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The design and performance analysis of optical wireless ACO-MC-CDMA system in the presence of clipping noise

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Dear editor,
With rapidly growing wireless data demand and the saturation of radio frequency (RF) spectrum, optical wireless communications (OWC) has become a viable candidate for indoor and medium range data transmission [1]. In general, the performance degradation of indoor OWC employing intensity modulation and direct detection (IM/DD) stems from the ambient light noise and multipath dispersion causing intersymbol interference (ISI). Furthermore, ambient light noise includes two different forms, i.e., the shot noise produced in the receivers and the interference caused by artificial light sources. In practice, the artificial light behaves like a narrow band jammer [2]. In addition, multiple access (MA) is another crucial issue to be considered for practical applications of OWC.

Code-division multiple-access (CDMA) is a promising MA technique which can combat the artificial light, multiuser and multipath interference. However, the use of CDMA becomes impractical due to the severe ISI at high data rate.

In IM/DD, a real-valued nonnegative signal modulates the intensity of a light emitting diode (LED), and it is detected by photodiode (PD). Orthogonal frequency division multiplexing (OFDM) provides inherent protection against ISI due to multipath dispersion. However, the conventional OFDM, which involves complex signalling, cannot be directly applied in an IM/DD. There are several popular methods to create the optical OFDM signal for IM/DD, such as DC-biased optical OFDM (DCO-OFDM), asymmetrically clipped optical OFDM (ACO-OFDM), unipolar OFDM (U-OFDM), and Flip-OFDM.

As a result, optical multi-carrier CDMA (MC-CDMA) has attracted significant interests to exploit the benefit of both optical OFDM and CDMA. The application of optical MC-CDMA in OWC is mainly motivated by the need to provide the MA capability, resist ambient light interference and combat ISI in highly hostile indoor optical channels. In [3, 4], the baseband complex MC-CDMA signaling was employed in IM/DD. In [5], an MC-CDMA system based on polarity reversed optical OFDM was presented.

In this letter, we propose the ACO-MC-CMDA which is a combination of CDMA and ACO-OFDM due to its high power efficiency and good bit error rate (BER) performance. However, the high peak-to-average power ratio (PAPR) also appears in ACO-MC-CDMA. Since the dynamic range of an LED is limited, the driving signal exceeding this range would be clipped. We would investigate the clipping distortion effect on the systems in this letter.

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Proposed ACO-MC-CDMA system. In the transmitter, the ith quadrature-amplitude modulation (QAM) symbol s_u^i of user u is copied and fed to G branches. Each branch is multiplied by a chip of a spreading code $\boldsymbol{c}_u = [c_u(0), c_u(1), \ldots, c_u(G-1)]^{\mathrm{T}}$, where $[\cdot]^{\mathrm{T}}$ denotes the matrix transpose. The spreading sequences for all the users are further added chip-by-chip, and the mth overlapped symbol is given by

$$X_{\mathrm{o}}^i(m) = \sum_{u=0}^{U-1} c_u(m) s_u^i, \tag{1}$$

where $m = 0, 1, 2, \ldots, G-1$.

In ACO-OFDM, only odd subcarriers are modulated as follows:

$$\boldsymbol{X}_{\mathrm{m}} = \left[0, X_{\mathrm{o}}^i(0), 0, \ldots, X_{\mathrm{o}}^i(G-1), \ldots, X_{\mathrm{o}}^{i*}(0)\right]^{\mathrm{T}}, \tag{2}$$

where $(\cdot)^*$ denotes the complex conjugation. The Hermitian symmetry of $\boldsymbol{X}_{\mathrm{m}}$ is required to ensure a real-valued inverse fast Fourier transform (IFFT).

Before carrying out IFFT, pre-scaling operation is applied to keep the average power of the output of IFFT constant, resulting in $\boldsymbol{X}_{\mathrm{s}} = \alpha \boldsymbol{X}_{\mathrm{m}}$, where α is the pre-scaling factor. Then taking an N-point IFFT gives $\boldsymbol{x}_{\mathrm{td}} = \alpha \boldsymbol{F}^{\mathrm{H}} \boldsymbol{X}_{\mathrm{m}}$, where $N = 4G$, $(\cdot)^{\mathrm{H}}$ denotes the complex conjugate transpose, and $\boldsymbol{F}$ is the unitary discrete Fourier transform matrix. This form of the IFFT/FFT has the advantage that the signals $\boldsymbol{x}_{\mathrm{td}}$ and $\boldsymbol{X}_{\mathrm{s}}$ have the same total energy. In fact for $N \geqslant 64$, the vector $\boldsymbol{x}_{\mathrm{td}}$ is approximately Gaussion with zero-mean and variance σ_0^2. In addition, α is given by [1]

$$\alpha = \sigma_0 \sqrt{\frac{N-1}{\sum_{n=0}^{N-1} |X_{\mathrm{m}}(n)|^2}}. \tag{3}$$

After IFFT, signal $\boldsymbol{x}_{\mathrm{td}}$ is clipped. The resulting signal $\boldsymbol{x}_{\mathrm{c}}$ is then converted from parallel to serial, and a cyclic prefix (CP) is appended. The resulting signal is digital to analog converted and DC biased resulting in $x_{\mathrm{LED}}(t) = x_{\mathrm{c}}(t) + B_{\mathrm{DC}}$, where B_{DC} is the DC bias.

A linear dynamic range of an LED is assumed between $i_{\min}$ and $i_{\max}$. Conditioning $x_{\mathrm{LED}}(t)$ within this range can result in non-linear distortion and/or double-sided signal clipping. In general, the clipping levels at the bottom and top are $\varepsilon_{\mathrm{b}} = \max(i_{\min} - B_{\mathrm{DC}},\ 0)$ and $\varepsilon_{\mathrm{t}} = i_{\max} - B_{\mathrm{DC}}$, and the clipping can be expressed as

$$x_{\mathrm{c}}(k) = \begin{cases} \varepsilon_{\mathrm{t}}, & x_{\mathrm{td}}(k) > \varepsilon_{\mathrm{t}}, \\ x_{\mathrm{td}}(k), & \varepsilon_{\mathrm{b}} \leqslant x_{\mathrm{td}}(k) \leqslant \varepsilon_{\mathrm{t}}, \\ \varepsilon_{\mathrm{b}}, & x_{\mathrm{td}}(k) < \varepsilon_{\mathrm{b}}. \end{cases} \tag{4}$$

The non-linear distortion can be modeled by means of the Bussgang theorem [6] as, $\boldsymbol{x}_{\mathrm{c}} = \rho \boldsymbol{x}_{\mathrm{td}} + \boldsymbol{n}_{\mathrm{c}}$, where $\boldsymbol{n}_{\mathrm{c}}$ is the additive non-Gaussian noise uncorrelated with $\boldsymbol{x}_{\mathrm{td}}$, ρ is the attenuation factor calculated by $\rho = \mathrm{Q}(\lambda_{\mathrm{b}}) - \mathrm{Q}(\lambda_{\mathrm{t}})$, where $\mathrm{Q}(\xi) = \frac{1}{\sqrt{2\pi}} \int_{\xi}^{\infty} \exp(\frac{-v^2}{2}) \mathrm{d}v$, $\lambda_{\mathrm{b}} = \varepsilon_{\mathrm{b}}/\sigma_0$ and $\lambda_{\mathrm{t}} = \varepsilon_{\mathrm{t}}/\sigma_0$. Especially the scenario with the ideal clipping is defined as $\varepsilon_{\mathrm{b}} = 0$ and $\varepsilon_{\mathrm{t}} = +\infty$.

At the receiver of the rth user, the electrical domain noisy analog signal $y(t)$ is given by

$$y(t) = h(t) \otimes x_{\mathrm{LED}}(t) + n_{\mathrm{A}}, \tag{5}$$

where $\otimes$ represents convolution, $h(t)$ is the impulse responses of the diffused optical channel which is modeled as ceiling-bounce model [7], n_{A} is an AWGN with two-sided power spectral density of $N_0/2$. $y(t)$ is further analog to digital converted and CP discarded. Next, an FFT is taken yielding

$$\boldsymbol{Y} = \boldsymbol{F}\boldsymbol{h}\left(\alpha\rho \boldsymbol{F}^{\mathrm{H}} \boldsymbol{X}_{\mathrm{m}} + \boldsymbol{N}_{\mathrm{c}} + \boldsymbol{B}_{\mathrm{DC}}\right) + \boldsymbol{N}_{\mathrm{A}}, \tag{6}$$

where $\boldsymbol{B}_{\mathrm{DC}} = [\sqrt{N} B_{\mathrm{DC}}, 0, \ldots, 0]^{\mathrm{T}}$, $\boldsymbol{h}$ is a circulant matrix, $\boldsymbol{N}_{\mathrm{c}}$ and $\boldsymbol{N}_{\mathrm{A}}$ are the clipping noise and AWGN in frequency domain, respectively. For large FFT size, $\boldsymbol{N}_{\mathrm{c}}$ is modeled as a zero-mean complex Gaussian noise with variance as follows:

$$\begin{aligned} \sigma_{\mathrm{c}}^2 = & \sigma_0^2 \Big(\rho(1 + \lambda_{\mathrm{b}}^2) - 2\rho^2 - \phi(\lambda_{\mathrm{t}})(\lambda_t - \lambda_{\mathrm{b}}) \\ & - \lambda_{\mathrm{b}}\big(\phi(\lambda_{\mathrm{b}}) - \phi(\lambda_{\mathrm{t}})\big) + \mathrm{Q}(\lambda_{\mathrm{t}})(\lambda_{\mathrm{t}} - \lambda_{\mathrm{b}})^2 \Big), \end{aligned} \tag{7}$$

where $\phi(\xi) = \frac{1}{\sqrt{2\pi}} \exp(\frac{-\xi^2}{2})$. Next, the odd subcarriers in first half of $\boldsymbol{Y}$ are extracted, the mth subcarrier is multiplied by a gain $d_r(m)$, where $d_r(m)$ is determined by applied diversity scheme. After adding the subcarrier signals together, the decision variable is obtained as

$$\begin{aligned} v_r = & \alpha\rho \sum_{m=0}^{G-1} s_r^i c_r(m) d_r(m) H(2m+1) \\ & + \alpha\rho \sum_{m=0}^{G-1} \sum_{u=0, u\neq r}^{U-1} s_u^i c_u(m) d_r(m) H(2m+1) \\ & + \sum_{m=0}^{G-1} \sigma_{\mathrm{c}} d_r(m) H(2m+1) N_{\mathrm{CN}}(2m+1) \\ & + \sum_{m=0}^{G-1} \sigma_{\mathrm{A}} d_r(m) N_{\mathrm{CN}}(2m+1), \end{aligned} \tag{8}$$

where $H(2m+1)$ is the diagonal element of $\boldsymbol{H}$, here $\boldsymbol{H}$ is an $N \times N$ diagonal matrix defined as $\boldsymbol{H} = \boldsymbol{F}\boldsymbol{h}\boldsymbol{F}^{\mathrm{H}}$, $N_{\mathrm{CN}}(2m+1)$ is the sample of AWGN with zero mean and unity variance, σ_{A} is the standard deviation of the AWGN. In (8), the 1st term is the desired signal, the 2nd term is the multiuser interference (MUI), the 3rd and 4th terms are the Gaussian clipping noise and AWGN, respectively.

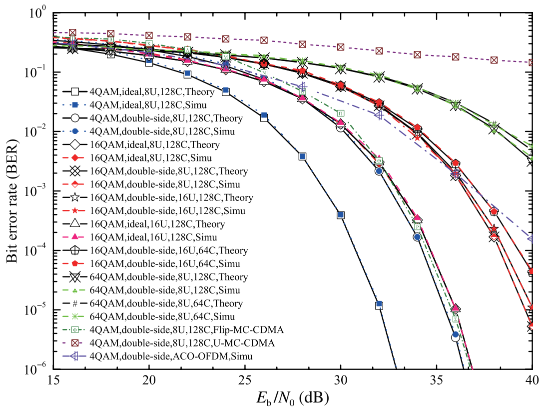

Figure 1 (Color online) BER performance of ACO-MC-CDMA.

Finally, v_r is demodulated using a maximum likelihood detector.

Performance analysis. The BER for M-QAM modulation can be approximated as [8, 9]

$$P_{\mathrm{b}} = \frac{4(\sqrt{M}-1)}{\sqrt{M}\log_2(M)}\mathrm{Q}\left(\sqrt{\frac{3\log_2(M)}{M-1}\frac{E_{\mathrm{b}}}{N_0}}\right), \quad (9)$$

where E_{b}/N_0 is the electrical signal to noise ratio (SNR) per bit.

In this letter, the orthogonality restoring combining (ORC) diversity scheme is considered. In a downlink channel, choosing the gain $d_r(m)$ as

$$d_r(m) = \frac{c_r(m)H^*(2m+1)}{|H(2m+1)|^2}, \quad (10)$$

the receiver can eliminate the MUI perfectly. Then E_{b}/N_0 can be calculated from (8) as

$$\frac{E_{\mathrm{b}}}{N_0} = \frac{(\alpha\rho G)^2}{\log_2(M)\left(G\sigma_{\mathrm{c}}^2 + \sum_{m=0}^{G-1}\frac{\sigma_{\mathrm{A}}^2}{|H(2m+1)|^2}\right)}. \quad (11)$$

Simulation results. The orthogonal gold (OG) codes are employed for different users. The simulated multipath channel is given as $h(t) = \sum_{l=0}^{L-1} h_l\delta(t - l\triangle\tau)$, where h_l is obtained by uniformly sampling the ceiling bounce model, $l\triangle\tau$ represents the delay of the lth path, and $h(t)$ is normalized so that $\|h(t)\|^2 = 1$. The following parameters are used: $D_{\mathrm{rms}} = 8$ ns, $L = 15$, $\sigma_0 = 0.2$, $B_{\mathrm{DC}} = 0.06$, $i_{\min} = 0.1$ and $i_{\max} = 1$.

Figure 1 shows the BER performance of ACO-MC-CDMA for ORC. It can be seen that the theoretical results agree well with the simulation results. Since the Gaussian clipping noise is added independently of the QAM symbol in (8), the higher order modulation is more vulnerable to signal clipping. Due to the MUI eliminated, as the number of the users increases from 8 to 16, the BER remains almost unchanged in presence of double-sided clipping. Thus, increasing the number of users does not dramatically impact on the BER. As the length of spreading codes increases from 64 to 128, BER performance is improved slightly. In addition, ACO-MC-CDMA gives the best performance, whereas U-MC-CDMA performs worst within the depicted SNR range. This is because Flip-MC-CDMA and U-MC-CDMA systems prove to be more vulnerable to signal clipping, and U-MC-CDMA systme is more vulnerable to diffused channel like U-OFDM. It is expected that ACO-MC-CDMA delivers a better BER performance as compared to ACO-OFDM, which can be explained that ACO-MC-CDMA combines the energy scattered in the frequency domain at the receiver so that the diversity gain is obtained.

Conclusion. The proposed ACO-MC-CDMA system is analyzed in this letter. Theoretical expression of SNR for ORC is derived in the presence of signal clipping. Simulation results show that the proposed system has the best BER performance compared to ACO-OFDM, Flip-MC-CDMA and U-MC-CDMA systems.

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