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Frequency Domain Equalization and Post Distortion for LED Communications with Orthogonal Polynomial Based Joint LED Nonlinearity and Channel Estimation

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Abstract

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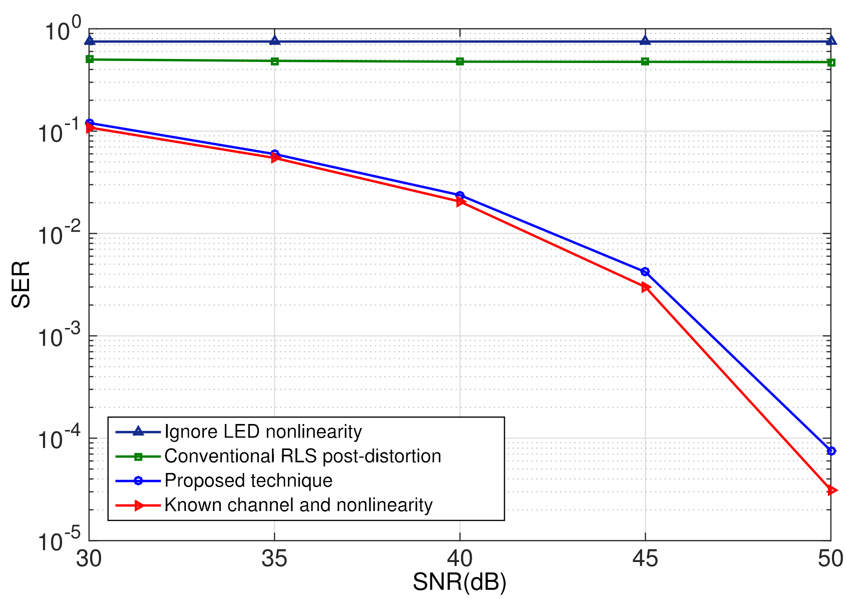

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Frequency Domain Equalization and Post Distortion for LED Communications With Orthogonal Polynomial Based Joint LED Nonlinearity and Channel Estimation

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Abstract: The light-emitting diode (LED) is the major source of nonlinearity in LED communications, and the nonlinearity needs to be effectively modeled and mitigated to avoid the degradation of communication performance. In indoor LED communications, multipath dispersion can lead to inter-symbol interference (ISI) at high data rates. In this paper, we jointly estimate the LED nonlinearity and ISI channel based on an orthogonal polynomial based technique. Then, frequency domain equalization and post-distortion are adopted to mitigate the ISI and nonlinearity. Simulation results are provided to demonstrate the effectiveness of the proposed technique.

Index Terms: Light-emitting diode (LED) communications, nonlinearity, memory polynomial, orthogonal polynomial, ISI channel, frequency domain equalization (FDE).

1. Introduction

Recently, with the rapid development of light-emitting diode (LED) technology, traditional incandescent and fluorescent lights have been gradually replaced by eco-friendly LEDs due to its low power dissipation, long operating life and small size [1]. Another outstanding advantage of LEDs is that they have very high response sensitivity [2], which enables simultaneous illumination and data transmission. LED communications can be achieved based on existing LED lighting facilities at low cost as there is no need for massive infrastructure construction [3]. Also, visible light has unregulated spectrum region and thereby can avoid interferences with current radio frequency (RF) communications [4]. For safety transmission, LED communications has nature merits over traditional RF communications as visible light cannot pass through walls. Hence, LED communication is a research focus, which provides a new option for high-speed indoor communications.

In LED communications, electric signals are modulated to optical ones using the light-intensity of LEDs, and the light intensity is transformed to electric signals by photo detector (PD) [5]. However, the nonlinearity of LED is a challenge in LED communications as the electric to optical conversion is a nonlinear process, which distorts the amplitude of the transmitted signals [6]–[8]. Also, the nonlinear behaviour has memory effects due to transport delay and rapid thermal effects, which cannot be neglected when transmission bandwidth is large [9]. Additionally, in indoor communication environment, optical detectors may receive signals propagated from emitters through multiple paths due to reflections of room surfaces or other objects [10]. The multipath propagation includes two types: line of sight (LOS) propagation which depends on the distance and orientation between emitter and receiver, and non-LOS which is also known as diffuse link due to reflections off the room surroundings. Such multipath dispersion can lead to inter-symbol interference (ISI) at high data rates [11].

Therefore, LED Nonlinearity modelling and mitigation are vital for LED communications to avoid system performance degradation. In [12], based on the least square (LS) method, the author constructed a post-distorter using memory polynomial. However, matrix inversion operation is required in finding the coefficients of the post-distorter, which often generates numerical instability problem [13]–[15]. This leads to inaccurate estimation of the post-distorter and thus the performance of nonlinearity mitigation is severely affected. Recently, we proposed an orthogonal polynomial-based LED nonlinearity modelling technique [16]. We built a set of orthogonal polynomial basis for PAM modulated LED signals to accomplish effective LED nonlinearity modelling. Due to the use of orthogonal polynomials, the basis matrix for our new model are quasi-orthogonal (leading to low complexity inversion of the corresponding correlation matrix), so the problem of numerical instability in doing the matrix inverse operation is solved, resulting in accurate LED nonlinearity modelling and significantly better performance [16]. In the above works, only LOS channels are considered; whilst the ISI effect in optical wireless communications is ignored. In [11], OFDM is used for the mitigation of ISI. However, as OFDM signals has inherent high peak-to-average-power (PAPR) ratio [5], the LED nonlinearity is a more serious issue. In [17], single carrier modulation with frequency domain equalization (FDE) is used to combat ISI in LED communications. However, the nonlinear distortion in LEDs is not considered.

In this paper, we will first extend our previous work on orthogonal polynomial based LED modelling [16] and demonstrate another advantage of this approach, i.e., it allows the simultaneous estimation of the LED nonlinearity and ISI channel coefficients. According to [18] and [19], the bandwidth limitation of LED induced fading will also affect the system performance. In this work, we consider both the LED nonlinearity and LED memory effect. We develop a joint estimation approach for LED nonlinearity and composite ISI channel. The orthogonal polynomial based joint estimation technique circumvents the numerical instability problem over the traditional one and it has low complexity thanks to the orthogonal property of the basis matrix. It is shown in [20] that single carrier LED communication systems can deliver better performance than OFDM LED systems, but conventional single carrier systems require higher computational complexity compared to OFDM ones. It is worth emphasizing that, single carrier system with FDE has similar complexity as OFDM one. Hence, in this work, the ISI effect due to the LED and channel memory is eliminated through frequency domain equalization. Moreover, a post-distorter is employed to mitigate memoryless LED nonlinearity. To the best of our knowledge, for the first time, both LED nonlinearity with memory effect and ISI optical channel are jointly estimated and mitigated. Our work shows that the proposed technique can effectively mitigate the nonlinearity and ISI distortion and significant system performance improvement can be achieved.

The organization of the paper is as follows. In Section 2, the orthogonal polynomial technique in our previous work is briefly introduced. In Section 3, optical wireless channel model is given, and the receiver design for joint LED nonlinearity and ISI channel estimation and mitigation is detailed. Simulation results are provided in Section 4 to verify the effectiveness of our proposed technique. The paper is concluded in Section 5.

2. Orthogonal Polynomial Based LED Nonlinearity Modelling

The memory polynomial model proposed by Kim [21] and Ding [22] is very popular for LED nonlinearity modelling and mitigation [23]. However, it suffers from numerical instability in determining the polynomial coefficients, which may degrade the system performance severely [13], [14]. We have addressed this issue in single carrier LED communication systems by using the orthogonal polynomial technique with PAM modulation in our previous work [16], leading to remarkable performance improvement. As this work is carried out based on the orthogonal polynomial technique, we briefly introduce it in this Section.

The memory polynomial model for LED nonlinearity can be represented as [20]

$$z(n) = \sum_{k=1}^{K} \sum_{m=0}^{M} a_{k,m} x(n-m)|x(n-m)|^{k-1}, \tag{1}$$

where $x(n) \in \chi$ is the PAM modulated input signal, K is the polynomials order, M is the memory length, and $\{a_{k,m}\}$ are two dimensional coefficients that stand for both the nonlinear and memory effects. In LED communications, the transmitted signals $x(n) \geq 0$, we have

$$z(n) = \sum_{k=1}^{K} \sum_{m=0}^{M} a_{k,m} x(n-m)^{k}, \tag{2}$$

which can be expressed as the following matrix form

$$\mathbf{z} = \Phi \mathbf{a}, \tag{3}$$

where $\mathbf{z} = [z(M), z(M+1), \ldots z(N)]^T$, $\mathbf{a} = [a_{1,0}, \ldots a_{K,0}, \ldots a_{1,M}, \ldots a_{K,M}]^T$, and $\Phi = [\Phi_0^1, \ldots, \Phi_0^K, \ldots, \Phi_m^1, \ldots, \Phi_m^K, \ldots, \Phi_M^1, \ldots, \Phi_M^K]$ with $\Phi_m^k = [x(M-m)^k, x(M+1-m)^k, \ldots, x(N-m)^k]^T$. According to the above memory polynomial model in (3), with training signals $\{x(n)\}$, the coefficient $\{a_{k,m}\}$ can be estimated using the LS approach with the following model

$$\mathbf{y} = \Phi \mathbf{a} + \mathbf{n}, \tag{4}$$

where $\mathbf{y} = [y(M), y(M+1), \ldots y(N)]^T$, and $\mathbf{n}$ denotes nonlinearity modelling error and measurement noise. The LS estimate of the parameters $\{a_{k,m}\}$ can be represented as [13]

$$\mathbf{a} = (\Phi^H \Phi)^{-1} \Phi^H \mathbf{y}. \tag{5}$$

However, it is known that $\Phi^H \Phi$ in (5) is usually ill-conditioned, which leads to numerical instability and noise enhancement in LS estimation [13]–[15].

To solve the numerical instability problem in the above conventional memory polynomial techniques, in [16], we proposed an orthogonal polynomial based technique for LED nonlinearity modelling. Instead of using (4) directly, we use

$$\mathbf{y} = \Psi \mathbf{b} + \mathbf{n}, \tag{6}$$

where the new basis matrix Ψ is shown as $\Psi = [\Psi_0^1, \ldots, \Psi_0^K, \ldots, \Psi_m^1, \ldots, \Psi_m^K, \ldots, \Psi_M^1, \ldots, \Psi_M^K]$ where $\Psi_m^k = [\Psi^k(x(M-m)), \Psi^k(x(M+1-m)), \ldots, \Psi^k(x(N-m))]^T$ and the polynomial $\Psi^k(x)$ is defined as $\Psi^k(x) = d_{k,k}x^k + d_{k,k-1}x^{k-1} + \cdots + d_{k,1}x$. As shown in [16], the first order columns of Ψ in the memory part $\Psi_M^1 (m \neq 0)$ are removed, forming a new matrix $\Psi^{'}$, so that the columns of $\Psi^{'}$ can be quasi-orthogonal, by properly choosing the values of the polynomial coefficients $\{d_{\{k,l\}}\}$. This leads to

$$\mathbf{y} = \Psi' \mathbf{b}' + \mathbf{n}. \tag{7}$$

The LS solution to (7) is given by

$$\mathbf{b} = (\Psi'^H \Psi')^{-1} \Psi'^H \mathbf{y}. \tag{8}$$

Although the above solution still involves a matrix inverse operation, the numerical instability problem is avoided due to the quasi-orthogonality of $\Psi^{'}$.

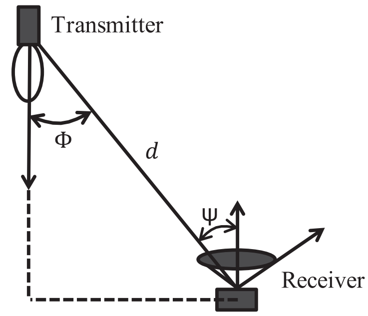

Fig. 1. LOS propagation.

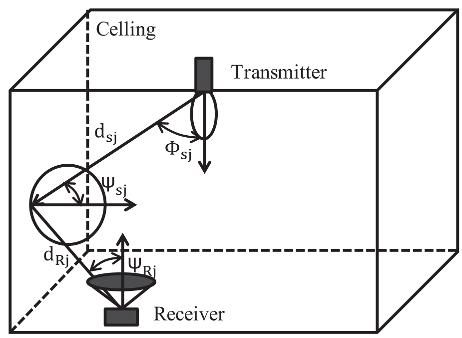

Fig. 2. NLOS propagation.

3. Equalization and Post Distortion With Orthogonal Polynomial Based Joint Nonlinearity and Channel Estimation

3.1 ISI Channel Model for Indoor LED Communications

In indoor LED communications, optical detectors may receive signals propagated from emitters through multiple paths due to reflections off the room surfaces or any other objects [10]. The optical wireless channel transfer function is given by [24]

$$h(t) = h_{los}(t) + h_{nlos}(t), \tag{9}$$

where $h_{los}(t)$ is the contribution due to LOS, which is independent of the modulation frequency but depends on the distance and orientation between emitter and receiver, and $h_{nlos}(t)$ represents non-LOS propagation, which is known as diffuse link due to reflections off the room surroundings. As shown in Fig. 1, where we consider a receiver located at a distance of d and angle Φ with respect to emitter, the impulse response of LOS link can be expressed as [25]–[27]

$$h_{los}(t) = \frac{A_r}{d^2} R_0(\Phi) T_s(\Psi) g(\Psi) \cos(\Psi) \delta(t - d/c), \tag{10}$$

where A_r is the active area of the photo-diode, $T_s(\Psi)$ is the optical band-pass filter, $g(\Psi)$ is the gain of non-imaging concentrator, c is the speed of the light in free space, δ is the Dirac function, and $R_0(\Phi)$ denotes the Lambertian radiant intensity.

For the non-LOS model, the surfaces of the room are divided by R reflecting elements with area of ΔA. As shown in Fig. 2, given a particular single source S and receiver, the channel response after one reflection can be approximated as [25]–[27]

$$h_{nlos}(t) = \sum_{j=1}^{R} \frac{\rho_j A_r \Delta A}{d_{sj}^2 d_{Rj}^2} R_0(\Phi_{sj}) \cos(\Psi_{sj}) \cos(\Psi_{Rj}) \delta\left(t - \frac{d_{sj} + d_{Rj}}{c}\right), \tag{11}$$

where ρ_j is the reflection coefficient of j.

3.2 Receiver Design

3.2.1) Joint LED Nonlinearity and ISI Channel Estimation: In this work, we employ the Hammerstein model for the nonlinearity and memory effect of the LED, it can be represented as [28]

$$z(n) = \sum_{m=1}^{M} b_m g(x(n-m)), \tag{12}$$

where $g(\cdot)$ is a memoryless polynomial and

$$g(x) = \sum_{k=1}^{K} a_k x^k. \tag{13}$$

After transmitting through the optical wireless channel, the received signal can be represented as

$$\mathbf{y} = \mathbf{z} * \mathbf{h} + \mathbf{n}, \tag{14}$$

where symbol $*$ denotes linear convolution, $\mathbf{z} = [z(M), z(M+1), \ldots z(N)]^T$ is the transmitted signal vector, $\mathbf{h} = [h_0, h_1, \ldots, h_{L-1}]^T$ is a length L vector representing the unknown optical channel taps, which can be obtained by sampling $g_{tx}(t)*h(t)*g_{rx}(t)$ with $g_{tx}(t)$ and $g_{rx}(t)$ being the transmit and receive filters, $\mathbf{y} = [y(M), y(M+1), \ldots y(N+L-1)]^T$ is the received signal vector, and $\mathbf{n} = [n(M), n(M+1), \ldots n(N+L-1)]^T$ denotes the additive white Gaussian noise (AWGN) with variance N_0.

It can be seen that the optical channel $\mathbf{h}$ can be absorbed into the Hammerstein model, so that $\mathbf{z} * \mathbf{h}$ in (14) can be written as

$$z'(n) = \sum_{k=1}^{K} \sum_{i=0}^{L+M-1} a_k c_i x(n-i)^k, \tag{15}$$

where c_i is the ith element of the length-(L+M-1) vector $\mathbf{c} = \mathbf{h} * \mathbf{b}$ and $\mathbf{c}$ is called the composite channel vector. It is clear that (15) is a special case of the memory polynomial model with $\{a_k\}$ representing the memoryless nonlinearity coefficients and $\{c_i\}$ representing the composite ISI channel coefficients, respectively. However, due to the removal of the first order columns of the memory parts in Ψ, we cannot obtain $\{a_k\}$ and $\{c_i\}$ directly by using our orthogonal polynomial technique. In the following, we show how to estimate the LED nonlinearity and composite ISI channel simultaneously. The technique was developed based on our previous work on orthogonal method. But it cannot be employed straightforwardly and some tricks are explained in detail. According to (7), the orthogonal polynomial model with basis matrix Ψ' can be expressed as

$$y_o(n) = \sum_{k=1}^{K} b_{k,0} \Psi^k(x(n)) + \sum_{k=2}^{K} \sum_{i=1}^{L+M-1} b_{k,i} \Psi^k(x(n-i)). \tag{16}$$

With training signals at the receiver side, we can get LS estimates of $\{b_{k,i}\}$, which are two dimensional parameters that represent both the nonlinearity and composite ISI channel. A difficult is that the composite channel coefficients and the nonlinearity coefficients are tangled together in the form of multiplication. Next, we show how to seperate $\{c_i\}$ with $\{a_k\}$ from $\{b_{k,i}\}$. In [16], we have proved that (16) and (15) have the following relationship

$$\sum_{k=1}^{K} a_k c_0 x(n)^k - \sum_{k=1}^{K} b_{k,0} \Psi^k(x(n)) \approx A_0, \tag{17}$$

$$\sum_{k=1}^{K} a_k c_i x(n-i)^k - \sum_{k=2}^{K} b_{k,i} \Psi^k(x(n-i)) \approx A_i, i \neq 0, \tag{18}$$

where $\{A_0, \ldots, A_i\}$ are all constants and $A_0 + \cdots + A_i \approx 0$.

In the following, we first calculate c_i and A_0 according to the above relationships based on $\{b_{k,i}\}$. Define

$$D^k = \{x_l^k - x_m^k, 1 \le m < l \le L\}, \tag{19}$$

where L is the size of the PAM alphabet and the size of set D^k is $L \times (L-1)/2$. We use D_j^k to denote the jth element in D^k and define

$$\Delta_D = \sum_{j=1}^{\frac{L\times(L-1)}{2}} \sum_{k=1}^{K} a_k D_j^k. \tag{20}$$

Similarly, define

$$E^k = \{\Psi^k(x_l) - \Psi^k(x_m), 1 \le m < l \le L\}, \tag{21}$$

where the size of set E^k is $L \times (L-1)/2$. We use E_j^k to denote the jth element in E^k and define

$$\Delta_{E_0} = \sum_{j=1}^{\frac{L\times(L-1)}{2}} \sum_{k=1}^{K} b_{k,0} E_j^k, \tag{22}$$

and

$$\Delta_{E_i} = \sum_{j=1}^{\frac{L\times(L-1)}{2}} \sum_{k=2}^{K} b_{k,i} E_j^k. \tag{23}$$

It can be seen that Δ_D, Δ_{E_0} and Δ_{E_i} depend on the PAM alphabet and they are constants when the alphabet is fixed.

Substituting different values of $x(n)$ into (17), we have the following for x_l and $x_m \in \chi$ $(x_l \ne x_m)$,

$$\sum_{k=1}^{K} a_k c_0 x_l^k - \sum_{k=1}^{K} a_k c_0 x_m^k \approx \sum_{k=1}^{K} b_{k,0}\Psi^k(x_l) - \sum_{k=1}^{K} b_{k,0}\Psi^k(x_m). \tag{24}$$

So we can have $L \times (L-1)/2$ equations for different combinations of x_l and x_m. Then, summing each side of these equations, leads to

$$c_0 \Delta_D = \Delta_{E_0}. \tag{25}$$

Similarly, based on (18), we have

$$c_i \Delta_D = \Delta_{E_i}. \tag{26}$$

From (25) and (26), we have the proportional relationship of the composite ISI channel coeffcients

$$\frac{c_0}{c_i} = \frac{\Delta_{E_0}}{\Delta_{E_i}}, \forall i. \tag{27}$$

Also, adding equation (17) and (18), leads to

$$\frac{c_0}{c_i} = \frac{E_0 + LA_0}{E_i + LA_i}, \forall i, \tag{28}$$

where $E_0 = \sum_{k=1}^{K}\sum_{l=1}^{L} b_{k,0}\Psi^k(x_l)$ and $E_i = \sum_{k=2}^{K}\sum_{l=1}^{L} b_{k,i}\Psi^k(x_l)$, which are all known. From (27) and (28), we have

$$\frac{\Delta_{E_0}}{\Delta_{E_i}} = \frac{E_0 + LA_0}{E_i + LA_i}, i = 1, 2 \ldots, I. \tag{29}$$

With $A_0 + \cdots + A_i \approx 0$, A_0 can be determined. In addition, we also know c_0/c_i based on (27), which will be used by frequency domain equalization at the receiver side. According to (17) the

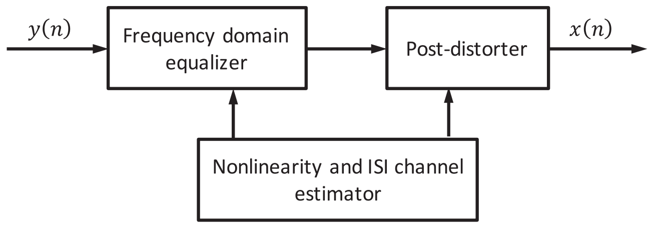

Fig. 3. Block diagram of receiver structure.

TABLE 1
Simulation Parameters for Optical Wireless Channel

Parameters	value
Room size	$5 \times 5 \times 3\ m^3$
Reflection coefficient	$\rho = 0.8$
Power	$1\ w$
Lambert's order	1
LED location (x, y, z)	$(2.5, 2.5, 2.15)$
Receiver location (x, y, z)	$(2.5, 2.5, 0)$
Semi-angle at half power	70°
FOV of a receiver	60°
Active area	$1\ cm^2$

memoryless nonlinear function can be represented as

$$g(x(n)) \approx \sum_{k=1}^{K} a_k x(n)^k \approx \frac{\sum_{k=1}^{K} b_{k,0} \Psi^k(x(n)) + A_0}{c_0}. \tag{30}$$

Although, we do not know the values of $\{a_k\}$ for the memoryless polynomial $g(x(n))$, we can use the polynomial at the right hand side of (30) as a replacement for $g(x(n))$. Note that, the value of c_0 is unknown. But it can be absorbed into the polynomial coefficients $\{a_k\}$, so we can simply let $c_0 = 1$.

3.2.2) Frequency Domain Equalization and Post Distortion: With the definition of the composite channel vector c and the Hammerstein model for LED, we can rewrite (14) as

$$\mathbf{y} = \mathbf{g} * \mathbf{c} + \mathbf{n}, \tag{31}$$

where $\mathbf{g} = [g(x(M)), g(x(M+1)), \ldots g(x(N)]^T$ given by (13). According to (15), the system can be regarded as a concatenation of memoryless nonlinear LED and an ISI channel. Hence, besides the nonlinearity and ISI channel estimator, we employ an equalizer to combat ISI, followed by a post-distorter to tackle the nonlinear distortion. The structure of the receiver is shown in Fig. 3. In particular, we propose to use frequency domain equalization due to its low complexity. Assuming that cyclic prefixing is used in the single carrier system, we construct an $(N-M+1)\times(N-M+1)$ circulant composite channel matrix $\mathbf{C}$ with its first column given by $\widetilde{\mathbf{c}} = [\mathbf{c}, \mathbf{0}]^T$, where the length of vector $\mathbf{0}$ is $(N-2M-L+2)$. The linear convolution model in (15) can be rewritten in a matrix form as

$$\mathbf{y} = \mathbf{C}\mathbf{g} + \mathbf{n}. \tag{32}$$

Since the circulant matrix $\mathbf{C}$ can be diagonalized by the DFT matrix, i.e., $\mathbf{F}\mathbf{C}\mathbf{F}^{\mathbf{H}} = \mathbf{D}$, where $\mathbf{F}$ is the normalized DFT matrix (the $(m, n)th$ element is given by $N^{-1/2}e^{-j2\pi mn/N}$, where $j = \sqrt{-1}$), and the diagonal matrix

$$\mathbf{D} = Diag\{d_M, d_{M+1}, \ldots, d_N\},$$

whose diagonal elements $[d_M, d_{M+1}, \ldots, d_N]^T = \sqrt{N-M+1}\mathbf{F}\widetilde{\mathbf{c}}$. Model (32) can be rewritten as

$$\mathbf{F}\mathbf{y} = \mathbf{F}\mathbf{C}\mathbf{F}^{\mathbf{H}}\mathbf{F}\mathbf{g} + \mathbf{F}\mathbf{n}. \tag{33}$$

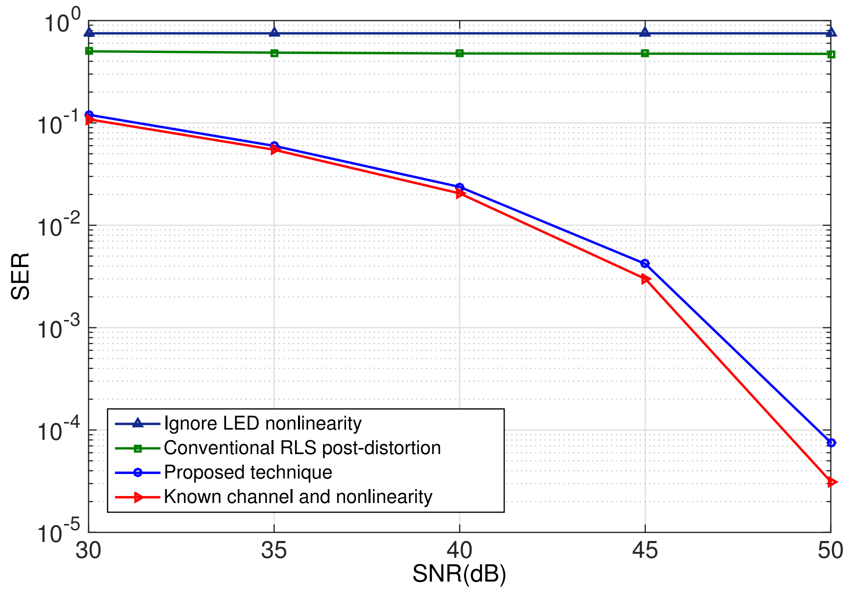

Fig. 4. SER performance comparison of receivers with 8-PAM and K = 4.

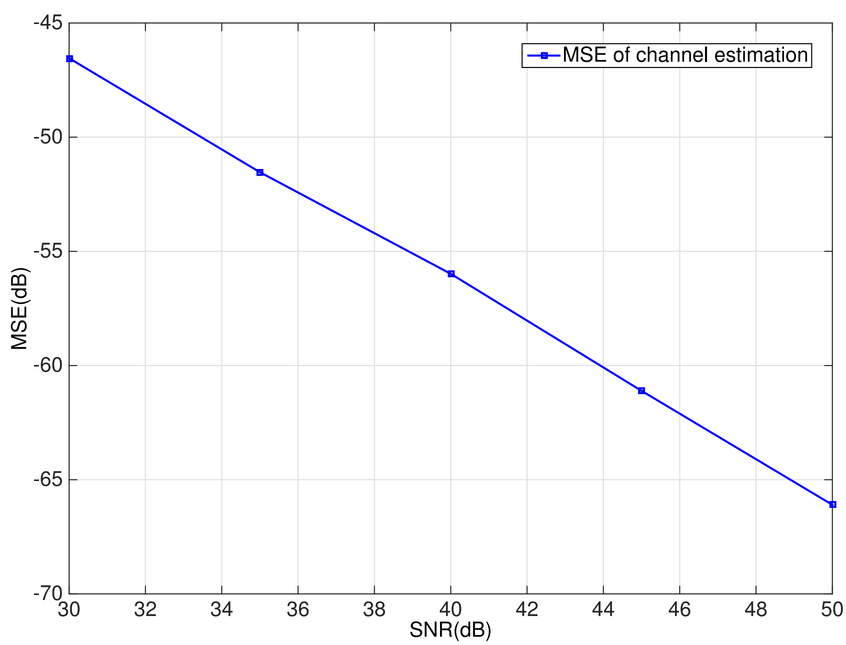

Fig. 5. MSE of channel estimation.

Then we have

$$\mathbf{Fy} = \mathbf{DFg} + \mathbf{Fn}. \tag{34}$$

With received signal vector $\mathbf{y}$ and combined channel and memory coefficients $\mathbf{c}$ estimated from the previous part, the LS estimate of the memoryless nonlinear signal $g(x(n))$ can be expressed as

$$\widehat{\mathbf{g}} = \mathbf{F}^{\mathbf{H}}\mathbf{D}^{-1}\mathbf{Fy}. \tag{35}$$

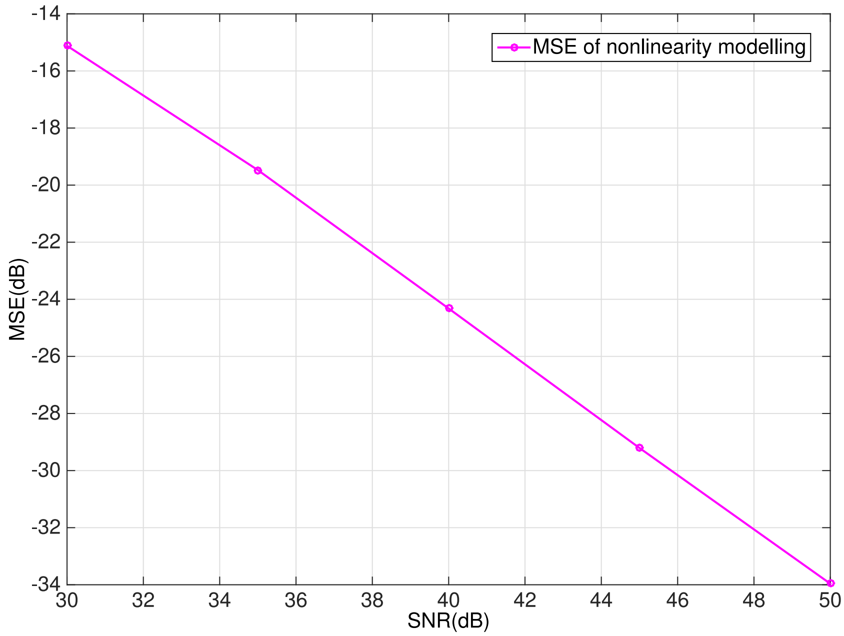

Fig. 6. MSE of nonlinearity modelling.

The output of the FDE is input to the post-distorter for nonlinearity mitigation, where the post-distorter is an inverse function of the estimated nonlinear model.

4. Simulation Results

In this section, we examine the system performance with the proposed technique in terms of symbol error rate (SER). Also, we compare the mean square error (MSE) of the estimated composite channel and the MSE of the nonlinearity modelling. To account for the limited LED bandwidth, the LED memory length is set to be 4. The LED nonlinear transfer function is given by the Hammerstein model [28]

$$z(n) = \sum_{k=1}^{K} a_k x(n)^k + \sum_{m=1}^{4} \lambda_m \left(\sum_{k=1}^{K} a_k x(n-m)^k \right), \tag{36}$$

where to ensure that the LED acts as a low-pass filter [19], the coefficients $\lambda_1 = 0.3$, $\lambda_2 = 0.25$, $\lambda_3 = 0.2$ and $\lambda_1 = 0.1$, and the polynomial coefficients $\{a_k\}$ are obtained based on an LED data sheet [29]. We choose $K = 4$ and the coefficients are $a_1 = 34.11$, $a_2 = -29.99$, $a_3 = 6.999$ and $a_4 = -0.1468$. We consider a room of $5 \times 5 \times$ 3 m^3 with a single light source on the centre of the celling. Details of the key simulation parameters for the optical wireless channel are summarized in Table I. Based on the channel model in Section 2, where both LOS component and the first order reflection as the NLOS component of the channel are considered, we can determine the value of the optical wireless channel taps, where we take 5 dominant channel taps and ignore the small ones. The length of training sequence is 100 and 8-PAM is used.

Fig. 4 shows the SER performance of the LED system with various receivers, where the performance of the receiver with the perfect knowledge of composite channel coefficients and LED nonlinearity is also shown for reference. In Fig. 4, the curve with legend "Ignore LED nonlinearity" denotes the performance of the conventional LS-based linear equalizer where both the LED nonlinearity and memory effect are ignored. It can be seen from the figure that, due to the severe impact of LED nonlinearity and memory effect, the receiver simply does not work. We also show the performance of the recursive LS (RLS) based post-distortion technique [12] in Fig. 4, where it is denoted by “conventional RLS post-distortion”. It is found that, due to the ill-condition of the correlation matrix, the technique does not work properly either. In contrast, our proposed technique (where the LED nonlinearity and composite channel coefficients are jointly estimated) works very well and its performance is very close to that of the receiver with known LED nonlinearity and composite channel coefficients. Fig. 5 shows the MSE of the composite channel estimation and Fig. 6 shows the MSE of nonlinearity modelling, where the simulation settings are the same as those in Fig. 4. The results again demonstrate the effectiveness of our proposed technique.

5. Conclusion

In this paper, we have investigated joint estimation of LED nonlinearity and optical wireless channel based on the orthogonal polynomial technique. The receiver has been designed, where frequency domain equalization is used to combat ISI, followed by a post-distorter to tackle the LED nonlinearity. The effectiveness of the proposed technique has been verified by simulation results.

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