PM-DCO-OFDM for PAPR Reduction in Visible Light Communications

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***Abstract*—Phase modulation is used in DCO-OFDM for visible light communications, whereby around 10 dB PAPR reduction is achieved. Considering the nonlinearity of LED emitters, our proposed method achieves better BER performance than that of conventional DCO-OFDM.**

I. Introduction

Visible light communication (VLC) has drawn great interests due to its distinctive merits. For example, abundant visible light spectrum relieves the pressure of increasingly crowded radio frequency spectrum [1]. Meanwhile, white light-emitting diodes (LEDs) are employed in VLC to provide both illumination and communication, which is friendly to environment [2], [3]. To reach the goal of communication, intensity modulation and direct-detection (IM/DD) are applied. Due to its inherent features, orthogonal frequency division modulation (OFDM) is recently introduced to VLC systems, in which high-speed data stream is divided into low-speed parallel data streams and transmitted simultaneously [4], [5]. However, OFDM suffers from high peak-to-average power ratio (PAPR), which causes nonlinear distortions and raises the requirement of power amplifiers. In VLC-OFDM system, high PAPR leads to performance loss due to the nonlinear characteristic of LED emitters. DC biased optical OFDM (DCO-OFDM) is widely used in VLC. In DCO-OFDM, the complex data signal in frequency domain is constrained by Hermitian symmetry [6], which generates real-valued signal in time domain after inverse fast Fourier transform (IFFT). After that, a DC bias is added to the signal before delivered to LED transmitter. Many techniques have been proposed to reduce the PAPR in DCO-OFDM. In [7], an exponential nonlinear companding function is applied to OFDM signal before fed into LED transmitter, achieving a PAPR of less than 1dB by sacrificing its bit-to-error rate (BER) performance. In [8], a pilot-assisted method is used, which has lower PAPR compared to the classical selected mapping (SLM) method. However, it inevitably suffers from higher complexity and relatively worse performance because of the distorted side information. Besides that, [9] proposes an algorithm based on semidefinite relaxation to optimize variables in tone injection (TI) dealing with PAPR issue in DCO-OFDM, which is an efficient technique with high complexity.

Since constant envelop OFDM (CE-OFDM) [10] can significantly decrease the PAPR of OFDM system, whereby phase modulation with constant envelop is used in traditional OFDM systems, leading to a 0 dB PAPR with a relatively good BER performance, we propose to apply phase modulation into conventional DCO-OFDM systems (PM-DCO-OFDM), whereby constant envelop is performed over real-valued signals. PAPR could be reduced by nearly 10 dB with negligible performance loss. In addition, there is no side information to be transmitted. Taking account of severe nonlinear characteristic of LED emitters, PM-DCO-OFDM achieves much better performance compared with its traditional DCO-OFDM counterpart.

The rest of paper is organized as follows. Section II describes the system model of DCO-OFDM. The proposed PM-DCO-OFDM is introduced in Section III. Simulation results are shown in Section IV, and Section V draws the conclusion.

II. System Model

For a DCO-OFDM system, the complex signal after QAM mapping, $\mathbf{X} = \{X[0], X[1], X[2], ..., X[N-1]\}$ is constrained to possess Hermitian symmetry property before it takes IFFT operation, where N denotes the number of subcarriers. The Hermitian symmetry is defined as

$$X[k] = X^*[N-k], \quad k = 1, 2, \cdots, N/2-1. \tag{1}$$

In DCO-OFDM, $X[0]$ and $X[N/2]$ are set to zero. After IFFT operation, the time-domain signal $x[n]$ is real-valued, which is given by

$$x[n] = \frac{1}{N}\sum_{k=0}^{N-1} X[k]\exp(\frac{j2\pi nk}{N}) \tag{2}$$

where $n = 0, 1, ..., N-1$. After that, the time-domain signal $\mathbf{x} = \{x[0], x[1], x[2], ..., x[N-1]\}$ passes a parallel-to-serial converter and then, cyclic prefix is added in order to avoid inter-symbol interference (ISI). After that, the signal is fed into a digital to analog converter and goes through a low-pass filter to generate the analytic signal $x(t)$. Finally, a moderate DC bias is added on $x(t)$ in order to generate the non-negative real-valued signal transmitted through LED emitters.

The PAPR of discrete OFDM signals is defined as the ratio of the maximum power of the signal to the average power. In DCO-OFDM, since OFDM signal $\mathbf{x}$ is real-valued due to Hermitian symmetry, the PAPR value of $\mathbf{x}$ is written as

$$PAPR = \frac{\max\{\|x[n]\|^2\}}{E\{\|x[n]\|^2\}} \quad n \in \{0, 1, \cdots, N-1\} \tag{3}$$

where $\|\cdot\|$ denotes the modulo of signal. In order to reduce the PAPR, it is essential to decrease the peak value of the signal while keeping the average power nearly constant. In this

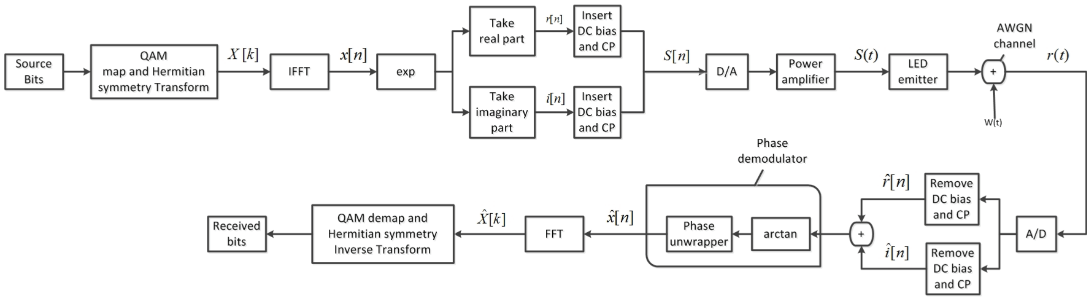

Fig. 1 PM-DCO-OFDM system model

paper, we propose to use constant envelop phase modulation into DCO-OFDM to solve this issue.

III. Proposed PM-DCO-OFDM

The system diagram of the proposed PM-DCO-OFDM is shown in Fig. 1. Binary source bits are mapped to M-ary quadrature amplitude modulation (M-QAM) constellation. Then a N-point IFFT is performed on the signal $\mathbf{X}$ in frequency domain and generates time samples $\mathbf{x}$. To use phase modulation, $\mathbf{x}$ has to be real-valued. So the frequency-domain signals need to meet Hermitian symmetry in (1). After the IFFT operation, the real-valued output $\mathbf{x}$ is fed into a phase modulator, and gets a 0 dB PAPR sequence $\{P[n] = \exp\{Cjx[n]+\theta\}\}_{n=0}^{N-1}$, where the envelop is constant to unit. C is a scaling factor that is set specially to make $Cjx[n]$ constraint in $[-\frac{2}{3}\pi, \frac{2}{3}\pi]$ to avoid phase ambiguity, and θ is an important rotation factor, which is used to reduce PAPR. The real part $\{r[n]\}_{n=0}^{N-1}$ and the imaginary part $\{i[n]\}_{n=0}^{N-1}$ are taken from $\{P[n]\}_{n=0}^{N-1}$ respectively, a DC-bias is added to make them positive, and a N_{cp}-sampled cyclic prefix(CP) is attached to both of $\{r[n]\}_{n=0}^{N-1}$ and $\{i[n]\}_{n=0}^{N-1}$. After that, the two sequences are concatenated together to be transmitted and are defined as

$$\begin{aligned}\mathbf{S} =&\{\underbrace{r[N-N_{cp}], \cdots, r[N-1]}_{CP_r}, r[0], r[1], \cdots, r[N-1],\\ &\underbrace{i[N-N_{cp}], \cdots, i[N-1]}_{CP_i}, i[0], i[1], \cdots, i[N-1]\}\end{aligned} \tag{4}$$

where CP_r and CP_i are the cyclic prefixes of $\{r[n]\}$ and $\{i[n]\}$ respectively, and N_{cp} is the length of them. The real non-negative signal $\mathbf{S}$ is then passed through a digital to analog converter as well as a power amplifier, and finally transmitted with a LED emitter, whose nonlinear characteristic can be modeled in [11]. For simplicity, the AWGN channel model is also assumed.

At the receiver, as shown in Fig. 1, the inverse operations are performed, including removing cyclic prefix and the DC bias, passing a phase demodulator which consists of an arctangent calculator as well as a phase unwrapper [10], and finally taking N-point FFT operation and the QAM demapping. The BER performance of the decoded bits and complementary cumulative distribution function (CCDF) of the signal PAPR are analyzed in Section IV.

The 0 dB PAPR cannot be achieved since the real and the imaginary parts of the constant-envelop signal are transmitted separately. Thus we have to calculate the PAPR of the two parts. In fact, if θ is equal to 0, it can be proved that the imaginary part has a rather higher PAPR than that of the real part. The PAPR of the imaginary part can be described as

$$PAPR_{\sin} = \frac{\max(\sin^2(Cx[n]))}{E(\sin^2(Cx[n]))} \quad n \in \{0, 1, \cdots, N-1\}. \tag{5}$$

In [6], the PDF of $\{Cx[n]\}_{n=0}^{N-1}$ is modeled as a Gaussian distribution with zero mean and variance of σ_D^2, when N is large. Therefore, its PDF can be written as

$$f_{Cx[n]}(v) = \frac{1}{\sigma_D\sqrt{2\pi}} \exp(-\frac{v^2}{2\sigma_D^2}). \tag{6}$$

When N is large enough, the maximum power is close to 1, since there exists $Cx[n]$ whose value is close to $\pi/2$. If taking the variance as 0.20, the average power of the imaginary part $Cx[n]$ can be calculated as [6]

$$\begin{aligned}E(\sin^2(Cx[n])) =& \int_{-\frac{2}{3}\pi}^{\frac{2}{3}\pi} \frac{\sin^2(v)}{\sqrt{2\pi \times 0.2}} \exp(-\frac{v^2}{2 \times 0.2}) dv \\ \approx& \, 0.1648.\end{aligned} \tag{7}$$

So the PAPR of the imaginary part will be $10\lg(1/0.1648) \approx 8.2$ dB. Furthermore, after the OFDM signal takes phase modulation with constant envelop over the complex plane, the data symbols on the unit circle will be often close to the point $(1, 0)$, which leads to a close-to-zero average power of the imaginary part and a relatively high PAPR. In order to solve this issue, a rotation of the interval $[-2/3\pi, 2/3\pi]$ is required. The value of θ is set to $\pi/4$ which allows both the real part and the imaginary part to have a rather low PAPR. Now, the PDF of $\{Cx[n]\}_{n=0}^{N-1}$ can be calculated as

$$f_{(Cx[n]+\theta)}(v) = \frac{1}{\sigma_D\sqrt{2\pi}} \exp(-\frac{(v-\pi/4)^2}{2\sigma_D^2}). \tag{8}$$

Take the average power of $\mathbf{x}$ as $\sigma_D^2 = 0.20$, and (8) is calculated as

$$\begin{aligned}E(\sin^2(Cx[n])) =& \int_{-\frac{2}{3}\pi}^{\frac{2}{3}\pi} \frac{\sin^2(v)}{\sqrt{2\pi \times 0.2}} \exp(-\frac{(v-\pi/4)^2}{2 \times 0.2}) dv \\ \approx& \, 0.4988.\end{aligned} \tag{9}$$

Finally, we have

$$PAPR_{\sin} \approx 10\lg(\frac{1}{0.4988}) \approx 3.0207\text{dB}. \tag{10}$$

As a result, it can be concluded that an about 3 dB PAPR of the system could be realized by the rotation of $\theta = \pi/4$.

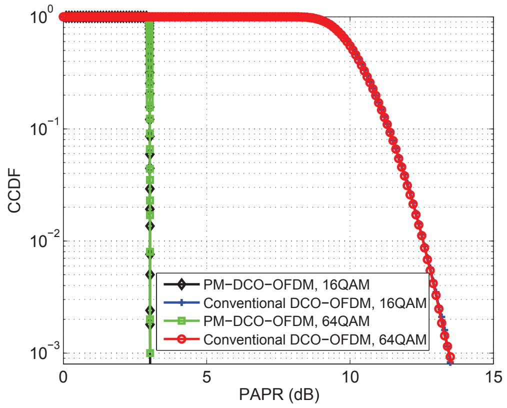

Fig. 2 CCDF of PAPR for DCO-OFDM and PM-DCO-OFDM using both 16QAM and 64QAM

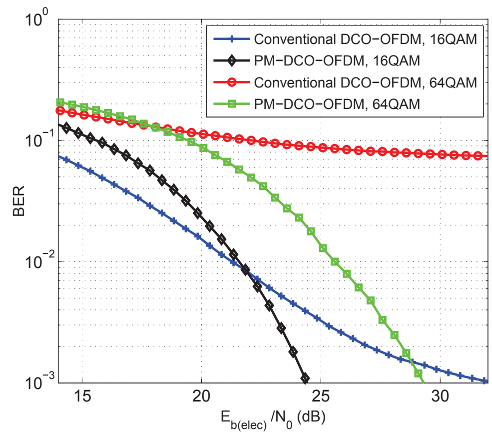

Fig. 3 BER performance for DCO-OFDM and PM-DCO-OFDM using both 16QAM and 64QAM

IV. Simulation Results and Discussion

The performance of the PM-DCO-OFDM is compared with its DCO-OFDM counterpart by simulations. The number of valid symbols per frame N_{qam} is set to 200, and $N = 512$. In the simulation, 16QAM and 64QAM are used, and the scaling factor of phase modulation C is set to $2\pi/(3 \times \max\{\|x[n]\|\}_{n=0}^{N-1})$, enabling the modulated phase within the range of $[(-2/3 + 1/4)\pi, (2/3 + 1/4)\pi]$. By using AWGN channel and considering the nonlinear model of LED emitters presented in [11], CCDF of PAPR and BER performance of DCO-OFDM with nonlinear distortion are simulated and compared with PM-DCO-OFDM.

Fig. 2 presents the simulation results for PAPRs of PM-DCO-OFDM and DCO-OFDM. For DCO-OFDM, PAPR is as much as 13.5 dB when CCDF is equal to 10^{-3}, which leads to huge performance loss when nonlinearity of LED exists. When PM-DCO-OFDM is applied, a significant PAPR reduction of over 10 dB could be achieved, which could improve the BER performance considerably for both 16QAM and 64QAM as the nonlinearity of LED exists.

The BER performance of PM-DCO-OFDM and its DCO-OFDM counterpart is shown in Fig. 3. When 16QAM is used, PM-DCO-OFDM achieves better BER performance than DCO-OFDM, because the SNR of PM-DCO-OFDM is much lower than that of conventional DCO-OFDM when BER reaches 10^{-3}. With 64QAM, the curve of traditional DCO-OFDM shows a severe error floor when BER equals 10^{-1}, while PM-DCO-OFDM definitely outperforms the conventional case. Therefore, if taking consideration of the nonlinear characteristic of LEDs, the PAPR reduction by PM-DCO-OFDM will lead to a significant BER performance gain.

V. Conclusion

To solve the nonlinearity issue of VLC communications, PM-DCO-OFDM is proposed which uses phase modulation into the conventional DCO-OFDM. In addition, the rotation phase of θ is added to reduce the inherent high PAPR of the imaginary part. As a result, more than 10 dB PAPR reduced could be realized.

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