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Low-complexity Receiver for HACO-OFDM in Optical Wireless Communications

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***Abstract*—In this letter, a low-complexity receiver (Rx) is proposed for hybrid asymmetrically clipped optical orthogonal frequency division multiplexing (HACO-OFDM). Owing to the special time-domain property of HACO-OFDM, overlaid asymmetrically clipped OFDM (ACO-OFDM) and pulse-amplitude-modulated discrete multitoned (PAM-DMT) signals can be distinguished in the time domain to reduce its computational complexity. Besides, the near-optimal optical power allocation is further applied to optimize the proposed system performance. Theoretical analysis and simulation results demonstrate that, the proposed Rx can achieve nearly the same bit error rate (BER) performance as the BER-optimal iterative Rx but with an effective complexity reduction, thus demonstrating its application potential in high-speed optical communication systems.**

***Index Terms*—HACO-OFDM, computational complexity, optical power allocation, BER**

I. Introduction

USING the visible band to transmit information is better known as visible light communications (VLC), which is part of the optical wireless communications [1]. With solid-state light-emitting diodes (LEDs) being widely used for general lighting and signage in both indoor and outdoor environments, there is a golden opportunity to use them as an optical antenna in VLC. In recent years, we are seeing a growing interest in development and applications of VLC in academia and industry. Compared with the conventional radio frequency (RF) wireless communications, VLC offer many advantages such as unregulated bandwidth, high-level security, environment-friendly (i.e., lower energy usage), low radiation levels, and multiple features (data communications, localization, and sensing). Such features, make VLC highly attractive for smart environments, Internet of things, device to device communications, inter-and intra-chip communications, etc [1,2].

In order to improve the spectral efficiency of VLC, a series of hybrid modulation schemes have been proposed recently [3-5]. In [3], asymmetrically clipped DC biased orthogonal frequency division multiplexing (ADO-OFDM), a new version of OFDM based on a combination of ACO-OFDM and DCO-OFDM was proposed to improve the optical power performance compared with the conventional OFDM techniques. In [4], pulse-amplitude-modulated discrete multitone-based hybrid optical OFDM (PHO-OFDM), where high-order 2-dimensional QAM is used in place of a 1-dimensional PAM on even subcarriers to improve the spectral efficiency of pulse-amplitude-modulated discrete multitone (PAM-DMT). Although schemes in [3,4] have the ability to improve the spectral efficiency of ACO-OFDM and PAM-DMT, the output signals are bipolar, thus the need for the additional DC-biasing in intensity modulation (IM). In [5], a hybrid ACO-OFDM and PAM-DMT without DC-biasing, which is less complex, power-efficient and higher data rates, was proposed as hybrid asymmetrically clipped optical orthogonal frequency division multiplexing (HACO-OFDM). However, the proposed system in [5] is limited by the clipping noise. Although the iterative receiver (Rx) shows the best bit error rate (BER) performance [6], its computational complexity is much higher than the conventional Rx, which is not preferred in high-speed communication scenarios.

Therefore, we propose a novel low-complexity Rx with a competitive BER performance in this letter. Unlike the conventional and iterative Rxs extract the ACO-OFDM and PAM-DMT components in the frequency domain, the proposed Rx distinguishes the ACO-OFDM and PAM-DMT signals by the time-domain processing.

The remainder of this letter is organized as follows. Section II introduces the proposed Rx with near-optimal optical power allocation and its computational complexity analysis. BER comparisons between the proposed and the other existing Rxs are given and discussed in section III. Finally, we conclude this letter in section VI.

II. The proposed HACO-OFDM systems

A. Rx of HACO-OFDM

The block diagram of conventional HACO-OFDM Rx is shown in Fig. 1. Since the clipping noise of odd subcarriers only falls on even subcarriers and the clipping noise of even subcarriers does not affect odd subcarriers, the ACO-OFDM branch can be detected directly and the PAM-DMT branch can be thereafter demodulated following removal of the clipping distortion of ACO-OFDM [5].

Since the conventional Rx requires additional inverse fast Fourier transform (IFFT) and FFT operations to regenerate the frequency-domain clipping distortion of the ACO-OFDM

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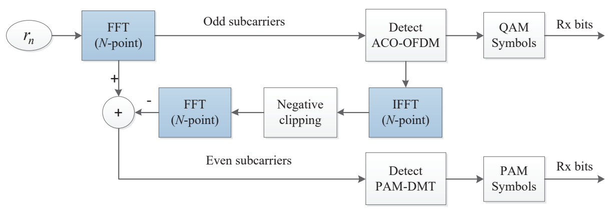

Fig. 1: Block diagram of the conventional HACO-OFDM Rx

branch, its computational complexity can be high. As the ACO-OFDM branch and PAM-DMT branch of HACO-OFDM have the special antisymmetric and periodic property, the separation of ACO-OFDM branch and PAM-DMT branch can be realized by subtraction in the time domain with reduced complexity compared with the frequency-domain counterpart. The block diagram of the proposed Rx is given in Fig. 2.

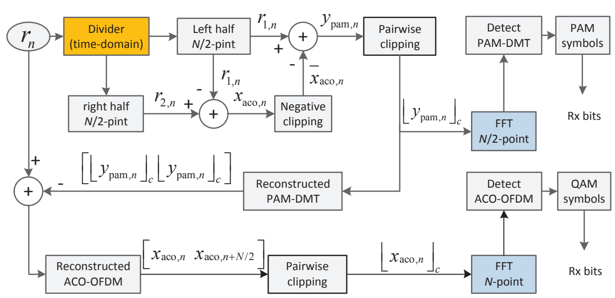

Fig. 2: Block diagram of the proposed HACO-OFDM Rx

Firstly, the received time-domain signal is split into two parts as $r_{1,n}$ and $r_{2,n}$, respectively. Since the time-domain ACO-OFDM and PAM-DMT have anti-symmetry and periodic property respectively, the bipolar half-frame ACO-OFDM $x_{\mathrm{aco},n}$ can be obtained by subtraction operation as $r_{1,n} - r_{2,n}$ [7]. Next, the first half-frame unipolar ACO-OFDM $\overline{x}_{\mathrm{aco},n}$ can be obtained by negative clipping operation. Therefore, the reconstructed unipolar PAM-DMT can be given by:

$$y_{\mathrm{pam},n} = r_{1,n} - \overline{x}_{\mathrm{aco},n}, \tag{1}$$

where $n = 0, 1, 2, ..., N/2 - 1$. After applying pairwise clipping on $y_{\mathrm{pam},n}$, we have:

$$\lfloor y_{\mathrm{pam},n} \rfloor_c = \begin{cases} y_{\mathrm{pam},n} I(y_{\mathrm{pam},n} \geq y_{\mathrm{pam},N/2-n}) \\ y_{\mathrm{pam},N/2-n} I(y_{\mathrm{pam},n} < y_{\mathrm{pam},N/2-n}) \end{cases}, \tag{2}$$

where $n = 1, 2, ..., N/4 - 1$ and $I(.)$ is an indicator function with $I(A) = 1$ if the event A is true and $I(A) = 0$ otherwise. In addition, $\lfloor y_{\mathrm{pam},0} \rfloor_c$ and $\lfloor y_{\mathrm{pam},N/4} \rfloor_c$ should be set to zero. Therefore, the noise of PAM-DMT branch can be reduced [6-7]. Following a $N/2$-point FFT, the PAM-DMT signal can be demodulated. Note that, the frequency-domain subcarriers should be multiplied by 4 to compensate for the power loss induced by the half-frame and negative clipping. Since PAM-DMT has a periodic property, ACO-OFDM can be reconstructed again in the time domain by subtracting PAM-DMT from the received signal r_n as given by:

$$\begin{cases} \widetilde{x}_{\mathrm{aco},n} = r_n - \lfloor y_{\mathrm{pam},n} \rfloor_c \\ \widetilde{x}_{\mathrm{aco},n+N/2} = r_{n+N/2} - \lfloor y_{\mathrm{pam},n} \rfloor_c \end{cases}, \tag{3}$$

where $n = 0, 1, 2, ..., N/2 - 1$. Since the time-domain ACO-OFDM signal is still antisymmetric, we use another pairwise clipping to further reduce the noise of ACO-OFDM branch as given by :

$$\left\lfloor \widetilde{x}_{\mathrm{aco},n} \right\rfloor_c = \begin{cases} \widetilde{x}_{\mathrm{aco},n} I(x_{\mathrm{aco},n} \geq x_{\mathrm{aco},n+N/2}) \\ \widetilde{x}_{\mathrm{aco},n+N/2} I(x_{\mathrm{aco},n} < x_{\mathrm{aco},n+N/2}) \end{cases}, \tag{4}$$

where $n = 0, 1, 2, ..., N/2 - 1$. After twice pairwise clipping, the total BER performance of HACO-OFDM can be further improved.

B. Near-optimal Power Allocation

As the different power factors of ACO-OFDM and PAM-DMT branches will affect the whole BER performance of HACO-OFDM, the near-optimal optical power allocation scheme is derived in this subsection. The BER equations of ACO-OFDM and PAM-DMT in HACO-OFDM are first given by [6,8]:

$$P_{\mathrm{b,ACO}} \approx \frac{4(\sqrt{M_1} - 1)}{\sqrt{M_1} \log_2 M_1} Q(\sqrt{\frac{3}{M_1 - 1} \frac{E_{s1}}{N_0}}), \tag{5}$$

$$P_{\mathrm{b,PAM}} \approx \frac{2(M_2 - 1)}{M_2 \log_2 M_2} Q(\sqrt{\frac{6}{M_2^2 - 1} \frac{E_{s2}}{N_0}}), \tag{6}$$

where M_1 and M_2 are the constellation size of QAM and PAM, and E_{s1} and E_{s2} denote the electrical energy of ACO-OFDM and PAM-DMT per symbols, respectively. At the same time, $Q(k)$ satisfy $Q(k) = \frac{1}{\sqrt{2\pi}} \int_k^\infty e^{-\frac{x^2}{2}} dx$. Note that, the total optical power is assumed to be constant and can be normalized to unity. Therefore, for ACO-OFDM the factor of optical power is defined as:

$$\alpha = \frac{P_{\mathrm{o,aco}}}{P_{\mathrm{o,aco}} + P_{\mathrm{o,pam}}} = P_{\mathrm{o,aco}}, \tag{7}$$

where $P_{\mathrm{o,aco}}$ and $P_{\mathrm{o,pam}}$ denote the normalized optical power allocated to the ACO-OFDM and PAM-DMT, respectively. Besides, they satisfy $E_{\mathrm{s1}} = \pi P_{\mathrm{o,aco}}^2$ and $E_{\mathrm{s2}} = \pi P_{\mathrm{o,pam}}^2$. Therefore, the total BER performance of HACO-OFDM can be given as:

$$P_{\mathrm{b,HACO}} = \frac{\frac{N}{4} \log_2 M_1 P_{\mathrm{b,ACO}} + (\frac{N}{4} - 1) \log_2 M_2 P_{b,PAM}}{\frac{N}{4} \log_2 M_1 + (\frac{N}{4} - 1) \log_2 M_2}. \tag{8}$$

Assume N is large enough, (8) can be approximated to:

$$\begin{array}{l} P_{\mathrm{b,HACO}} \approx \frac{4(\sqrt{M_1}-1)}{\sqrt{M_1} \log_2 (M_1 M_2)} Q(\alpha \sqrt{\frac{3\pi}{N_0 (M_1 - 1)}}) \\ \quad + \frac{2(M_2 - 1)}{M_2 \log_2 (M_1 M_2)} Q((1-\alpha) \sqrt{\frac{6\pi}{N_0 (M_2^2 - 1)}}) \end{array}. \tag{9}$$

As the second derivative of $P_{\mathrm{b,HACO}}$ is $P''_{\mathrm{b,HACO}}$, and $P''_{\mathrm{b,HACO}} > 0$ for $0 < \alpha < 1$, we can obtain the minimum of $P_{\mathrm{b,HACO}}$ on the condition that the derivative of (9) is zero. Since N_0 can be ignored when the SNR is large enough, we can have the relationship of α and the modulation combinations (M_1, M_2) as:

$$\alpha^2 (M_2^2 - 1) - 2(1 - \alpha)^2 (M_1 - 1) = 0. \tag{10}$$

According to (10), we can obtain the near-optimal power allocation of HACO-OFDM.

C. Computational complexity

Since the transmission time of the optical signal is very short (the distance of indoor VLC system is usually smaller than 5 m), the main delay of communication speed comes from the processing time of the receiver. Therefore, the computation complexity is a key factor in high-speed communication systems, we outline complexity comparisons of the proposed, conventional, and iterative Rxs in this subsection. Here, we first define the computation complexity as the number of complex-value multiplications in FFT/IFFT as adopted in [9]. Therefore, the N-point IFFT and FFT operations can be thought of as $O(N\log_2 N)$ and $O(N/2\log_2 N)$ respectively. As the conventional HACO-OFDM Rx requires two N-point FFT and one N-point IFFT operations, the computation complexity can be given by:

$$\begin{aligned} O(\mathrm{Conv}) = & O(N\log_2 N) + 2O(N/2\log_2 N) \\ & = 2O(N\log_2 N) \end{aligned} . \tag{11}$$

As for the iterative receiver [6], its BER performance can be improved by increasing the iterations. However, twice iterations usually result in a very tiny SNR improvement ($<$ 0.3 dB) than the once iteration, but with nearly twice times computational complexity increase. Therefore, we choose the iterative Rx with once iteration for comparisons in this paper. As once iterations require two N-point FFT and two N-point IFFT operations, the computational complexity can be given as:

$$\begin{aligned} O(\mathrm{1st-Iter}) &= 2O(N\log_2 N) + 2O(N/2\log_2 N) \\ &= 3O(N\log_2 N) \end{aligned} . \tag{12}$$

Since the proposed Rx requires only one single N- and $N/2$-point FFT operations at the Rx, the computational complexity can be given by:

$$\begin{aligned} O(\Pr\mathrm{op}) &= O(N/2\log 2(N)) + O(N/4\log_2(N/2)) \\ &= 1/2O(N\log 2(N)) + 1/4O(N\log_2(N/2)) \end{aligned} . \tag{13}$$

From the above complexity equations, we can conclude that the proposed Rx has the lowest computational complexity than the conventional and iterative Rxs in terms of complex-value multiplexing of FFT and IFFT. Assume $N = 512$, complexity reduction ratio between the proposed Rx and iterative Rx is nearly 75.9$\%$ [7].

III. Simulation results

The numerical simulations are performed on the computer with AMD Ryzen Threadripper 2990WX 32-core processer and 64-GB RAM. The near-optimal optical power allocations for the modulation combinations of 16QAM with 16PAM, 64QAM with 16PAM, and 64QAM-32PAM are first calculated by (13) and shown in Table I.

In order to demonstrate the near-optimal power allocation, the BER results of the above modulation combinations with different power allocations are drawn in Fig.3. We can clearly see that, the derived near-optimal optical power allocation factor always display the best BER performance for the proposed Rx, which further verifies the optical power allocation theory. Note that, the following BER comparisons between the proposed Rx and other Rxs are all performed with the near-optimal power allocation.

TABLE I: The near-optimal optical power allocation factors

Order of ACO-OFDM branch	Order of PAM-DMT branch	Power factor α
16-QAM	16-PAM	0.26
64-QAM	16-PAM	0.41
64-QAM	32-PAM	0.26

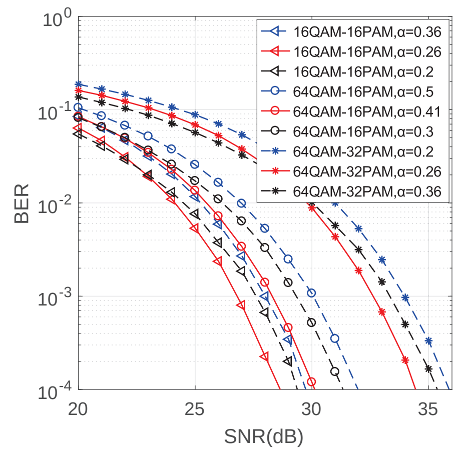

Fig. 3: BER comparisons of the proposed Rx under the different optical power allocations

As the total BER of HACO-OFDM is affected by the ACO-OFDM branch and PAM-DMT branch, the BER results of the PAM-DMT branch with the proposed, conventional and iterative Rxs are first drawn in Fig.4. It can be seen that the proposed Rx can achieve significant BER performance gain to the conventional Rx in the PAM-DMT branch. However, as the proposed Rx only uses the time-domain signal separation method, it is inherently weak at removing the constellation noise compared with the iterative Rx. Therefore, the BER performance of the proposed Rx is inferior to the iterative Rx in the PAM-DMT branch.

Next, the BER results of the ACO-OFDM branch with the conventional, iterative, and proposed Rxs are given in Fig.5. Since the conventional and 1st iterative Rxs use the same way to demodulate the ACO-OFDM branch, their BER curves fit well all the time. On the other hand, the proposed Rx shows better BER performance than the other two Rxs in the ACO-OFDM branch, which is nearly 1 dB, 1.1 dB, and 1.4 dB improvement respectively for the constellation combinations of 16QAM with 16PAM, 64QAM with 16PAM, and 64QAM with 32 PAM at the BER of 10^{-4}. The performance gain is mainly obtained from the time-domain pairwise clipping operation employed in the reconstructed ACO-OFDM branch.

Finally, the total BER results of HACO-OFDM with different Rxs are depicted in Fig.6. It can be seen that, at the BER of 10^{-4}, the proposed Rx can achieve about 2.8, 2.3 dB, and 3.0 dB SNR gain compared with the conventional Rx respectively for 16QAM with 16PAM, 64QAM with 16PAM, and 64QAM with 32PAM. Besides, we also find that the proposed Rx shows

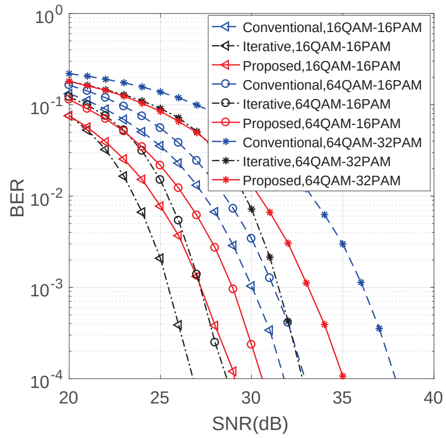

Fig. 4: BER comparisons of the PAM-DMT branch with the conventional, 1st iteration and proposed Rxs

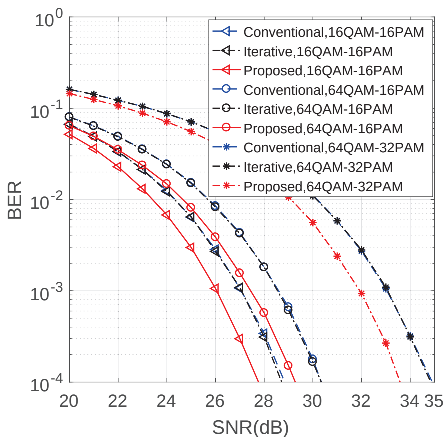

Fig. 5: BER comparisons of the ACO-OFDM branch with the conventional, 1st iteration and proposed Rxs

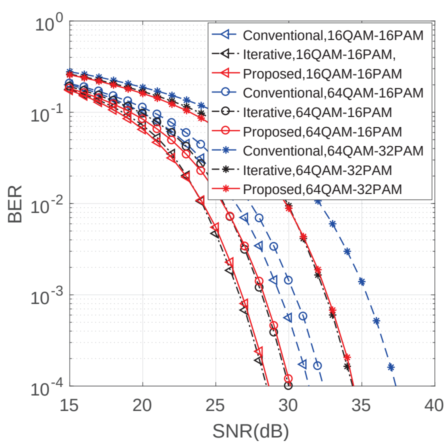

Fig. 6: BER comparisons of the total HACO-OFDM with the conventional, 1st iteration and proposed Rxs

a tiny SNR penalty compared with the 1st iterative Rx under the above constellation combinations. As the SNR differences are always smaller than 0.1 dB, the BER performance of the proposed Rx can be acceptable in most communication scenarios.

Since the computational complexity analyzed by the multiplication times of FFT and IFFT cannot fully reflect the total complexity of the above Rxs[9-10], the running times of the conventional, iterative, and proposed Rxs with the above constellation combinations are given in Table II.

TABLE II: Running time comparisons based on modulation combination M_1-QAM with M_2-PAM for different Rxs

Modulation	No. of symbols	Conventional(s)	1st iteration(s)	Proposed(s)
64-16	2.286×10^{6}	21.78	22.61	20.18
64-32	2.286×10^{6}	28.06	29.88	26.05
64-16	4.572×10^{6}	42.79	45.36	39.69
64-32	4.572×10^{6}	55.71	59.21	51.19

From Table II,we can see the time-domain processing of the proposed scheme only results in $10\% \sim 15\%$ time reduction compared with the 1st iterative Rx. This is because the total computational complexity of Rx is also affected by the constellation demodulation (e.g. maximum likelihood algorithm). However, the proposed Rx has the shortest running time among these three Rxs.

IV. Conclusion

In this letter, a low-complexity Rx is proposed for the classical HACO-OFDM systems. The noise and estimation error of HACO-OFDM can be successfully reduced by the time-domain processing with reduced complexity compared with the conventional and iterative Rxs. Besides, the near-optimal optical power allocation is also derived and applied to optimize the total BER performance. Comprehensive comparisons and analysis indicate that, the proposed Rx offer almost the same BER performance as the BER-optimal iterative-based Rx but with reduced complexity, thus making it much more attractive in high-speed optical communications.

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