

Dynamic Range Constrained Clipping in Visible Light OFDM Systems with Brightness Control

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Abstract—Visible light communication (VLC) systems can provide illumination and communication simultaneously by way of light emitting diodes (LEDs). The communication function is achieved by employing simple and low-cost intensity modulation and direct detection (IM/DD) schemes. Brightness control and flicker mitigation are two main challenges for the illumination function. Orthogonal frequency division multiplexing (OFDM) has been considered for VLC for its ability to boost data rates. However, OFDM waveforms have high peak-to-average power ratio, and will be clipped if its magnitude is beyond the dynamic range of LEDs. Clipping can cause the performance degradation of communication as well as illumination. In this paper, we will propose an iterative clipping method considering brightness control and flicker mitigation. We will investigate the performance in terms of error vector magnitude (EVM) as well as computational complexity. We will formulate the EVM minimization problem as a convex optimization problem to compare with the iterative clipping method.

I. INTRODUCTION

Visible light communication (VLC) has drawn extensive attention recently for its potential to complement the conventional RF communication [1], [2]. VLC uses the visible light spectrum to transmit information; it can provide illumination and communication simultaneously by way of light emitting diodes (LEDs). VLC has many advantages including low-cost front-ends, energy-efficient transmission, huge (THz) bandwidth, no electromagnetic interference, and no eye safety constraints like infrared [3]. To achieve the goal of communication, simple and low-cost intensity modulation and direct detection (IM/DD) techniques are employed, thus only signal intensity information, not phase information, is modulated. IM/DD requires the electric signal to be real-valued and unipolar (positive-valued). Brightness control and flicker mitigation are two main challenges for the illumination function in VLC. Generally, there are two ways to control the brightness: (i) adjust the average forward voltage or current; (ii) change the duty cycle of pulse width modulation (PWM). The standard IEEE 802.15.7 [4] has applied the above two ways to control the brightness for on-off keying (OOK) and variable pulse position modulation (VPPM). Flicker refers to the situation that the changing variations of the light intensity is noticeable to human eye. Run length limited (RLL) code is utilized in IEEE 802.15.7 [4] to mitigate flicker for OOK and VPPM.

Orthogonal frequency division multiplexing (OFDM) has been considered for VLC due to its ability to boost data rates

and efficiently combat inter-symbol-interference (ISI) [5], [6], [7]. The authors in [8] have demonstrated a data rate of 1 Gb/s at a standard illumination level by using OFDM. To ensure that the OFDM time-domain signal is real-valued, Hermitian symmetry must be satisfied in the frequency-domain. However, OFDM is known for its disadvantage of high peak-to-average power ratio (PAPR) and thus is very sensitive to nonlinear distortions. The LED is the main source of non-linearity in VLC. Although LEDs can be linearized by a predistorter [9], the dynamic range is limited by the turn-on voltage (TOV) and maximum permissible alternating current. The input signal outside this range will be clipped. Clipping will not only degrade the performance of communication, but also affect the brightness control and cause inter-symbol flicker.

In this paper, we will propose an iterative clipping method considering brightness control and flicker mitigation. We will focus on DC biased optical OFDM (DCO-OFDM) [5]. We will investigate the performance in terms of error vector magnitude (EVM) as well as computational complexity. We will formulate the EVM minimization problem as a convex optimization problem to compare with the iterative clipping method.

II. DYNAMIC RANGE CONSTRAINED VISIBLE LIGHT OFDM SYSTEM

In VLC, an LED is utilized to simultaneously transmit information and provide illumination. The principle is that the human eye cannot perceive fast-changing variations of the light intensity, and only responds to the average light intensity. To implement the communication function, the simple and low-cost intensity modulation (IM) and direct detection (DD) schemes are employed. At the transmitter, the forward signal $y[n]$ drives the LED which in turn converts the magnitude of the input electric signals $y[n]$ into optical intensity. At the receiver, a photodiode (PD) transforms received optical power into the amplitude of an electrical signal.

LEDs are the main source of non-linearity in VLC. With predistortion, the input-output characteristic of the LED can be linearized, but only within a limited interval $[V_{\text{tov}}, V_{\text{sat}}]$, where V_{tov} denotes the turn on voltage and V_{sat} denotes the saturation input voltage [9]. The Dynamic range can be denoted by $D \triangleq V_{\text{sat}} - V_{\text{tov}}$. Fig. 1 shows the input-output characteristic of an ideal LED. O_{sat} denotes the output optical power corresponding to the input voltage V_{sat} . The input signal

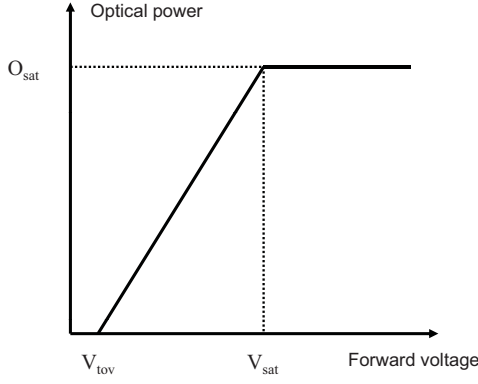


Fig. 1. Ideal linear LED characteristic.

$y[n]$ will be clipped if its magnitude is beyond the dynamic range of LED. The clipped signal $\bar{y}[n]$ is given by

$$\bar{y}[n] = \begin{cases} V_{\text{sat}}, & y[n] > V_{\text{sat}} \\ y[n], & V_{\text{tov}} \leq y[n] \leq V_{\text{sat}} \\ V_{\text{tov}}, & y[n] < V_{\text{tov}} \end{cases} \quad (1)$$

In an OFDM system, a discrete time-domain symbol $\mathbf{x} = [x[0], x[1], \dots, x[N-1]]$ is generated by applying the inverse FFT (IFFT) operation to a frequency-domain symbol $\mathbf{X} = [X_0, X_1, \dots, X_{N-1}]$ as

$$x[n] = \text{IDFT}(X_k) = \frac{1}{\sqrt{N}} \sum_{k=0}^{N-1} X_k \exp\left(\frac{j2\pi kn}{N}\right), \quad n = 0, 1, \dots, N-1, \quad (2)$$

where $j = \sqrt{-1}$ and N are the size of IFFT. In VLC, IM/DD schemes require the baseband signal in VLC to be real-valued. To generate real-valued baseband OFDM signal, DC biased optical OFDM (DCO-OFDM) [5] was introduced for VLC. According to the property of inverse Fourier transform, a real-valued time-domain signal $x[n]$ corresponds to a frequency-domain signal X_k that is Hermitian symmetric; i.e.,

$$X_k = X_{N-k}^*, \quad 1 \leq k \leq N-1, \quad (3)$$

where $*$ denotes complex conjugate. In DCO-OFDM, the 0th and $N/2$ th subcarrier are null; i.e., $X_0 = 0$, $X_{N/2} = 0$. The time-domain signal $x[n]$ can be obtained as

$$x[n] = \frac{2}{\sqrt{N}} \sum_{k=1}^{N/2-1} \left(\Re(X_k) \cos\left(\frac{2\pi kn}{N}\right) - \Im(X_k) \sin\left(\frac{2\pi kn}{N}\right) \right), \quad n = 0, 1, \dots, N-1, \quad (4)$$

where $\Re(\cdot)$ denotes the real part of X_k and $\Im(\cdot)$ denotes the imaginary part of X_k . Since the DC component is zero ($X_0 = 0$), $x[t]$ has zero average value. Let us denote by σ_x^2 the variance of $x[n]$. It is well-known that the OFDM time-domain signal has high peak-to-average power ratio (PAPR) [10], which is defined as

$$\text{PAPR} \triangleq \frac{\max x^2[n]}{\sum_{n=0}^{N-1} x^2[n]/N}. \quad (5)$$

To convert $x[n]$ into a unipolar signal, a biasing B is added to $x[n]$ as

$$y[n] = x[n] + B. \quad (6)$$

The resulting signal $y[n]$ will have a average value B . Let us define the power back-off as $\gamma \triangleq D^2/\sigma_x^2$, and biasing ratio as $\varsigma \triangleq (B - V_{\text{tov}})/D$.

Brightness control is essential for the illumination function in VLC. Let O_{des} denote the desired emitted average optical power, which is determined by the illumination scenario, and let V_{des} denote the corresponding input voltage. The principle of brightness control is to make the average amplitude of the input forward signal equal to V_{des} . To avoid flicker of LED, the average value of each N -length symbol $y[n]$ must be kept constant. Since the average value of the biased signal $y[n]$ is equal to B , it is straightforward to set biasing level equal to V_{des} for each DCO-OFDM symbol; i.e., $B = V_{\text{des}}$, which is called the biasing adjustment method. We can also control the brightness via PWM. A PWM signal with period N_{pwm} is expressed as

$$p[n] = \begin{cases} V_{\text{pwm}}, & 0 \leq n < N \\ 0, & N \leq n < N_{\text{pwm}} \end{cases}, \quad (7)$$

where N is the “on” duration and $N_{\text{pwm}} - N$ is the “off” duration. V_{pwm} denotes the input forward voltage in the “on” duration. The output optical power can be adjusted by changing the duty cycle defined as $d \triangleq N/N_{\text{pwm}}$. To generate the optical intensity with average power O_{des} , the duty cycle of PWM has to be chosen as

$$d = \frac{V_{\text{des}} - V_{\text{tov}}}{V_{\text{pwm}} - V_{\text{tov}}}, \quad (8)$$

where $d \leq 1$, and $V_{\text{pwm}} \geq V_{\text{des}}$. We propose to combine the DCO-OFDM signal with PWM as

$$y[n] = \begin{cases} x[n] + V_{\text{pwm}}, & 0 \leq n < N \\ 0, & N \leq n < N_{\text{pwm}} \end{cases}, \quad (9)$$

which can be seen as a DCO-OFDM symbol with biasing level $B = V_{\text{pwm}}$ followed by $N_{\text{pwm}} - N$ length “0” compensations.

Since $y[n]$ is constrained by the dynamic range $[V_{\text{tov}}, V_{\text{sat}}]$, the constraints will in turn apply to $x[n]$ as $[V_{\text{tov}} - B, V_{\text{sat}} - B]$. Therefore, the maximum $x[n]$ is limited by $V_{\text{sat}} - B$, and the minimum $x[n]$ is limited by $V_{\text{tov}} - B$. The clipped signal $\bar{y}[n]$ in (1) can be written as

$$\bar{y}[n] = \bar{x}[n] + B, \quad (10)$$

where $\bar{x}[n]$ denotes the clipped version of $x[n]$ as

$$\bar{x}[n] = \begin{cases} V_{\text{sat}} - B, & x[n] > V_{\text{sat}} - B \\ x[n], & V_{\text{tov}} - B \leq x[n] \leq V_{\text{sat}} - B \\ V_{\text{tov}} - B, & x[n] < V_{\text{tov}} - B \end{cases} \quad (11)$$

Let us define the upper peak-to-average power ratio (UPA-PR) of $x[n]$ as

$$\mathcal{U} \triangleq \frac{(\max x[n])^2}{\sum_{n=0}^{N-1} x^2[n]/N}, \quad (12)$$

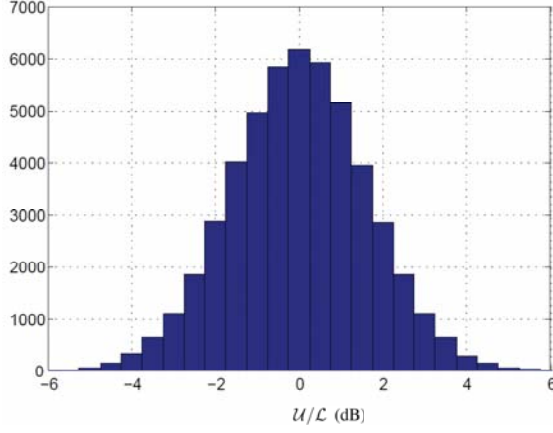


Fig. 2. Histogram of the ratio \mathcal{U}/\mathcal{L} from 50000 DCO-OFDM symbols (QPSK modulation, $N = 256$).

and the lower peak-to-average power ratio (LPAPR) of $x[n]$ as

$$\mathcal{L} \triangleq \frac{(\min x[n])^2}{\sum_{n=0}^{N-1} x^2[n]/N}. \quad (13)$$

It is worth mentioning that PAPR, UPAPR, and LPAPR obey the following relationship:

$$\text{PAPR} = \max \{ \text{UPAPR}, \text{LPAPR} \}. \quad (14)$$

Fig. 2 shows the histogram of the ratio \mathcal{U}/\mathcal{L} , from 50000 DCO-OFDM symbols. We chose QPSK modulation and $N = 256$. In the histogram, the maximum probability occurs around the ratio $\mathcal{U}/\mathcal{L} = 0$ dB since DCO-OFDM are symmetric distributed. We can observe that the ratio \mathcal{U}/\mathcal{L} has a wide range, which means that the difference between UPAPR and LPAPR can be very large for some symbols. Therefore, depending on the biasing ratio, UPAPR, and LPAPR, the clipping effects vary symbol by symbol.

After clipping, the average value of the clipped signal $\bar{y}[n]$ will become

$$\frac{1}{N} \sum_{n=0}^{N-1} \bar{y}[n] = B + \frac{1}{N} \sum_{n=0}^{N-1} \bar{x}[n]. \quad (15)$$

where $\frac{1}{N} \sum_{n=0}^{N-1} \bar{x}[n]$, the average value of $\bar{x}[n]$, will not necessarily equal to zero. For example, when the LED is dimly lit, the biasing level B will be very close to V_{tov} . Then the absolute value of upper clipping level $|V_{\text{sat}} - B|$ will be much greater than the absolute value of lower clipping level $|V_{\text{tov}} - B|$. Because $x[n]$ are symmetric distributed, the $x[n]$ will be clipped at the negative tails more than at the positive tails. In this case, the average value of $\bar{x}[n]$ will be greater than zero. Besides, because the \mathcal{U} and \mathcal{L} are random variables, the average value of $\bar{x}[n]$ will change symbol by symbol. In summary, the clipping can affect the illumination function in two ways. First, the emitted average optical power will deviate the desired value O_{des} because the average value of the clipped symbol $\bar{y}[n]$ is not equal to B . Second, the clipping may cause

LED to flicker since the average value of $\bar{y}[n]$ changes symbol by symbol.

III. ITERATIVE CLIPPING METHOD

To avoid the deviation of the output optical power and flicker, an iterative clipping method is proposed in the paper. The objective is to generate a forward signal $\tilde{y}[n]$ as

$$\tilde{y}[n] = \tilde{x}[n] + B, \quad (16)$$

where $\tilde{x}[n]$ has a zero average value and a limited dynamic range as

$$\frac{1}{N} \sum_{n=0}^{N-1} \tilde{x}[n] = 0, \quad (17)$$

$$V_{\text{tov}} - B \leq \tilde{x}[n] \leq V_{\text{sat}} - B. \quad (18)$$

Let $\tilde{x}^{(i-1)}[n]$ denote the output signal from the $(i-1)$ th iteration, which has a zero average value. In the i th iteration, we first compare $\max \tilde{x}^{(i-1)}[n]$ with $V_{\text{sat}} - B$, and compare $\min \tilde{x}^{(i-1)}[n]$ with $V_{\text{tov}} - B$.

If $\max \tilde{x}^{(i-1)}[n] > V_{\text{sat}} - B$ and $\min \tilde{x}^{(i-1)}[n] < V_{\text{tov}} - B$, we operate upper clipping and lower clipping separately as

$$\bar{x}^u[n] = \begin{cases} V_{\text{sat}} - B, & \tilde{x}^{(i-1)}[n] > V_{\text{sat}} - B \\ \tilde{x}^{(i-1)}[n], & \tilde{x}^{(i-1)}[n] \leq V_{\text{sat}} - B \end{cases} \quad (19)$$

$$\bar{x}^l[n] = \begin{cases} V_{\text{tov}} - B, & \tilde{x}^{(i-1)}[n] < V_{\text{tov}} - B \\ \tilde{x}^{(i-1)}[n], & \tilde{x}^{(i-1)}[n] \geq V_{\text{tov}} - B \end{cases} \quad (20)$$

Then we take FFT of both $\bar{x}^u[n]$ and $\bar{x}^l[n]$ to obtain the frequency-domain signal \bar{X}_k^u and \bar{X}_k^l , respectively. We compare the distortions power $\bar{P}^u = \sum_{k=1}^{N/2-1} |\bar{X}_k^u - X_k|^2$ and $\bar{P}^l = \sum_{k=1}^{N/2-1} |\bar{X}_k^l - X_k|^2$, and obtain $\tilde{x}^{(i)}[n]$ as

$$\tilde{x}^{(i)}[n] = \begin{cases} \bar{x}^u[n], & \bar{P}^u < \bar{P}^l \\ \bar{x}^l[n], & \text{otherwise} \end{cases} \quad (21)$$

If $\max \tilde{x}^{(i-1)}[n] > V_{\text{sat}} - B$, and $\min \tilde{x}^{(i-1)}[n] \geq V_{\text{tov}} - B$, we only operate upper clipping and obtain $\tilde{x}^{(i)}[n]$ as $\tilde{x}^{(i)}[n] = \bar{x}^u[n]$.

If $\max \tilde{x}^{(i-1)}[n] \leq V_{\text{sat}} - B$, and $\min \tilde{x}^{(i-1)}[n] < V_{\text{tov}} - B$, we only operate lower clipping and obtain $\tilde{x}^{(i)}[n]$ as $\tilde{x}^{(i)}[n] = \bar{x}^l[n]$.

Second, we remove the DC component from $\tilde{x}^{(i)}[n]$ as

$$\tilde{x}^{(i)}[n] = \tilde{x}^{(i)}[n] - \frac{1}{N} \sum_{n=0}^{N-1} \tilde{x}^{(i)}[n]. \quad (22)$$

Third, examine whether $\tilde{x}^{(i)}[n]$ satisfies the dynamic range as

$$V_{\text{tov}} - B \leq \tilde{x}^{(i)}[n] \leq V_{\text{sat}} - B. \quad (23)$$

If it does not, go to the next iteration; otherwise, we obtain the desired signal $\hat{x}[n] = \tilde{x}^{(i)}[n]$.

In practice, to facilitate the convergence of the iteration, the upper clipping level $V_{\text{sat}} - B$ in (19) and lower clipping level $V_{\text{tov}} - B$ in (20) are both multiplied by a scaling factor β which is very close to 1.

IV. EVM MINIMIZATION

EVM is a figure-of-merit for distortions. Let $\mathbf{X}^\dagger = [X_0^\dagger, X_1^\dagger, \dots, X_{N-1}^\dagger]$ denote the N -length FFT of the modified time-domain symbol $\mathbf{x}^\dagger = [x^\dagger[0], x^\dagger[1], \dots, x^\dagger[N-1]]$. EVM can be defined as

$$\xi(\mathbf{X}, \mathbf{X}^\dagger) \triangleq \sqrt{\frac{\sum_{k=1}^{N/2-1} |X_k - X_k^\dagger|^2}{\sum_{k=1}^{N/2-1} |X_k|^2}}, \quad (24)$$

Let us consider the setting

$$\hat{x}[n] = x[n] + c[n], \quad 0 \leq n \leq N-1, \quad (25)$$

where $c[n]$ is a distortion signal, and the resulting $\hat{x}[n]$ is expected to be zero average value and have a limited dynamic range as

$$\frac{1}{N} \sum_{n=0}^{N-1} \hat{x}[n] = 0, \quad (26)$$

$$V_{\text{tov}} - B \leq \hat{x}[n] \leq V_{\text{sat}} - B. \quad (27)$$

Clipping can produce one such $\hat{x}[n]$ signal, but there are other less straightforward algorithms that can generate other $\hat{x}[n]$ waveforms that also satisfy (26) and (27). In the frequency-domain,

$$\hat{X}_k = X_k + C_k. \quad (28)$$

Since $x[n]$, $c[n]$, and $\hat{x}[n]$ are all real-valued, X_k , C_k , and \hat{X}_k all should satisfy the Hermitian symmetry condition (3). Therefore, $c[n]$ has the form

$$c[n] = \frac{2}{\sqrt{N}} \sum_{k=1}^{N/2-1} \left(\Re(C_k) \cos\left(\frac{2\pi kn}{N}\right) - \Im(C_k) \sin\left(\frac{2\pi kn}{N}\right) \right) + \frac{1}{\sqrt{N}} C_0 + \frac{1}{\sqrt{N}} C_{N/2} \cos(\pi n). \quad (29)$$

The condition (26) implies $\frac{1}{N} \sum_{n=0}^{N-1} c[n] = 0$. Hence, we have $C_0 = 0$.

We are interested in knowing the lowest possible EVM,

$$\xi = \sqrt{\frac{\sum_{k=1}^{N/2-1} |C_k|^2}{\sum_{k=1}^{N/2-1} |X_k|^2}}, \quad (30)$$

among all such $\hat{x}[n]$ waveforms.

We formulate the following convex optimization problem:

$$\begin{aligned} & \text{minimize} && \sum_{k=1}^{N/2-1} |C_k|^2 \\ & \text{subject to} && \hat{x}[n] \leq V_{\text{sat}} - B \\ & && \hat{x}[n] \geq V_{\text{tov}} - B \\ & && \hat{x}[n] = x[n] + \frac{2}{\sqrt{N}} \sum_{k=1}^{N/2-1} \left(\Re(C_k) \cos\left(\frac{2\pi kn}{N}\right) - \Im(C_k) \sin\left(\frac{2\pi kn}{N}\right) \right) + \frac{1}{\sqrt{N}} C_{N/2} \cos(\pi n), \\ & && 0 \leq n \leq N-1 \\ & && C_{N/2} \in \mathbb{R} \end{aligned} \quad (31)$$

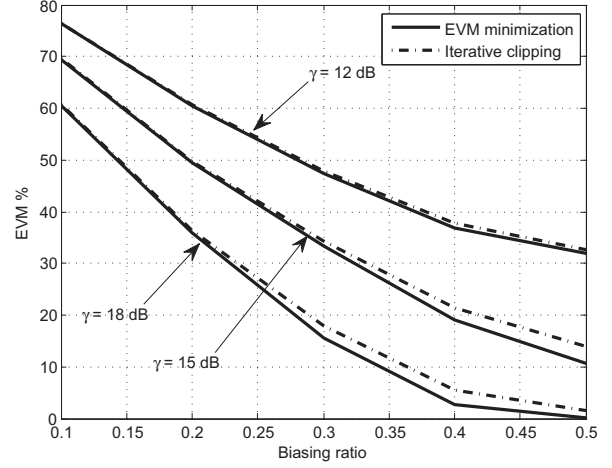


Fig. 3. EVM comparison between iterative clipping method and EVM minimization scheme (QPSK modulation, $N = 256$).

When the distortion of each OFDM symbol is minimized by the above convex optimization approach, the corresponding EVM of $\hat{x}[n]$ (which is proportional to $\sqrt{\sum_{k=1}^{N/2-1} |C_k|^2}$) serves as the lower bound for the given dynamic range. The convex optimization problem can be solved by the interior point method (IPM) as in [11], [12], [13]. In this paper we used CVX, a package for specifying and solving convex programs [14], [15]. Fig. 3 compared the EVM between iterative clipping method and EVM minimization scheme with various power back-off and biasing ratios. In the simulation, we generated 1000 DCO-OFDM symbols with QPSK modulation and $N = 256$. The ζ is assumed between 0 and 0.5. We can observe that the EVM has a minimum value at biasing ratio $\zeta = 0.5$ regardless of the power back-off. When $\zeta = 0.5$, we infer that $V_{\text{sat}} - B = -(V_{\text{tov}} - B)$, i.e., when the $x[n]$ waveform is symmetrically clipped at the negative and positive tails, the clipping error power is always less than that when the two tails are asymmetrically clipped. The proposed iterative clipping method has a no more than 4% gap compared with EVM minimization scheme. Fig. 3 can serve as a reference for choosing an appropriate biasing ratio and power back-off pair given the EVM threshold. For example, if the EVM threshold is 20%, for a power back-off $\gamma = 15$ dB, the biasing ratio has to be chosen greater than 0.4 with iterative clipping method or 0.4 with EVM minimization scheme. However, the biasing adjustment method does not have the freedom to choose the biasing ratio since it is determined by the illumination requirement. The PWM method can increase or decrease the biasing ratio in the “on” interval while keeping the brightness unchanged by decreasing or increasing the duty cycle.

V. COMPLEXITY ANALYSIS

In this section, we will investigate the complexity of the proposed iterative clipping algorithm. The computational complexity will be quantified by the number of instructions per OFDM symbol and iteration.

At the first step, $2N$ comparisons were made to decide which samples need to be upper clipped or lower clipped. Assume that the number of clipped sample is K_t , there are K_t assignment operations. Assuming either comparison operation or assignment operation only requires a single instruction, the number of instructions required to implement clipping is $I_{\text{clipping}} = 2N + K_t$. If $\max \tilde{x}^{(i-1)}[n] > V_{\text{sat}} - B$ and $\min \tilde{x}^{(i-1)}[n] < V_{\text{tov}} - B$, the FFT has be operated. It is known that a N -point FFT requires $N \log N$ multiplication and $N \log N$ additions. In order to calculate \bar{P}_u and \bar{P}_l , we need $4(N/2 - 1)$ subtractions, $4(N/2 - 1)$ multiplications, and $4(N/2 - 1)$ additions. Assuming that the multiplication operation requires α_M instructions and addition/subtraction operation requires α_A instruction, the total number of additional instructions for the case “ $\max \tilde{x}^{(i-1)}[n] > V_{\text{sat}} - B$ and $\min \tilde{x}^{(i-1)}[n] < V_{\text{tov}} - B$ ” is $I_p = \alpha_M(N \log N + 2N - 4) + \alpha_A(N \log N + 4N - 8)$. Actually, the case “ $\max \tilde{x}^{(i-1)}[n] > V_{\text{sat}} - B$ and $\min \tilde{x}^{(i-1)}[n] < V_{\text{tov}} - B$ ” can only happen in the first iteration. The probability depends on the power back-off and biasing ratio.

At the second step, N additions and one multiplication are required to calculate the DC component and N subtractions required to remove the DC-component. Thus, the number of instructions required for the second step is $I_{\text{dc}} = 2\alpha_A N + \alpha_M$.

At the third step, we need $2N$ comparisons to examine whether the dynamic range is obeyed. Hence, the number of instructions required for the third step is $I_{\text{comp}} = 2N$.

In summary, the overall computational complexity of the iterative clipping algorithm is $I_{\text{total}} = I_{\text{clipping}} + I_p + I_{\text{dc}} + I_{\text{comp}}$, where $I_p = 0$ after the first iteration. Either the parameter K_t or the number of iterations depends on the power back-off and biasing ratio. Table I shows the average number of iterations of 1000 OFDM symbols ($N = 256$ subcarriers, QPSK modulation, and $\beta = 0.998$) with various power back-off and biasing ratio.

When solving (31) with IPM, the computation is dominated by one FFT and inversion of a $N \times N$ matrix per iteration. Therefore, the complexity has the order $O(N \log N + N^3)$, which is much more complicated than the iterative clipping method, which has a complexity of order $O(N \log N + N)$.

TABLE I
AVERAGE NUMBER OF ITERATIONS

Biasing ratio	0.1	0.2	0.3	0.4	0.5
Power back-off					
12 dB	9.29	6.01	6.64	9.42	11.49
15 dB	6.55	3.00	2.06	1.97	2.28
18 dB	4.82	2.00	1.11	0.19	0.02

VI. CONCLUSION

In this paper, we investigated the clipping effects on illumination in visible light OFDM systems. To avoid brightness deviation and flicker of simple clipping, we proposed an iterative clipping method. We formulated the EVM minimization problem as a convex optimization problem, which requires high-complexity computation to solve. The simulation

shows that, compared with the EVM minimization method, the iterative clipping method has a less than 4% EVM gap.

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