}_\Omega$ can be obtained by multiplying an $M \times N$ measurement matrix $\mathbf{\Phi}$ with $\tilde{\mathbf{Y}}_0$, expressed as

$$\begin{aligned}\tilde{\mathbf{Y}}_\Omega = \mathbf{\Phi}\tilde{\mathbf{Y}}_0 &= \mathbf{\Phi}\left(\mathbf{X} + \mathbf{C} + \mathbf{n}_0 - \hat{\mathbf{X}}_0\right) \\ &= \mathbf{\Phi}\mathbf{F}\mathbf{c} + \mathbf{\Phi}\left(\mathbf{X} - \hat{\mathbf{X}}_0 + \mathbf{n}_0\right) \\ &= \mathbf{A}\mathbf{c} + \boldsymbol{\eta}\end{aligned} \tag{25}$$

where $\mathbf{A} = \mathbf{\Phi}\mathbf{F}$ is the sensing matrix which shows a good RIP property, and $\boldsymbol{\eta} = \mathbf{\Phi}\left(\mathbf{X} - \hat{\mathbf{X}}_0 + \mathbf{n}_0\right)$ is the residual observation noise.

Therefore, the problem that recovering an K_c-sparse signal $\mathbf{c}$ from an $M \times 1$ compressed observations can be considered as a typical CS problem and be solved by using CS reconstruction algorithm with the specified ρ from the perspective of CS. In this paper, the orthogonal matching pursuit (OMP) algorithm is adopted. It should be mentioned that if ρ is set too small, the clipping noise cannot be corrected by OMP accurately because there are insufficient reliable observations. Therefore, the proposed scheme only works well when the required number M exceeds a certain threshold M_{min}. Here we set the threshold as

$$M_{\text{min}} = \min\left\{NC \cdot \text{erf}\left(\frac{\gamma}{\sqrt{2}}\right), 0.6N\right\} \tag{26}$$

where $C = 0.6$ is suitable for DCO-OFDM transmission from our numerical observations.

IV. Simulation Results

Now we make simulations to evaluate the BER performance of the proposed CS scheme for uncoded DCO-OFDM systems over an IM/DD optical channel. Note that the IM/DD optical channel can be modeled as a Gaussian low-pass filter which is widely adopted in many literatures such as [5] and [15]. For simplicity, the channel model in [15] with $f_{3\text{dB}}$ =80MHz is adopted and the influence of electro-optic device on system BER is not considered here. The parameters of DCO-OFDM used in the following simulations are depicted in Table. I.

TABLE I
SIMULATION PARAMETERS

DCO-OFDM Parameters	Values
Sampling frequency (GHz)	1
Modulation scheme	16QAM
IFFT/FFT length	2048
Subcarrier spacing (MHz)	0.488
Subcarrier number N	512
Cyclic Prefix (CP) ratio	1/64
Clipping ratio γ (dB)	2∼6

Fig. 3 shows BER performance of our proposed scheme with deep clipping γ = 4dB for various ρ. In addition, the

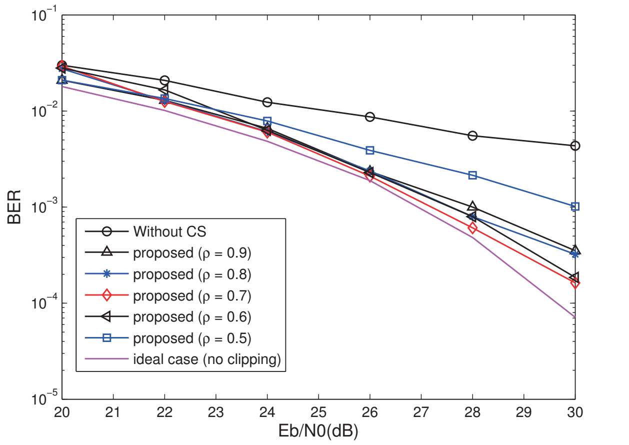

Fig. 3. BER performance versus Eb/N0 for the proposed CS scheme with different compression ratio when $\gamma = 4$dB.

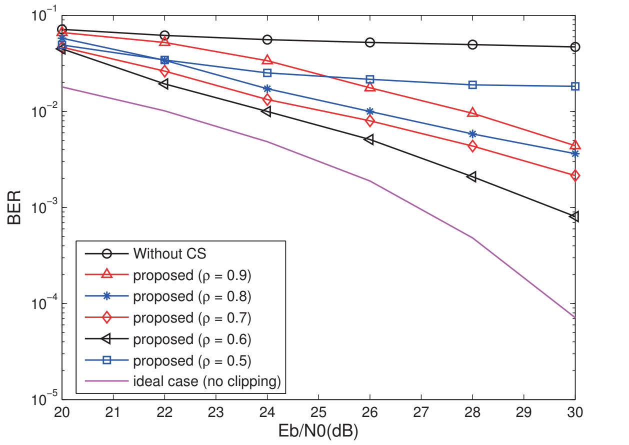

Fig. 4. BER performance versus Eb/N0 for the proposed CS scheme with different compression ratio when $\gamma = 2$dB.

original case without CS and the ideal case without clipping are also depicted here. Compared with the original case, the BER performance of the clipped DCO-OFDM has been greatly improved by the proposed scheme for $\rho = 0.6$, 0.7, 0.8 and 0.9. The gap of BER curves between the proposed scheme and the ideal case is small, e.g., 0.1dB, 0.3dB and 0.3dB for $\rho = 0.7$, 0.8 and 0.6, respectively, which indicates that the clipped signal can be exactly corrected by CS algorithm. However, it can be also observed from the figure that the BER performance of the proposed scheme with $\rho = 0.5$ is little improved because there are insufficient reliable observations which would induce the performance degradation during CS reconstruction.

Fig. 4. demonstrates the BER performance of the proposed scheme for $\gamma = 2$dB. It is seen that these curves are gradually moving away from the ideal case and the overall performance drops seriously when the extremely severe clipping are adopted. However, the performance of BER $< 10^{-3}$ can still be achieved by the proposed CS scheme for some compressed

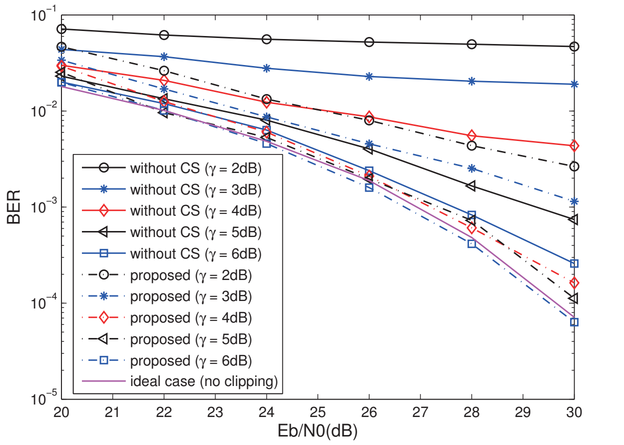

Fig. 5. BER comparisons between the original case and the proposed CS scheme with different clipping ration when $\rho = 0.7$.

ratio with favorable value. At Eb/N0 $=$ 30dB, the BER of the original case is obtained as 0.0234, while that of the proposed scheme for $\rho = 0.6$ is 0.0008, resulting in the reduction of at least an order of magnitude. The curve with $\rho = 0.5$ has worse performance when compared to other cases, while the curve with $\rho = 0.6$ has better performance than that with $\rho = 0.7$, $\rho = 0.8$ and $\rho = 0.9$. Therefore, the proposed CS scheme can still be used to improve the BER performance under extremely severe clipping conditions which would induce more nonlinear distortion. Furthermore, the moderate ρ can guarantee the validity and feasibility of the CS algorithm and minimize the BER, which is more appreciated in extremely deep clippings. Note that our scheme sets the used compressed ratio according to the actual clipping.

As γ varies from 2dB to 6dB, the BER performance comparisons between the original case without CS and the proposed CS scheme with $\rho = 0.7$ is presented in Fig. 5. It is demonstrated that the BER performance is significantly improved by the proposed CS scheme. At a given BER $= 10^{-3}$, the required Eb/N0 is reduced at least by 5dB, 2.2dB and 0.6dB for $\gamma = 4$dB, 5dB and 6dB, respectively. In the case of low CRs, e.g., $\gamma < 4$dB, even though the nonzero components of the clipping noise becomes too large thus degrades its sparsity level, the proposed scheme also performs well because the dominant observation interference is mitigated before the CS procedure, which is favorable for the clipping noise correction. While for the high CRs, e.g., $\gamma \geq 4$ dB, the nonlinear noise induced by clipping process is minor and the corresponding BER improvement is relatively smaller. Obviously, for $\gamma \geq 4$dB, the BER curves of the proposed scheme gradually approaches the ideal case, and the Eb/N0 gap between the proposed method and the ideal case become smaller. Note that the proposed CS scheme with $\gamma = 6$dB is better than the ideal case. Compared to the ideal case, the beneficial power gain is obtained from the observation noise mitigation and the average transmission power reduction by clipping.

V. Conclusions

In this paper, a modified scheme using CS technique with observation interference mitigation has been proposed in DCO-OFDM system for clipping noise cancellation. The nonlinear distortions caused by clipping has been effectively corrected and the BER performance of DCO-OFDM is significantly improved, which presents an effective solution for nonlinearity mitigation in IM/DD optical communications. Moreover, the proposed scheme can work under severe clipping conditions with moderate compressed ratios. At high CRs, the proposed scheme offers better BER performance than the ideal transmission, which validates the accuracy and effectiveness of this scheme.

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