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Iterative Receiver for Flip-OFDM in Optical Wireless Communication

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***Abstract*—To guarantee nonnegative signals in optical wireless communication (OWC) systems, flipped orthogonal frequency division multiplexing (Flip-OFDM) transmits the positive and negative parts of the signal over two consecutive OFDM subframes (positive subframe and negative subfame, respectively). The conventional receiver for Flip-OFDM recovers the data by subtracting the negative subframe from the positive one. However, the signal analysis shows that the signals in the two subframes both contain the information of the transmitted data and can be used together to decode the data. An iterative receiver is then proposed to improve the transmission performance of Flip-OFDM by exploiting the signals in both subframes. Simulation results show that the proposed iterative receiver provides significant signal to noise ratio (SNR) gain over the conventional receiver. Moreover, the iterative receiver also outperforms the existing advanced receiver.**

***Index Terms*—Optical wireless communication (OWC), orthogonal frequency division multiplexing (OFDM), Flip-OFDM, iterative receiver.**

I. Introduction

WITH the widespread deployment of light-emitting diodes (LEDs), optical wireless communication (OWC) has attracted an increasing interest in academia and industry recently [1]. Due to its distinct advantages such as rich spectrum resources and high communication security, OWC has been anticipated to be an attractive alternative to radio frequency (RF) systems, especially in indoor scenarios. In addition, the convergence of illumination and communication makes OWC one of the most important green technologies [2].

In order to achieve high data rates and alleviate inter-symbol interference (ISI), orthogonal frequency division multiplexing (OFDM) has been employed in OWC [3]–[5]. Since intensity modulation and direct detection (IM/DD) is commonly used in OWC systems, the transmitted signals must be real and nonnegative. Real time-domain signals can be obtained by imposing Hermitian symmetry on the OFDM subcarriers. Furthermore, to deal with the issue of bipolarity in OFDM signals, several OFDM schemes have been proposed for OWC. For example, direct current (DC) biased optical OFDM (DCO-OFDM) [6], asymmetrically clipped optical OFDM (ACO-OFDM) [7] and pulse-amplitude-modulated discrete multitone (PAM-DMT) [8]. There is a price paid with these three schemes. DCO-OFDM adds a DC bias to the OFDM symbols, which increases the power dissipation of the signal significantly. ACO-OFDM and PAM-DMT do not need DC bias because of the clipping operation, but each has only half the spectral efficiency of DCO-OFDM. The authors in [9] proposed a novel OFDM technique named as Flip-OFDM in which positive and negative parts of the signal are separately transmitted over two consecutive OFDM subframes. In [10], it was shown that Flip-OFDM can be modified to approach the spectral efficiency of DCO-OFDM without biasing, which contributes to the practical applications of Flip-OFDM in OWC.

In the conventional receiver for Flip-OFDM, the data is recovered by subtracting the negative signal block from the positive one [9]. This method is simple and straightforward. However, it increases the noise variance of the received symbols, making the performance much worse than that of bipolar OFDM with the same modulation method. To improve the performance of Flip-OFDM, a time-domain noise filtering technique was proposed in [11] and investigated in [12]. However, the algorithm does not make full use of the signal structures. In this letter, an iterative receiver is proposed for Flip-OFDM by fully exploiting the structures of the received signals. Simulations confirm that the proposed iterative receiver is superior to other receivers.

Notations: Bold italic letters denote column vectors. More specifically, a lowercase letter such as $\boldsymbol{v}$ denotes a time-domain vector, and the uppercase letter such as $\boldsymbol{V}$ denotes the corresponding frequency-domain vector. A bold but not italic letter such as $\mathbf{A}$ denotes a matrix. Specially, $\mathbf{0}$ and $\mathbf{I}$ represent the zero and identity matrices with appropriate dimensions, respectively. $(\cdot)^*$, $(\cdot)^T$, $(\cdot)^H$, $|\cdot|$ and $\mathrm{sign}(\cdot)$ denote conjugate, transpose, Hermitian transpose, absolute value and sign (for convenience, $\mathrm{sign}(0) = 1$ is defined in this paper), respectively. The n-th element of a vector $\boldsymbol{v}$ is denoted by $v(n)$. $\mathrm{diag}(\boldsymbol{v})$ is a diagonal matrix with $\boldsymbol{v}$ on the main diagonal.

II. System Model

The block diagram of a Flip-OFDM transmitter with N subcarriers is illustrated in Fig. 1 [9]. To ensure that the time-domain signal is real in IM/DD systems, the input data vector $\boldsymbol{X} = [X(0), X(1), \cdots, X(N-1)]^T$ should satisfy the Hermitian symmetry property, i.e.,

$$X(k) = X^*(N-k),\, k = 1, 2, \cdots, N/2-1. \qquad (1)$$

Note that $X(0)$ and $X(N/2)$ are usually set to zero since the DC part of OFDM signal is left unused in practical applications. Hence, the time-domain signal vector $\boldsymbol{x}$ =

This work was supported in part by the National Nature Science Foundation of China under Grants 61223001 and 61372106, and in part by the National 863 High Technology Development Project under Grant 2013AA013601.

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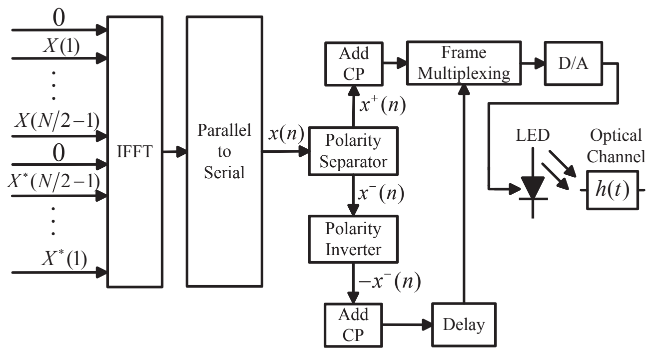

Fig. 1. Block diagram of a Flip-OFDM transmitter.

$[x(0), x(1), \cdots, x(N-1)]$ after inverse fast Fourier transform (IFFT) operation can be represented as

$$\begin{aligned} x(n) &= \frac{1}{\sqrt{N}} \sum_{k=0}^{N-1} X(k) \exp(j2\pi kn/N) \\ &= \frac{2}{\sqrt{N}} \sum_{k=1}^{N/2-1} \mathrm{Re}\left[X(k) \exp(j2\pi kn/N)\right] \\ & n = 0, 1, \cdots, N-1. \end{aligned} \tag{2}$$

The signal $x(n)$, which is real and bipolar, can be decomposed as

$$x(n) = x^{+}(n) + x^{-}(n), \tag{3}$$

where the positive part $x^{+}(n)$ and the negative part $x^{-}(n)$ are defined as

$$x^{+}(n) = \begin{cases} x(n), & x(n) \geq 0 \\ 0, & x(n) < 0 \end{cases}, \tag{4}$$

$$x^{-}(n) = \begin{cases} x(n), & x(n) < 0 \\ 0, & x(n) \geq 0 \end{cases}. \tag{5}$$

To guarantee a nonnegative time-domain signal, the two components $\boldsymbol{x}^{+} = [x^{+}(0), x^{+}(1), \cdots, x^{+}(N-1)]^{T}$ and $\boldsymbol{x}^{-} = [x^{-}(0), x^{-}(1), \cdots, x^{-}(N-1)]^{T}$ are separately transmitted over two successive OFDM subframes. The positive component $\boldsymbol{x}^{+}$ is transmitted in the first subframe (positive subframe), while the flipped negative component $-\boldsymbol{x}^{-}$ is transmitted in the second subframe (negative subframe). An example of the time-domain signals in Flip-OFDM is illustrated in Fig. 2. After propagating through the optical channel, the unipolar time-domain signal is received by a photodetector. Assuming the channel impulse response $\boldsymbol{h} = [h(0), h(1), \cdots, h(N-1)]^{T}$ is constant over two consecutive OFDM subframes, the received signal vectors in the frequency domain are given by

$$\boldsymbol{Y}^{+} = \mathbf{H}\boldsymbol{X}^{+} + \boldsymbol{Z}^{+}, \tag{6}$$

$$\boldsymbol{Y}^{-} = -\mathbf{H}\boldsymbol{X}^{-} + \boldsymbol{Z}^{-}, \tag{7}$$

where $\mathbf{H} = \mathrm{diag}(\mathbf{W}_N \boldsymbol{h})$, $\boldsymbol{X}^{+} = \mathbf{W}_N \boldsymbol{x}^{+}$, $\boldsymbol{X}^{-} = \mathbf{W}_N \boldsymbol{x}^{-}$, $\mathbf{W}_N$ is the $N \times N$ discrete Fourier transform (DFT) matrix, $\boldsymbol{Z}^{+}$ and $\boldsymbol{Z}^{-} \sim N(\mathbf{0}, \sigma^2 \mathbf{I})$, represent the noise vectors of the two subframes, respectively.

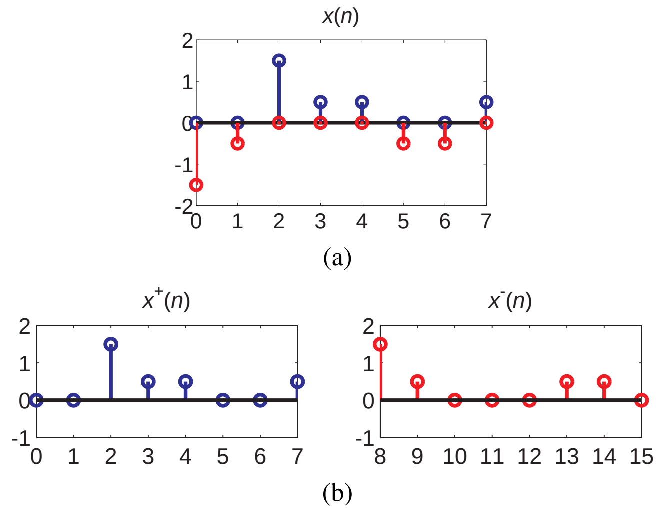

Fig. 2. An example of the time domain signals in Flip-OFDM ($N = 8$). (a) $x(n)$. (b) $x^{+}(n)$ and $x^{-}(n)$.

III. RECEIVER DESIGN

A. Conventional Receiver

In [9], a receiver is proposed for Flip-OFDM. This receiver subtracts the negative subframe from the positive one to decode the data. To facilitate the following descriptions, the receiver proposed in [9] is named as the conventional receiver in this paper. Subtracting (7) from (6) yields

$$\begin{aligned} \boldsymbol{Y}^{+} - \boldsymbol{Y}^{-} &= \mathbf{H}\boldsymbol{X}^{+} + \mathbf{H}\boldsymbol{X}^{-} + \boldsymbol{Z}^{+} - \boldsymbol{Z}^{-} \\ &= \mathbf{H}\boldsymbol{X} + \boldsymbol{Z}, \end{aligned} \tag{8}$$

where $\boldsymbol{Z} = \boldsymbol{Z}^{+} - \boldsymbol{Z}^{-}$. Then, the conventional receiver can be described as

$$\hat{\boldsymbol{X}}_{\mathrm{conv}} = \mathrm{dec}[\mathbf{H}^{-1}\boldsymbol{Y}^{+} - \mathbf{H}^{-1}\boldsymbol{Y}^{-}]. \tag{9}$$

where $\mathrm{dec}[\cdot]$ represents the decision of the detector.

B. Proposed Receiver

The conventional receiver is simple and straightforward, but it does not fully exploit the structures of the received signals. In the following, a new receiver is proposed by establishing the relationship between the received signals $\boldsymbol{Y}^{+}$, $\boldsymbol{Y}^{-}$ and the input data $\boldsymbol{X}$.

Since $|\boldsymbol{x}|$ can be expressed as

$$\begin{aligned} |\boldsymbol{x}| &= \mathbf{S}(\boldsymbol{X})\boldsymbol{x} \\ &= \mathbf{S}(\boldsymbol{X})\mathbf{W}_N^{H}\boldsymbol{X}, \end{aligned} \tag{10}$$

where $\mathbf{S}(\boldsymbol{X})$ is defined as

$$\begin{aligned} \mathbf{S}(\boldsymbol{X}) &= \mathrm{diag}\{\mathrm{sign}(\boldsymbol{x})\} \\ &= \mathrm{diag}\{\mathrm{sign}(\mathbf{W}_N^{H}\boldsymbol{X})\}, \end{aligned} \tag{11}$$

one can rewrite (4) as

$$\begin{aligned} \boldsymbol{x}^{+} &= \frac{\boldsymbol{x} + |\boldsymbol{x}|}{2} \\ &= \frac{\boldsymbol{x} + \mathbf{S}(X)\mathbf{W}_N^{H}\boldsymbol{X}}{2}. \end{aligned} \tag{12}$$

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From (12) and $\boldsymbol{X}^+ = \mathbf{W}_N\boldsymbol{x}^+$, it is easy to get

$$\begin{aligned}\boldsymbol{X}^+ &= \mathbf{W}_N \frac{\boldsymbol{x} + \mathbf{S}(\boldsymbol{X})\mathbf{W}_N^H\boldsymbol{X}}{2} \\ &= \frac{\mathbf{I} + \mathbf{W}_N\mathbf{S}(\boldsymbol{X})\mathbf{W}_N^H}{2}\boldsymbol{X}.\end{aligned} \tag{13}$$

Substituting (13) into (6) gives

$$\boldsymbol{Y}^+ = \frac{\mathbf{H} + \mathbf{H}\mathbf{W}_N\mathbf{S}(\boldsymbol{X})\mathbf{W}_N^H}{2}\boldsymbol{X} + \mathbf{Z}^+. \tag{14}$$

Similarly, by using (7) and $\boldsymbol{x}^- = \frac{\boldsymbol{x}-|\boldsymbol{x}|}{2}$, the relationship between $\boldsymbol{Y}^-$ and $\boldsymbol{X}$ can be derived as

$$\boldsymbol{Y}^- = \frac{\mathbf{H}\mathbf{W}_N\mathbf{S}(\boldsymbol{X})\mathbf{W}_N^H - \mathbf{H}}{2}\boldsymbol{X} + \mathbf{Z}^-. \tag{15}$$

(14) and (15) show the structures of the received signals in the two subframes, respectively. They can be concatenated to

$$\begin{bmatrix} \boldsymbol{Y}^+ \\ \boldsymbol{Y}^- \end{bmatrix} = \frac{1}{2}\begin{bmatrix} \mathbf{H} + \mathbf{H}\mathbf{W}_N\mathbf{S}(\boldsymbol{X})\mathbf{W}_N^H \\ \mathbf{H}\mathbf{W}_N\mathbf{S}(\boldsymbol{X})\mathbf{W}_N^H - \mathbf{H} \end{bmatrix}\boldsymbol{X} + \begin{bmatrix} \mathbf{Z}^+ \\ \mathbf{Z}^- \end{bmatrix}. \tag{16}$$

Based on (16), several methods are available to recover the data $\boldsymbol{X}$. In this paper, the zero-forcing (ZF) estimator is chosen for its low complexity and simplicity [13]. Denoting

$$\mathbf{G}(\boldsymbol{X}) = \frac{1}{2}\begin{bmatrix} \mathbf{H} + \mathbf{H}\mathbf{W}_N\mathbf{S}(\boldsymbol{X})\mathbf{W}_N^H \\ \mathbf{H}\mathbf{W}_N\mathbf{S}(\boldsymbol{X})\mathbf{W}_N^H - \mathbf{H} \end{bmatrix}, \tag{17}$$

the ZF matrix can be derived as

$$\mathbf{T}_{\mathrm{ZF}}(\boldsymbol{X}) = \left[\mathbf{G}^H(\boldsymbol{X})\mathbf{G}(\boldsymbol{X})\right]^{-1}\mathbf{G}^H(\boldsymbol{X}). \tag{18}$$

Then, the estimate of $\boldsymbol{X}$ is given by

$$\tilde{\boldsymbol{X}} = \mathbf{T}_{\mathrm{ZF}}(\boldsymbol{X})\begin{bmatrix} \boldsymbol{Y}^+ \\ \boldsymbol{Y}^- \end{bmatrix}. \tag{19}$$

The matrix $\mathbf{T}_{\mathrm{ZF}}(\boldsymbol{X})$ is dependent on the data vector $\boldsymbol{X}$, an iterative receiver can thus be proposed as

$$\hat{\boldsymbol{X}}^{(i)} = \begin{cases} \mathrm{dec}[\mathbf{H}^{-1}\boldsymbol{Y}^+ - \mathbf{H}^{-1}\boldsymbol{Y}^-], & i = 0 \\ \mathrm{dec}\left\{\mathbf{T}_{\mathrm{ZF}}(\hat{\boldsymbol{X}}^{(i-1)})\begin{bmatrix} \boldsymbol{Y}^+ \\ \boldsymbol{Y}^- \end{bmatrix}\right\}, & i = 1, \cdots, K \end{cases}, \tag{20}$$

where i is the index of iteration and K is the maximum number of iterations. Note that, when $K = 0$, the proposed iterative receiver in (20) is equivalent to the conventional receiver in (9).

Particularly, in line-of-sight (LOS) channels, the channel response can be expressed as

$$h(n) = c\delta(n), \tag{21}$$

where c is a constant and $\delta(n)$ is the dirac delta function. Since c scales the signal-to-noise ratio (SNR) only, without loss of generality, c is set to one to simplify the derivations [14]. In this case, one can obtain $\mathbf{H} = \mathbf{I}$, $\mathbf{G}^H(\boldsymbol{X})\mathbf{G}(\boldsymbol{X}) = \mathbf{I}$ and $\mathbf{T}_{\mathrm{ZF}}(\boldsymbol{X}) = \mathbf{G}^H(\boldsymbol{X})$. Therefore, the estimate of $\boldsymbol{X}$ can be simplified as

$$\begin{aligned}\tilde{\boldsymbol{X}}_{\mathrm{LOS}} =& \frac{1}{2}[\mathbf{I} + \mathbf{W}_N\mathbf{S}(\boldsymbol{X})\mathbf{W}_N^H]\boldsymbol{Y}^+ \\ &+ \frac{1}{2}[\mathbf{W}_N\mathbf{S}(\boldsymbol{X})\mathbf{W}_N^H - \mathbf{I}]\boldsymbol{Y}^-,\end{aligned} \tag{22}$$

and the iterative receiver becomes

$$\hat{\boldsymbol{X}}^{(i)}_{\mathrm{LOS}} = \begin{cases} \mathrm{dec}[\boldsymbol{Y}^+ - \boldsymbol{Y}^-], \quad i = 0 \\ \mathrm{dec}\left\{\frac{1}{2}[\mathbf{I} + \mathbf{W}_{\mathrm{N}}\mathbf{S}(\hat{\boldsymbol{X}}^{(i-1)}_{\mathrm{LOS}})\mathbf{W}_{\mathrm{N}}^{\mathrm{H}}]\boldsymbol{Y}^+ \right. \\ \left. \quad + \frac{1}{2}[\mathbf{W}_N\mathbf{S}(\hat{\boldsymbol{X}}^{(i-1)}_{\mathrm{LOS}})\mathbf{W}_N^H - \mathbf{I}]\boldsymbol{Y}^-\right\}, \; i = 1, \cdots, K \end{cases}. \tag{23}$$

By applying the additive property of DFT, the equivalent time-domain form of (22) can be expressed as

$$\begin{aligned}\tilde{x}_{\mathrm{LOS}}(n) =& \frac{1}{2}\{1 + \mathrm{sign}[x(n)]\}y^+(n) \\ &+ \frac{1}{2}\{\mathrm{sign}[x(n)] - 1\}y^-(n).\end{aligned} \tag{24}$$

It can be observed from (24) that in LOS channels, the main difference between the conventional receiver and the proposed receiver is the weight multiplied by $y^+(n)$ or $y^-(n)$. This is very useful for reducing the implementation cost of the proposed receiver in LOS channels.

C. Computational Complexity

In this subsection, we analyze the computational complexity of different receivers with order notation. Since the conventional receiver and the noise filtering receiver in [11] both include a single fast Fourier transform (FFT) operation, each of them has a complexity of $O(N \log N)$. For the proposed iterative receiver, the computational complexity is related to the channel characteristics. In non-line-of-sight (NLOS) channels, the iterative receiver has a complexity of $O(N^3)$ per iteration due to the matrix inversion of the ZF estimator. In LOS channels, the complexity of the iterative receiver is $O(N \log N)$ per iteration because the matrix inversion is avoided. It should be noted that the iterative receiver needs additional IFFT operations with complexity of $O(N \log N)$ to estimate $\mathbf{S}(\boldsymbol{X})$ in either LOS channels or NLOS channels.

IV. Numerical Results

In this section, the bit error rate (BER) performance of the proposed iterative receiver is evaluated by simulations. The number of subcarriers is 64, the length of cyclic prefix (CP) is 16, and 16-QAM is used for constellation mapping. Both LOS and NLOS channels are tested with the parameters of configuration A in [15]. For comparison, the BER curves corresponding to the noise filtering receiver and the case in (19) where the sign matrices are perfectly known ("lower bound") are also plotted.

Fig. 3 shows the BER performance of the iterative receiver with different iteration numbers. It can be seen that the performance of the iterative receiver can be improved by increasing the iteration number K. However, when $K \geq 2$, almost no additional gain can be achieved by further increasing K. Therefore, setting $K = 2$ is enough to achieve a noticeable performance gain while keeping relatively low complexity and delay.

Fig. 4 presents the BER performance of different receivers. It can be seen that the iterative receiver achieves a noticeable gain over the conventional receiver in any case. For example, in NLOS channels, the SNR gain over the conventional receiver is about 4 dB at the BER of 10^{-4}. The significant

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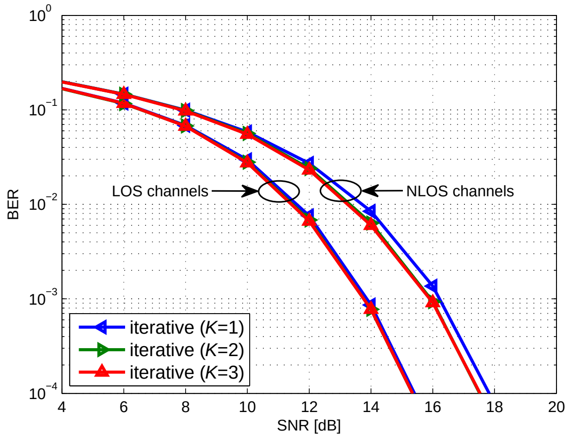
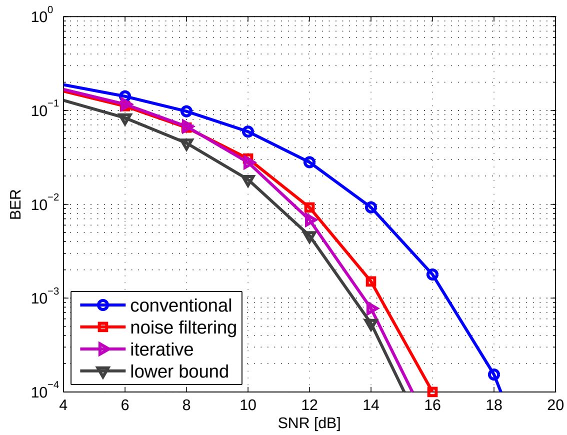
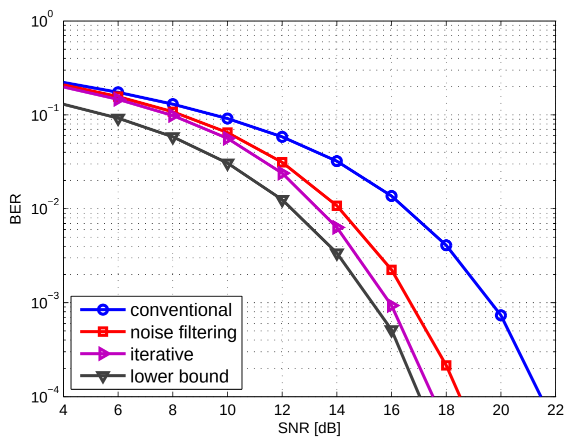


Fig. 3. BER performance of the iterative receiver with different iteration numbers.



(a)



(b)

Fig. 4. BER comparison of different receivers for Flip-OFDM. (a) LOS channels. (b) NLOS channels.

performance improvement results from the fact that the iterative receiver fully exploits the signal structures. Moreover, the proposed receiver also outperforms the noise filtering receiver. Specifically, at the BER of 10^{-4}, the iterative receiver provides 0.6 dB and 1 dB gains over the noise filtering receiver in LOS channels and NLOS channels, respectively.

V. Conclusion

In this paper, an iterative receiver is proposed for Flip-OFDM in IM/DD based OWC systems. In order to improve the receiver performance, the iterative receiver obtains the additional diversity gain by exploiting the signals in both the positive subframe and negative subfame. The simulation results show that the iterative receiver with only two iterations provides a significant SNR gain over the conventional receiver. Moreover, the receiver is also superior to the existing advanced receiver. It should be noted that the proposed iterative receiver can also be applied to the modified Flip-OFDM system proposed in [10] and the multiple-input multiple-output (MIMO) system proposed in [16].

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