Advantage of CAP Signaling for VLC Systems Under Non-Linear LED Characteristics

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Abstract—We consider the use of carrier-less amplitude and phase (CAP) modulation together with frequency domain equalization (FDE) for high-speed data transmission in indoor visible-light communication systems and investigate the suitability of this transmission scheme under non-linear LED characteristics, which has to be dealt with in practical implementations of these networks. We focus on the non-linear characteristic of the LED within its dynamic-range, while neglecting the effect of upper signal clipping which has been subject of extensive research so far. The LED is modeled by the Hammerstein model, which consists of a linear time-invariant system followed by a memoryless non-linear part. Through Monte-Carlo simulations, we show that, for a given spectral efficiency, the CAP-FDE scheme suffers less from LED non-linearity in terms of electrical power, as compared with the classical DC-biased optical orthogonal-frequency-division-multiplexing (DCO-OFDM).

I. INTRODUCTION

Visible-light communications (VLC) are well known as a promising solution for providing high data-rate wireless access in indoor environments, with a number of advantages over the traditional radio-frequency (RF) technologies [1]. Assuming the existence of a line-of-sight (LOS) between the transmitter (Tx) and the receiver (Rx), the main limitation on the transmission data rate arises from the limited bandwidth (BW) of the emitter [2], [3], which is typically a white LED with a modulation BW of a few MHz. So, to realize high data-rate transmission, spectrally efficient signaling schemes should be used. For this reason, modulations based on multiple subcarriers including DC-biased optical orthogonal-frequency-division-multiplexing (DCO-OFDM) and asymmetrically-clipped O-OFDM (ACO-OFDM) have been proposed within this context [4], [5]. Recently, single subcarrier schemes such as pulse amplitude modulation (PAM) and carrier-less amplitude and phase (CAP) modulation have been proposed, which have the main advantage of exhibiting low peak-to-average power ratio (PAPR), as compared with O-OFDM-based techniques [6]–[8]. Note that a high PAPR is disadvantageous for two reasons: it decreases the power efficiency of the system, and in addition, it requires a large dynamic range (DR) of the optical front-ends, otherwise resulting in severe performance degradation due to non-linear distortion. Another disadvantage of O-OFDM-based schemes is the loss in the spectral efficiency due to the Hermitian symmetry constraint, in particular for ACO-OFDM [7]. PAM- and CAP-based transmission also offer simpler transceiver structure and more robustness to frequency and phase noise. Meanwhile, due to

the LED BW limitations and probably time-dispersive channel conditions (due to the existence of multiple LOS paths or absence of LOS [3]), there is a need to channel equalization at the Rx in order to make high data-rate communication possible. The merits of frequency domain equalization (FDE) was recently shown in [7] where the channel equalization is performed in the frequency domain, allowing low complexity implementation of the Rx. This is done through adding a cyclic prefix (CP) to the frames of symbols at the Tx to avoid inter-frame interference; at the Rx, after CP removing, FDE is performed through taking a fast-Fourier transform (FFT) on the frame of received symbols, single-tap equalization, and an inverse FFT to come back to the time domain [8], [9].

In [7], we analyzed in particular the performance of CAP-FDE and showed its advantage over O-OFDM-based signaling in terms of reduced PAPR and Tx implementation complexity. We also showed that this scheme outperforms PAM-FDE for high transmission rates, quantified in terms of the ratio of bit rate R_b to LED 3-dB modulation BW f_c . Therein, we assumed a linear characteristic within the LED DR and only took into account signal clipping. In practice, in addition to clipping, the signal undergoes further distortion due to the non-linear “output power - driving current” characteristics of the LED as well as the non-linearity of the power amplifier at the Tx [10]. Here, our aim is to investigate the suitability of CAP-FDE when taking into account the non-linear LED characteristics. Meanwhile, we consider accurate modeling of this non-linearity as it is essential for designing appropriate mitigation techniques, which is the subject of a future work.

The remainder of this paper is organized as follows: After describing the LED non-linearity in Section II, we review in Section III the commonly-used approaches to model the non-linear LED transfer function (TF) and discuss their pros and cons in terms of accuracy and complexity. Then, in Section IV, we attempt to investigate the performance degradation caused by this non-linearity for DCO-OFDM and CAP-FDE signaling schemes. Lastly, we summarize our findings in Section V.

II. LED NON-LINEAR CHARACTERISTICS

In addition to signal clipping due to the limited DR of the Tx, the link performance can be degraded due to the non-linear characteristics of the Tx/Rx front-end, including the LED, the photo-detector (PD), and D/A

and A/D converters. Among these, the LED is the major source of non-linearity in VLC systems. The non-linearity of the LED TF arises from two phenomena. Firstly, the imperfection of the driving circuits of the LED leads to a non-linear voltage-current relationship. In particular, the LED has a limited voltage dynamic range, the lower limit being known as the turn-on voltage (TOV) and upper limit as the saturation voltage. Below the TOV, the LED is considered to be in the cut-off region and not conducting any current. The saturation voltage corresponds to the maximum permissible AC/pulsed current above which the LED encounters overheating that can damage it. Signal amplitudes that fall outside this operation range should be clipped before being fed to the LED, which results in non-linear distortion called the clipping noise [11]. If the signal amplitude is approximately modeled by a zero-mean Gaussian random variable, then this double-side clipping noise can be well modeled as an attenuation factor plus a non-correlated additive noise according to the Bussgang theorem [10]. In addition, the non-linear TF within the DR induces additional distortion, which should be treated as dynamic-range-limited non-linearity [12]. On the other hand, the number of emitted photons of the LED is not directly proportional to the amplitude of the injected electrical current in its active region, resulting in a non-linear transfer function between the input current and the output optical intensity. Figure 1 shows an example of the non-linearity between the voltage, current and output optical intensity of a typical white LED (LE CW E2B from ORSAM). Furthermore, note that the response of an LED depends not only on the current value of the injected electrical signal, but also on the previous input, giving rise to a memory effect. In fact, this memory effect is due to the fact that the carrier-density response of the LED depends on the frequency of the injected current [13], [14].

It is worth mentioning that the non-linear characteristics of the LEDs may change over time due to component aging and temperature drift [14]. The electrical-optical conversion efficiency slowly degrades due to temperature rise caused by LED self-heating. That means the output intensity decreases over time for the same input driving current. Moreover, the temperature may change suddenly for different illumination levels for supporting the demand of dimming control in VLC systems, which in turn results in variations of the non-linear characteristics of the LED [12].

III. MATHEMATICAL MODELING OF LED NON-LINEARITY

Depending on whether or not we take into account the memory effect in the non-linearity of an LED, there exist two types of memory-less and memory models. For memory-less models, the common approach is the so-called memory-less polynomial model by which the non-linear TF of the LED is approximated using a power series [15], [16]. This approach is very popular in VLC systems due to its simplicity but it has a very limited accuracy, especially at high transmission rates. As a matter of fact,

Fig. 1: The relationship of (a) voltage-current (b) current-optical power of a commercial white LED (OSRAM LE UW Q9WP).

since the response of the LED depends on the frequency of the input current, the LED exhibits strong memory effects as the signal BW increases. For this reason, we focus on the memory models in the following.

A. Volterra series based model

Volterra series, which are in fact the generalization of power series with memory [17], have been longly used to model non-linear systems with memory. They were first introduced by Vito Volterra as a generalization of the Taylor series of a given function. In fact, the output $y(t)$ for an linear time-invariant (LTI) system can be written as $y(t) = \int_{-\infty}^{\infty} h(\tau) x(t-\tau) d\tau$, where $h(t)$ is the system impulse response and $x(t)$ the input signal. This LTI system is essentially a *memory* system since the output $y(t_0)$ depends not only on the present value of the input $x(t_0)$, but also on its past and future values. Obviously, for a causal system, $h(\tau) = 0, \tau < 0$. For a non-linear memoryless system, the input-output relationship can be approached using power series as $y(t) = \sum_{n=0}^{\infty} c_n x^n(t)$. The Volterra series is a combination of non-linear mem-

oryless and linear memory models, defined as [17]:

$$y(t) = h_0 + \sum_{k=1}^{\infty} \int_{-\infty}^{\infty} \dots \int_{-\infty}^{\infty} h_k(\tau_1, \dots, \tau_k) \times \quad (1)$$

$$x(t - \tau_1) \dots x(t - \tau_k) d\tau_1 \dots d\tau_k$$

where h_0 is a constant and $h_k(\tau_1, \dots, \tau_k)$ are the kernels of the Volterra series, which determine the non-linearity and are of infinite length. For practical implementations, a discrete-time truncated version with finite LTI length and linearity order is more commonly used as follows:

$$y(n) = h_0 + \sum_{m=1}^M \sum_{k_1=0}^{K_1} \dots \sum_{k_m=0}^{K_m} h_m(k_1, \dots, k_m) \times \quad (2)$$

$$x(n - k_1) \dots x(n - k_m),$$

where M is the highest non-linearity order, and K_m is the memory length of the m -th order Volterra kernel.

It was shown in [13] that a second-order Volterra series can already provide a fair approximation to the power spectrum of the LED up to 14 MHz, which achieves the best accuracy compared with other models. However, the main drawback of this model is its relatively high complexity. In fact, the number of coefficients of the n -th order discrete Volterra kernel $h_n(\tau_1, \dots, \tau_n)$ with a memory filter length of L is L^n . Thus, the total number of coefficients of the corresponding model increases exponentially as the non-linearity order and the memory length increase, which makes it impractical for real-time applications. The use of Volterra series based modeling for VLC systems was considered in [18], [19], where a second-order Volterra series was used, making a good compromise between complexity and performance.

B. Memory polynomial model

The generalized memory polynomial model is a simplification of the Volterra-based model by considering only the diagonal coefficients in Volterra kernels [18], [20]. In other words, for the n -th order Volterra kernel, we set $h_n(k_1, \dots, k_n) = 0$ except for $k_1 = \dots = k_n$. Then, from (2),

$$y(n) = h_0 + \sum_{m=1}^M \sum_{k=0}^K a_{mk} x(n - k)^m, \quad (3)$$

where $a_{mk} = h_m(k, k, \dots, k)$, M is the non-linearity order, and K is the memory length. The number of the coefficients of this model is on the order of $(K + 1)$, compared to $(K + 1)^M$ for the Volterra series model. This approach, considered for instance in [21] for the VLC context, offers a good compromise between accuracy and complexity, compared with memoryless and the Volterra series based models.

C. Wiener and Hammerstein models

Further simplification of the Volterra series leads to the so-called Wiener [21], [22] and Hammerstein [19], [23] models. The Wiener model consists of an LTI system followed by a memory-less non-linear function. The Hammerstein model is similar to the Wiener model, while

TABLE I: Fitted polynomial coefficients c_k of degree k of the LED according to the Hammerstein model [10].

c_3	c_2	c_1	c_0
0.29	-1.09	2.06	-0.0003

TABLE II: Simulation parameters.

Parameter		Value
Tx/Rx	LED DR	[0.1, 1] A
	LED 3-dB BW	20 MHz
	PD type	PIN
	N_0	10^{-21} W/Hz
Modulation	FFT size	256
	CP length	12
	Clipping factor	10 dB
VLC channel	Type	Flat
	LOS gain	10^{-4}
Pulse shaping filters	Type	Square-root raised cos
	Up-sampling factor	6
	Filter span	16

the order of the LTI system and the memoryless non-linear function is reversed. In addition to their simplicity, the separation of the memoryless non-linear and linear memory parts allows a simple equalization at the Rx. The memoryless non-linear TFs can be compensated using a simple pre-distorter at the Tx or a post-distorter at the Rx, as studied in [23]. The linear memory part, on the other hand, can be equalized using time-domain or frequency-domain equalization [24]. One drawback of these models is their relatively low accuracy, compared to other non-linear models, in particular, for large modulation bandwidths [13].

IV. IMPACT OF LED NON-LINEARITY

Here we present a set of numerical results to investigate the impact of the non-linear LED TF on the performances of DCO-OFDM and CAP-FDE. For the reason of simplicity, we consider a Hammerstein model for the LED, where the memoryless non-linearity block of is approximated by a third degree polynomial with the coefficients provided in Table I for the considered white LED. The LTI block of the model is assumed to be a first-order low pass filter, as described in [7]. The other simulation parameters are provided in Table II. As explained previously, in order to investigate the impact of non-linear LED TF, we do not consider the high SNR region, which is dominated by excessive clipping noise, see [7].

Figure 2 shows the BER performance of 32-QAM DCO-OFDM and 32-CAP-FDE as a function of the electrical signal-to-noise ratio (SNR) *calculated with reference to the Tx power* for different values of R_b/f_c . Here, $E_{b,elec}$ denotes the *transmitted* electrical energy per information bit, and N_0 stands for the Rx noise unilateral power spectral density.

The comparison with the case of a linear LED TF is somehow delicate. In fact, several approaches could be used to consider a linear approximation of the non-linear TF given in Table I. For instance, we may use the linear fitted line in purple in Fig. 3 or the simple $y = x$ line (in blue) to represent a *hypothetic* linear LED case. Here, we consider the latter case for the input current - optical

Fig. 2: BER performance for the cases of the considered linear TF and the non-linear LED for different signal BWs (a) 32-QAM DCO-OFDM (b) 32-CAP-FDE.

power relationship of the LED. Note that, this way, from Fig. 3 for the same input current, the non-linear LED has a larger output optical power than the linear one. As a result, in Fig. 2, we see a better performance for the non-linear case, which could be misleading. In fact, a direct comparison of the BER performances makes no sense here. Instead, what we do here, is to compare the difference of the required SNR to achieve a given BER for the linear and non-linear cases for different modulation schemes. This way, we can determine which transmission scheme is more sensitive to the signal distortion caused by the LED non-linearity. For example, considering a target BER of 10^{-5} , the SNR penalties for DCO-OFDM are about 5.5 and 4.8 dB for $R_b/f_c = 5$ and 10, respectively, whereas for CAP-FDE, they are about 5 and 4 dB. The fact that the SNR penalties are larger for the case of DCO-OFDM, testifies its higher sensitivity to LED non-linearity, as compared with CAP-FDE.

We have also compared the BER performance of DCO-

Fig. 3: TFs for the case of linear and non-linear LED.

OFDM and CAP-FDE for different constellation sizes for a fixed $R_b/f_c = 10$, see Fig. 4. As before, we compare the performances of linear and non-linear cases in terms of SNR penalty. We notice that DCO-OFDM suffers from more SNR penalty with increased constellation size. For instance, the SNR penalty between the linear and non-linear LED cases is about 5, 5, and 4.8 dB at $\text{BER} \approx 10^{-5}$ for the cases of 4-QAM, 16-QAM, and 64-QAM DCO-OFDM, respectively. The CAP-FDE scheme benefits from a lower penalty, which is around 3.8, 3.8, and 2.9 dB for the cases of 4-CAP, 16-CAP and 64-CAP, respectively.

V. CONCLUSIONS

In this paper, after presenting the different approaches for modeling LED non-linearity for VLC applications, we investigated the impact of this non-linearity on the performance of DCO-OFDM and CAP-FDE transmission schemes by considering a Hammerstein model. We showed that DCO-OFDM is more sensitive to signal distortion caused by this LED non-linearity than CAP-FDE. However, due to the limitations of the approach we used, we quantified indirectly the SNR penalty arising from the non-linear LED TF, compared to the linear case. The on-going research focuses on proposing efficient techniques to reduce the impact of this non-linearity.

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Fig. 4: BER performance for the case of the considered linear TF and the non-linear LED for different constellation sizes of (a) DCO-OFDM (b) CAP-FDE; $R_b/f_c = 10$.

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