# An Improved FD-DFE Structure for Downlink VLC Systems Based on SC-FDMA

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***Abstract*—Visible light communication (VLC) technology is considered to be an attractive solution for future wireless indoor multimedia communication (WIMC). To reduce the peak to average power ratio (PAPR), single carrier frequency division multiple access (SC-FDMA) with time domain asymmetrical clipping is applied for the downlink multi-user VLC system. An improved frequency domain decision feedback equalizer (FD-DFE) structure is proposed for the SC-FDMA based VLC system in this letter. Different kinds of subcarrier distributions are considered in the analyses and simulations. Thanks to the clipping noise information on the even subcarriers which can reduce the decision error, the proposed algorithm provides a lower bit error rate (BER) and better transmission performance compared with traditional schemes.**

***Index Terms*—Visible light communication (VLC), single carrier frequency division multiple access (SC-FDMA), frequency domain decision feedback equalization (FD-DFE).**

## I. Introduction

Visible light communication (VLC) has become an attractive choice for future wireless indoor multimedia communication (WIMC), saving the limited radio frequency resources. With the advantages of high transmission data rate, the absence of electromagnetic interference and the convenience of being applied on existing lighting facilities, VLC systems are applicable to a variety of complex indoor communication environments.

Light emitting diode (LED) and photo detector (PD) are used in a typical VLC system. Since the optical signal must be both real and non-negative, intensity modulation/direct detection (IM/DD) method is applied commonly. Compared with relatively simple modulation schemes such as on-off keying (OOK) and pulse-position modulation (PPM), orthogonal frequency division multiplexing (OFDM) is considered to be a better modulation technique for optical wireless communication, with the advantages of resistance to multipath channel and high spectrum efficiency. In an IM/DD VLC system, traditional OFDM signal needs to be adjusted to be real and non-negative which is called optical OFDM (O-OFDM). Several O-OFDM modulation methods are widely used in VLC systems. Normally, the transmitted signal is first processed to satisfy Hermitian symmetry, ensuring that the signal after inverse fast Fourier transform (IFFT) is real rather than complex. In DC biased O-OFDM (DCO-OFDM), the data symbols are carried on all the subcarriers. A DC bias is added to the OFDM signal in the time domain to form the non-negative signal. In asymmetrically clipped O-OFDM (ACO-OFDM), the data symbols are only assigned to the odd subcarriers, and the signal components lower than zero after IFFT are directly clipped, reserving the positive parts [1]. Compared with DCO-OFDM, ACO-OFDM without a DC bias has a better power efficiency. Furthermore, the clipping noise on the even subcarriers can help decode the transmitted data in ACO-OFDM [2].

In a VLC system, the nonlinearity of the LED will have a great influence on the transmission performance of O-OFDM with high peak to average power ratio (PAPR). Many researches have been done on PAPR reduction for O-OFDM. In this letter, single carrier frequency division multiple access (SC-FDMA) is applied for downlink multi-user VLC systems, which can reduce the PAPR while ensuring a high data rate [3].

In an optical SC-FDMA system, the most direct method for equalization is the first-order frequency domain linear equalizer (FD-LE) which will have a great performance degradation when there exist obvious spectral nulls on the channel frequency response (CFR). In [4], the single-carrier frequency-domain equalization (SC-FDE) based on minimum mean square error (MMSE) is proposed for ACO-OFDM. Moreover, SC-FDE methods for OOK based on MMSE are also well studied [5] [6], greatly improving the system performance. Decision feedback equalization (DFE) is a commonly used technique in a SC-FDMA system to avoid the effects of high frequency selective and deep fading channels [7] [8]. In [9], a DFE method is proposed for the optical SC-FDE schemes and the performance improvement in multi-tap indoor VLC channels is demonstrated.

In this letter, an improved FD-DFE structure for downlink VLC systems is proposed based on SC-FDMA. By using the clipping noise information of the signal after time domain asymmetrical clipping, the decision error of FD-DFE can be reduced remarkably, providing a lower bit error rate (BER) and improving the transmission performance. Different kinds of subcarrier allocations are also considered.

## II. System Model

Fig. 1 shows the diagrams of the transmitter and the receiver for the $k$th user in a downlink VLC system based on SC-FDMA. The double lined arrow represents vector signal, while the single lined arrow represents scalar signal (keep the same meaning below). Assume the number of the users is $K$, each user occupies $M$ subcarriers, thus there will be $P = KM$

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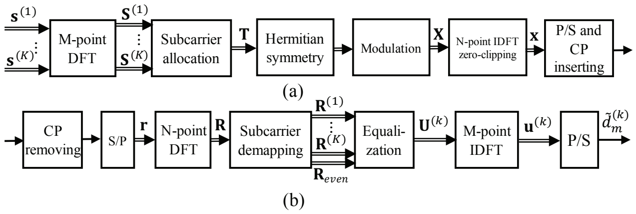


Fig. 1. Block diagram of the transmitter and the receiver for the $k$th user in a downlink VLC system based on SC-FDMA.

subcarriers that carry valid data symbols. The data symbols of the $k$th user is denoted as $\mathbf{s}^{(k)} = [s_0^{(k)}, s_1^{(k)}, ..., s_{M-1}^{(k)}]^{\mathrm{T}}$. After $M$-point discrete Fourier transform (DFT), $\mathbf{s}^{(k)}$ is transformed to $\mathbf{S}^{(k)}$. Then, $\mathbf{S}^{(k)}$ is assigned to a data vector $\mathbf{T}$ with a length of $P$ which is shown as follows [8]

$$\mathbf{T} = \sum_{k=1}^{K} \mathbf{D}^{(k)} \mathbf{S}^{(k)} \tag{1}$$

The size of the resource allocation matrix $\mathbf{D}^{(k)}$ for the $k$th user is $P \times M$. The resource allocation matrices of localized, distributed and frequency hoping (FH) allocation methods are as follows

$$\mathbf{D}_{Loc,(m,n)}^{(k)} = \begin{cases} 1, & m = (k-1)M + n, 0 \le n \le M-1 \\ 0, & \text{otherwise} \end{cases} \tag{2}$$

$$\mathbf{D}_{Dis,(m,n)}^{(k)} = \begin{cases} 1, & m = (k-1) + Kn, 0 \le n \le M-1 \\ 0, & \text{otherwise} \end{cases} \tag{3}$$

$$\mathbf{D}_{FH,(m,n)}^{(k)} = \begin{cases} 1, & m = \mathrm{FH}^{(k)}(n), 0 \le n \le M-1 \\ 0, & \text{otherwise} \end{cases} \tag{4}$$

where $\mathbf{D}_{Loc,(m,n)}^{(k)}$, $\mathbf{D}_{Dis,(m,n)}^{(k)}$ and $\mathbf{D}_{FH,(m,n)}^{(k)}$ represent three different allocation methods for the $k$th user respectively, and $(m,n)$ is the element index. $\mathrm{FH}^{(k)}$ is the frequency hoping function for the $k$th user. And a general expression of the resource allocation matrices can be expressed as

$$\mathbf{D}_{(m,n)}^{(k)} = \begin{cases} 1, & m = \mathrm{MP}^{(k)}(n), 0 \le n \le M-1 \\ 0, & \text{otherwise} \end{cases} \tag{5}$$

where $\mathbf{D}_{(m,n)}^{(k)}$ is the $(m,n)$ element of $\mathbf{D}^{(k)}$ and $\mathrm{MP}^{(k)}(n)$ is the subcarrier allocation function.

According to the properties of O-OFDM, the transmitted data vector must satisfy Hermitian symmetry. The data symbols are assigned only on the odd subcarriers to avoid the interference of clipping noise [1]. The data vector $\mathbf{T}$ can be transformed as follows

$$\mathbf{X} = [0, T_1, 0, T_2, ..., T_{P-1}, 0, T_{P-1}^*, 0, ..., T_2^*, 0, T_1^*]^{\mathrm{T}} \tag{6}$$

where $T_i$ is the $i$th component of $\mathbf{T}$ and $\mathbf{X}$ is the transmitted signal in frequency domain with a length of $N = 4P$. After $N$-point inverse DFT (IDFT) and zero-clipping, $\mathbf{X}$ is transformed to $\mathbf{x}$ in the time domain, which is both real and positive, and unaffected by the clipping noise which is only on the even subcarriers.

At the receiver, after the CP removing and $N$-point DFT, the received signal $\mathbf{r}$ is transformed to frequency domain as

$$\mathbf{R} = \mathrm{DFT}(\mathbf{r}) + \mathbf{V} \tag{7}$$

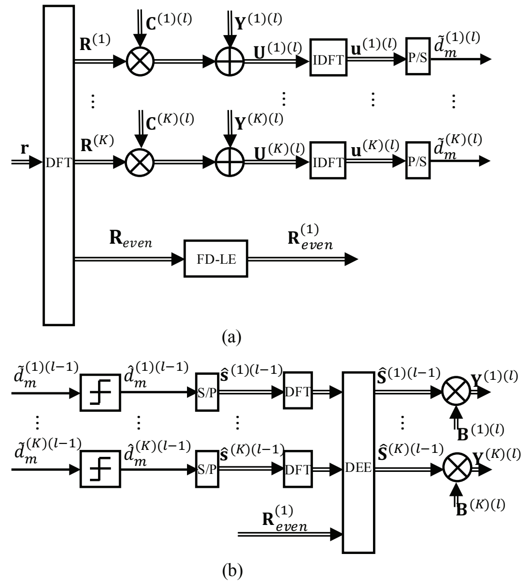


Fig. 2. Block diagram of the proposed FD-DFE structure.

where $\mathbf{V}$ is the noise vector and $\mathrm{DFT}(\mathbf{r}) = \mathbf{HX}$ with $\mathbf{H}$ the $N \times N$ CFR matrix whose diagonal elements are $\mathrm{DFT}(\mathbf{h})$. $\mathbf{h}$ is the channel impulse response whose maximum delay is assumed to be shorter than the length of CP, making the signal unaffected by the inter-symbol interference (ISI).

The received data on the odd subcarriers is $\mathbf{R}_{odd} = \mathbf{G}_{odd}\mathbf{R}$, where $\mathbf{G}_{odd}$ is a $P \times N$ matrix selecting the first half odd subcarriers of $\mathbf{X}$ as following

$$\mathbf{G}_{odd,(m,n)} = \begin{cases} 1, & n = 2m+1, 0 \le m \le P-1 \\ 0, & \text{otherwise} \end{cases} \tag{8}$$

Then, the received data vector for the $k$th user is

$$\mathbf{R}^{(k)} = \mathbf{D}^{(k)\mathrm{T}} \mathbf{R}_{odd} \tag{9}$$

From previous description, the $M \times M$ CFR diagonal matrix $\tilde{\mathbf{H}}^{(k)}$ for the $k$th user can be expressed as

$$\tilde{\mathbf{H}}^{(k)} = \mathbf{D}^{(k)\mathrm{T}} \mathbf{G}_{odd} \mathbf{H} \mathbf{G}_{odd}^{\mathrm{T}} \mathbf{D}^{(k)} \tag{10}$$

therefore, $R_m^{(k)} = \tilde{H}_m^{(k)} S_m^{(k)} + V_m^{(k)}$, $m = 0, 1, ...M-1$.

Note that the first half of the even subcarriers of $\mathbf{X}$ contains the same information as the last half, so the information of the received signal on the even subcarriers is $\mathbf{R}_{even} = \mathbf{G}_{even}\mathbf{R}$, where $\mathbf{G}_{even}$ is a $P \times N$ matrix selecting the first half even subcarriers of $\mathbf{X}$ as following

$$\mathbf{G}_{even,(m,n)} = \begin{cases} 1, & n = 2m, 0 \le m \le P-1 \\ 0, & \text{otherwise} \end{cases} \tag{11}$$

The received clipping noise information on the even subcarriers $\mathbf{R}_{even}$ contains the information of the transmitted data which can be used to improve the decoding accuracy [2].

## III. Improved FD-DFE Structure

The improved structure and algorithm are analyzed in this section. [7] proposed a traditional DFE structure in RF with both the feedforward (FF) filter and the feedback (FB) filter. In this letter, the noise elimination method based on [2] is introduced to improve the system performance of the SC-FDMA VLC system. From the description in Section II, the received signal after CP removing and $N$-point DFT is transformed and selected to generate $\mathbf{R}^{(k)}, k=1,2...K$, and $\mathbf{R}_{even}$ in the frequency domain.

Fig. 2 shows the block diagrams of the proposed structure. The proposed equalizer contains two parts, the FF filter and the FB filter. In the FF filter, the received signal $\mathbf{r}$ is first transformed to frequency domain. $\mathbf{R}^{(k)}$ is the received information for the $k$th user and $\mathbf{R}_{even}$ is the information on the even subcarriers. $l$ is the iteration index. The FF filter coefficients $\mathbf{C}^{(k)(l)}$ and the FB vector signal $\mathbf{Y}^{(k)(l)}$ are used to eliminate the interference of the received signal. $\mathbf{u}^{(k)(l)}$ is the estimated data vector for the $k$th user in the $l$th iteration after $M$-point IDFT. $\mathbf{R}_{even}^{(1)}$ is the result of $\mathbf{R}_{even}$ after FD-LE. In the FB filter, after a threshold detector and S/P, the estimated signals $\tilde{d}_m^{(k)(l-1)}, k=0,1,\ldots,K$, at iteration $(l-1)$ are used to generate the FB vectors $\mathbf{Y}^{(k)(l)}$ with FB filter coefficients $\mathbf{B}^{(k)(l)}$. The decision error elimination (DEE) module is used to reduce the decision error of $\hat{\mathbf{S}}^{(k)(l-1)}$ in which the information from the even subcarriers $\mathbf{R}_{even}^{(1)}$ is involved.

As illustrated in Fig. 2, for the $k$th user, the $m$th component of the feedback vector $\mathbf{Y}^{(k)(l)}$ in the FF filter is

$$Y_m^{(k)(l)} = B_m^{(k)(l)} \hat{S}_m^{(k)(l-1)}, \qquad m=0,1,...M-1. \tag{12}$$

The detected data block $\hat{\mathbf{s}}^{(k)(l-1)}$, with the iteration index $(l-1)$, is transformed to frequency domain by $M$-point DFT which yields $\hat{\mathbf{S}}^{(k)(l-1)}$. Then, the output vector $\mathbf{u}^{(k)(l)}$ of the equalizer is as follows

$$\mathbf{u}^{(k)(l)} = \mathrm{IDFT}(\mathbf{C}^{(k)(l)}\mathbf{R}^{(k)} + \mathbf{Y}^{(k)(l)}) \tag{13}$$

The power of transmitted signal and detected data are

$$\mathrm{M}_{S_m^{(k)}} = \mathrm{E}[|S_m^{(k)}|^2], \qquad \mathrm{M}_{\hat{S}_m^{(k)(l)}} = \mathrm{E}[|\hat{S}_m^{(k)(l)}|^2] \tag{14}$$

The correlation between these two signals is

$$\mathrm{r}_{S_m^{(k)},\hat{S}_m^{(k)(l-1)}} = \mathrm{E}[S_m^{(k)} \hat{S}_m^{(k)(l-1)*}] \tag{15}$$

The mean square error (MSE) can be written as [7]

$$\begin{aligned} MSE^{(k)(l)} &= \mathrm{E}[|\tilde{d}_n^{(k)(l)} - d_n^{(k)}|^2] \\ &= \frac{\mathrm{M}_W}{M^2}\sum_{m=0}^{M-1}|C_m^{(k)(l)}|^2 \\ &+ |C_m^{(k)(l)}\tilde{H}_m^{(k)} - 1|^2 \mathrm{M}_{S_m^{(k)}} + |B_m^{(k)(l)}|^2 \mathrm{M}_{\hat{S}_m^{(k)(l-1)}} \\ &+ 2\mathrm{Re}[B_m^{(k)(l)*}(C_m^{(k)(l)}\tilde{H}_m^{(k)} - 1)\mathrm{r}_{S_m^{(k)},\hat{S}_m^{(k)(l-1)}}] \end{aligned} \tag{16}$$

where $\mathrm{M}_W = M\sigma_w^2$ is the noise power in the frequency domain. To minimize $MSE^{(k)(l)}$, let

$$\sum_{m=0}^{M-1} B_m^{(k)(l)} = 0 \tag{17}$$

the solution of $B_m^{(k)(l)}$ is

$$B_m^{(k)(l)} = -\frac{\mathrm{r}_{S_m^{(k)},\hat{S}_m^{(k)(l-1)}}}{\mathrm{M}_{\hat{S}_m^{(k)(l-1)}}}[\tilde{H}_m^{(k)} C_m^{(k)(l)} - \gamma^{(k)(l)}] \tag{18}$$

and

$$\gamma^{(k)(l)} = \sum_{m=0}^{M-1} \tilde{H}_m^{(k)} C_m^{(k)(l)} \tag{19}$$

The solution of $C_m^{(k)(l)}$ can be calculated as

$$C_m^{(k)(l)} = \frac{\tilde{H}_m^{(k)*}}{\mathrm{M}_W + \mathrm{M}_{S_m^{(k)}}\left(1 - \frac{|\mathrm{r}_{S_m^{(k)},\hat{S}_m^{(k)(l-1)}}|^2}{\mathrm{M}_{\hat{S}_m^{(k)(l-1)}}\mathrm{M}_{S_m^{(k)}}}\right)|\tilde{H}_m^{(k)}|^2} \tag{20}$$

The parameters $\mathrm{M}_{\hat{S}_m^{(k)(l)}}$ and $\mathrm{r}_{S_m^{(k)},\hat{S}_m^{(k)(l-1)}}$ can be estimated as follows [7]

$$\mathrm{M}_{\hat{S}_m^{(k)(l)}} = \frac{1}{M^2}\sum_{m=0}^{M-1}|\hat{S}_m^{(k)(l)}|^2 \tag{21}$$

$$\mathrm{r}_{S_m^{(k)},\hat{S}_m^{(k)(l-1)}} = \frac{1}{M^2}\sum_{m=0}^{M-1}\frac{R_m^{(k)}}{\tilde{H}_m^{(k)}}\hat{S}_m^{(k)*} \tag{22}$$

The performance of the FD-DFE strongly depends on the decision error of $\hat{S}_m^{(k)(l)}$. The structure proposed in this letter improves the performance of the equalizer by reducing the decision error with the use of clipping noise information contained in $\mathbf{R}_{even}$ [2]. As shown in Fig. 2, the DEE module is adopted to reduce the decision error and provide a better estimation of the detected data.

For all the users, when $l=1$ in the first iteration, there is no detected data block previously. Accordingly, $\hat{d}_m^{(0)}$ should be set to be a zero vector, which means the FD-DFE turns out to be a FD-LE.

$\mathbf{R}_{even}^{(1)}$ is the result of $\mathbf{R}_{even}$ after FD-LE as follows

$$\begin{aligned} R_{even,p}^{(1)} &= C_{even,p}^{(1)} R_{even,p} \\ &= \frac{\tilde{H}_{even,p}^{*} R_{even,p}}{\mathrm{M}_W' + \mathrm{M}_{R_{even,p}}|\tilde{H}_{even,p}|^2}, \quad p=0,1,...,P \end{aligned} \tag{23}$$

where $\mathrm{M}_W' = P\sigma_w^2$ and $\tilde{\mathbf{H}}_{even}$ is the CFR matrix of even subcarriers that can be obtained from $\mathbf{H}$ easily. $\mathrm{M}_{R_{even,p}}$ is the power of the signal $\mathbf{R}_{even}$. The reconstructed data vector $\mathbf{T}'$ at the receiver can be written as

$$\mathbf{T}' = \sum_{k=1}^{K}\mathbf{D}^{(k)}\mathbf{U}^{(k)(1)} \tag{24}$$

where $\mathbf{U}^{(k)(1)}$ is the estimated data vector after FD-LE for the $k$th user of the first iteration in the frequency domain. The reconstructed transmitted signal $\mathbf{X}'$ can be expressed as

$$\mathbf{X}' = [0, T_1', 0, T_2', ..., T_{P-1}', 0, T_{P-1}'^*, 0, ..., T_2'^*, 0, T_1'^*]^{\mathrm{T}} \tag{25}$$

$\mathbf{X}'$ can be used to estimate the sign matrix $\mathbf{S}(\mathbf{X})$ which is a diagonal matrix with the diagonal components of the sign of the transmitted signal $\mathbf{x}$ in the time domain [10]

$$\mathbf{S}(\mathbf{X}) = \mathrm{diag}\{\mathrm{sign}(\mathbf{x})\} = \mathrm{diag}\{\mathrm{sign}(\mathbf{W}_N^{\mathrm{H}}\mathbf{X})\} \tag{26}$$

where, superscript $^{\mathrm{H}}$ denotes the Hermitian transpose, $\mathbf{W}_N$ is a $N \times N$ DFT matrix, i.e. $W_{N,(m,n)} = e^{-j2\pi/N \cdot mn}, m, n = 0, 1, \ldots N-1$.

The clipping noise information on the even subcarriers after time domain asymmetrical clipping is $\mathbf{W}_N|\mathbf{x}|$, and we have

$$|\mathbf{x}| = \mathbf{S}(\mathbf{X})\mathbf{x} = \mathbf{S}(\mathbf{X})\mathbf{W}_N^{\mathrm{H}}\mathbf{X} \tag{27}$$

By using the zero-forcing (ZF) estimator, the improved estimation of the data on the odd subcarriers is [10]

$$\mathbf{R}_{odd,improved}^{(1)} = \mathbf{R}_{odd}^{(1)} + \frac{1}{2}\mathbf{G}_{odd}\mathbf{W}_N\mathbf{S}(\mathbf{X}')\mathbf{W}_N^{\mathrm{H}}\mathbf{R}_{even}^{(1)} \tag{28}$$

where $\mathbf{R}_{odd}^{(1)}$ is the result of $\mathbf{R}_{odd}$ after FD-LE, which can be obtained from $\mathbf{T}'$ easily.

The detected data in the first iteration of the FB filter for the $k$th user can be obtained as

$$\hat{\mathbf{S}}^{(k)(1)} = \mathbf{D}^{(k)\mathrm{T}}\mathbf{R}_{odd,improved}^{(1)} \tag{29}$$

Note that for the clipping noise on the even subcarriers, there is no corresponding data symbols, so the information can only be used in the first iteration. The DEE module can be described as

$$\hat{\mathbf{S}}^{(k)(l)} = \begin{cases} 0, & l = 0 \\ \mathbf{D}^{(k)\mathrm{T}}\mathbf{R}_{odd,improved}^{(1)}, & l = 1 \\ \mathrm{DFT}(\hat{\mathbf{s}}^{(k)(l)}), & l = 2,3,4... \end{cases} \tag{30}$$

The DEE module can eliminate the decision error in the first iteration effectively and provide a more accurate estimation of $\hat{\mathbf{S}}^{(k)(1)}$. Compared with traditional DFE structure, the application of DEE module requires two additional $N$-point IFFT and one $N$-point FFT.

## IV. Simulation Results

In the simulation, it is assumed that the channel estimation and synchronization are ideal. Quadrature phase shift keying (QPSK) constellation is applied. The simulation is based on a coded system with a $1/2$-rate convolutional encoder (133,171) and a soft-decision Viterbi decoder. The total number of subcarriers is $N = 1024$. The length of CP is set to be $N/8 = 128$. Non-line of sight (NLOS) channel model of diffuse reflections (combined impulse response of $k$-bounce pulses, $k \in \{1,2,3\}$) is used in the simulations, representing the worst case of optical channel [11]. The subcarrier FH pattern is randomly generated.

The simulation results are shown in Fig. 3. For different structures, the performance of FD-LE without feedback is the worst among all the structures. Traditional FD-DFE performs much better because of the decision feedback technique. Note that for traditional FD-DFE structure, the FB filter is fed with previously detected data. It is clear that the improved structure in this letter provides a better performance than FD-LE and traditional FD-DFE when FB filter is fed with more accurate detected data utilizing the information of both the odd and even subcarriers. At $\mathrm{BER} = 1.0\mathrm{e}-4$, the performance of the improved FD-DFE is around $2.7$ dB better than FD-LE, and $1.4$ dB better than traditional FD-DFE in the localized and distributed allocation. With the ideal feedback, the ideal FD-DFE provides the best performance. Compared with localized and distributed allocation, FH allocation improves the performance of every structure respectively because of the randomness of subcarrier distribution. At $\mathrm{BER} = 1.0\mathrm{e}-4$, the performance of the improved FD-DFE is about $2.4$ dB better than FD-LE, and $1.0$ dB better than traditional FD-DFE in the FH allocation.

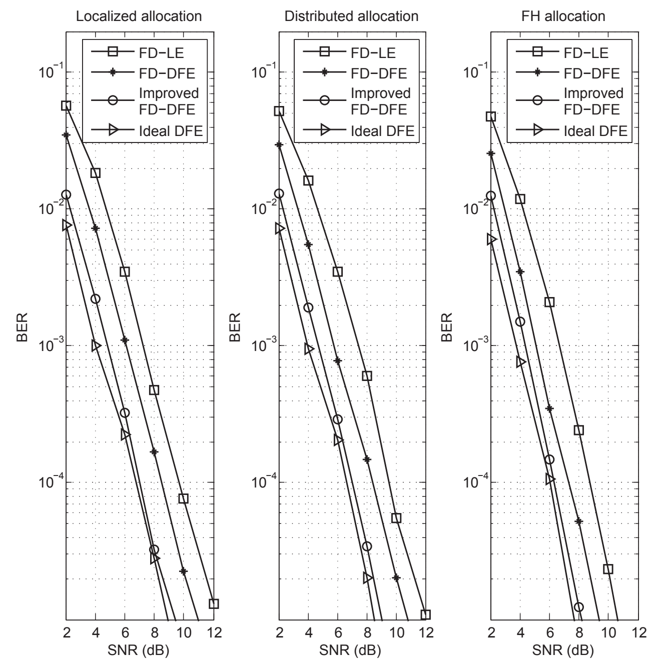

Fig. 3. BER performance against SNR. K=4, M=64, l=2.

## V. Conclusion

This letter proposed an improved FD-DFE structure for downlink VLC systems based on SC-FDMA. Time domain asymmetrical clipping method is applied in the proposed structure. With the use of the clipping noise information on the even subcarriers, the performance of the algorithm is improved with lower feedback decision error, reducing the BER of the system. Different subcarrier distributions are analyzed and simulated. The proposed structure provides a better transmission performance than traditional FD-DFE structures.

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