On the DC balance of multi-level PAM VLC systems

Mike Wolf and Martin Haardt
Communications Research Laboratory, Ilmenau University of Technology
Tel: (+49) 3677 69 2619, e-mail: mike.wolf@tu-ilmenau.de

ABSTRACT
Lighting requirements and their impact on visible light communication (VLC) systems based on multi-level pulse-amplitude modulation are discussed. We will show that the moving average of the signal must not fluctuate too much. Two solutions are presented. One is based on a simple extension of the well-known 5B6B line coding to codes with M-ary symbols. The other is based on the so-called Hadamard-coded modulation. Closely related to the topic of DC balance is the resistance of the received signal to high-pass filtering. The impact of such filtering on multi-level PAM systems with frequency domain equalization is investigated. We will show that the demand for flicker-free lighting can easily be met with line codes. Together with an equalization, which has to be used for multi-level PAM anyway, the AC coupling at the receiver is no problem either. The feasibility of an adaptive modulation is also discussed.
Keywords: DC-balance, 5B6B line coding, visible light communication, wireless optical communication, plastic optical fiber communication, AC-coupling

1 INTRODUCTION

In recent years, the interest in wireless optical communication as a supplement to radio has increased enormously. First and foremost is Visible Light Communication (VLC), where light sources based on LEDs are used not only for their original purpose, lighting, but also for communication. With regard to wireless infrared (IR) communication, which has been intensively discussed in the 1990s, there are many similarities such as the lack of small scale fading, the spatial limitation of radiation, or, from a technical point of view, the necessity of intensity modulation and direct detection (IM/DD) [1]. A major difference, however, is the possible modulation bandwidth of the source and its optical power. The output power of LED light sources is (usually) much higher than that of IR transmitters, but on the other hand the modulation bandwidth is much smaller. Therefore, it is not surprising that on-off keying (OOK) and pulse-position modulation have been favoured as IR modulation schemes [1], while higher-order modulation schemes are particularly interesting for VLC. These higher-order schemes counteract the limited modulation bandwidth and the associated susceptibility to inter-symbol interferences (ISI).

A question that always arises is whether the transmission should be parallel, i.e., based on OFDM or discrete multitone transmission (DMT) [2], or simply serial, based on PAM [3] or single-subcarrier modulation [4]. In this paper we restrict our attention to PAM, or rather, to *two specific problems* that arise in connection with VLC. *I.)* Is it possible to prevent any visible LED flickering by means of coding? An uncoded multi-level PAM source signal is by nature not DC balanced, but a DC balance can be achieved, e.g., by a suitable line coding scheme or the Hadamard-coded modulation technique proposed by [10]. *II.)* A typical VLC receiver is AC-coupled to suppress the current generated by ambient light or to decouple amplifier stages. But does the unipolar optical signal at the receiver input allow high-pass filtering? Such a signal may suffer from a baseline wander [1].

With DMT, of course, the question of a possible flickering of the LED intensity does not arise, because a DC balance can be forced by not using the subcarrier with frequency 0. Flickering was not an issue with wireless IR transmission either, as the IR radiation is not visible after all. Nevertheless, a DC balance, which can be guaranteed by Manchester coding, for example, can also be a great advantage for IR or plastic optical fiber transmission [5]. If the moving average of the optical signal is almost constant, which is synonymous with the fact that the continuous portion of the spectrum disappears near DC, then the baseline wander at the output of a high-pass filter is also correspondingly small.

As already mentioned, the main reason for using higher-level PAM is to achieve a very high data rate, even considering the limited LED modulation bandwidth. This purpose, however, requires an equalization anyway, because any higher-level modulation method is inherently very susceptible to ISI. An important assumption for the evaluation of the bit error rate (BER) will therefore be that the transmission is combined with a zero-forcing frequency domain equalization [3, 6]. This clearly allows to compensate any residual linear distortion caused by the high pass filter, too. Section 4 examines whether this also causes additional problems, e.g., increased noise. In Section 2 we will first derive the requirements for PAM from the point of view of good, absolutely flicker-free lighting.

2 LIGHTING REQUIREMENTS AND THEIR IMPACT ON THE DC BALANCE OF PAM

A human sense organ can perceive a change in the intensity of a stimulus above a certain relative threshold. With regard to the perception of changes in the luminous intensity, the threshold is around 1 % [7]. However, the flickering of a light source can only be perceived if the modulation frequency is below about 200 Hz [8]. Which requirements result from these two criteria for the DC balance of a multi-level PAM signal?

Let $\{z_n\}$ be an indexed sequence of unipolar PAM symbols with $z_n \in \{0, 1, \ldots, M-1\}$. Since the instantaneous optical power is proportional to z_n, the moving average $P_N(\ell)$ of the optical power over N symbols is

$$P_N(\ell) = \frac{1}{N} \sum_{n=\ell-N/2}^{\ell+N/2-1} P_0 \cdot z_n,$$

where P_0 should be the optical power for a series of ones. The length N corresponds to the integration time of at least $1/(200\text{ Hz})$. For equally probable symbol states, the expected value $P_0 \cdot E\big(z_n\big)$ is $P_0(M-1)/2$. This is the value we also want to achieve for $E\big(P_N(\ell)\big)$. If the luminous intensity is to vary by max. 1 %, the following must apply

$$0.995 \cdot P_0(M-1)/2 \leq P_N(\ell) \leq 1.005 \cdot P_0(M-1)/2$$

and thus

$$0.995 \cdot \frac{M-1}{2} \cdot N \leq \sum_{n=\ell-N/2}^{\ell+N/2-1} z_n \leq 1.005 \cdot \frac{M-1}{2} \cdot N. \tag{1}$$

How large is N? If we apply a very conservative PAM symbol rate of 2 MHz, $N = 2 \cdot 10^6/200 = 10\,000$ follows. Thus, the moving sum over 10 000 successive symbols z_n may fluctuate by a maximum of $\pm 25 \cdot (M-1)$. The following section shows whether this requirement can be met using fixed-length line codes, our first approach.

3 LINE CODING FOR MULTI-LEVEL PAM: AN EXTENSION OF THE 5B6B CODE

The 8B10B IBM code [9] is, besides the Manchester code, certainly the best known and most important DC balanced line code. The IBM code consists of a 5B6B code (i.e., 5 source bits are mapped into 6 coded bits) and a 3B4B code. For 2-PAM (i.e., OOK) we use exactly the above 5B6B code. Since the running disparity[1] after each codeword is at most ± 1, and the codeword itself consists of only 6 symbols, the requirements according to (1) are met with ease.

The basic idea of the 5B6B code is to use all 20 perfectly balanced 6 bit code words, i.e., those that have exactly 3 ones and 3 zeros. But since 32 codewords must be available, codewords with 4 ones and 2 zeros (disparity +1) and 2 ones and 4 zeros (disparity -1) are also used. To avoid a disparity accumulation, the code words $[101011]$ and $[010100]$, for example, are assigned to the same source code word. If the running disparity, which is permanently updated, is negative, $[101011]$ is assigned, otherwise $[010100]$.

It is absolutely obvious to use the same strategy to generate balanced 4-PAM signals. Exactly this was suggested by Gaudino and Nespola [5]. The authors denote the code as "5S6S", where "S" stands for symbol. According to their suggestion, all 580 possible perfectly balanced codes are used, regardless of the running disparity. The still missing $(4^5 - 580) = 444$ codewords are selected from all 546 available codewords with a disparity of $+1$, whereby the complementary versions of these codewords are used as described above depending on the running disparity. For the bipolar version $[+3+3-1-1-3+1]$ of a unipolar code word $[331102]$ this would be $[-3-3+1+1+3-1]$, for example.

We propose a slightly modified version here. Instead of assigning uncoded M-PAM symbols of a source alphabet to DC-balanced M-PAM symbols, the encoded M-ary symbols are generated directly from the binary source code words. For example, our code is denoted as 10B6Q if 10 bits are directly assigned to a 6-digit code word with quaternary symbols (4-PAM). It is called 8B6T if 8 source bits are mapped to 6 ternary symbols. We have generated all codes up to 9-ary symbols. Fig. 1 (a) shows the power spectral densities of the coded sequences for the 5B6B code, the 8B6T code, the 12B6Qiu code (Qui=Quinary, 5-ary), and the 16B6N code (N=Novenary, 9-ary). All codes have a similar redundancy of about 20 %.

The modulation orders $M = 2, 3, 5$ and 9 are particularly interesting because at the same average optical power each increase of M, $M \in \{2, 3, 5, 9\}$, exactly halves the distances between two neighbouring symbol states, see Fig. 1 (b) (blue dots). In other words, if, for example, 8 LEDs are available which can only be switched on or off, 2-PAM, 3-PAM, 5-PAM and 9-PAM can be realized easily with a constant average optical power. For 2-PAM, all 8 diodes are switched on or off together . For 3-PAM, groups of 4 diodes are addressed, for 5-PAM groups of 2 diodes and, finally, for 9-PAM, each individual LED can be activated or not. Therefore, for a constant PAM symbol rate $1/T$, the data rate is increased by the factor $\log_2(9)$ when changing from 2-PAM to 9-PAM, while the minimum Euclidian distance decreases by the factor 8.

What are the properties of the codes we have constructed? Except for the 12B6Qui and 16B6N code, all codes have a maximum disparity of ± 1. The codes for $M = 5$ and $M = 9$ have a maximum disparity of ± 2 assuming symbols from the alphabet $\{0, 1, \ldots, M-1\}$. Accordingly, the sum of all symbol states within a code word, which

[1] the accumulated disparity up to the current time point

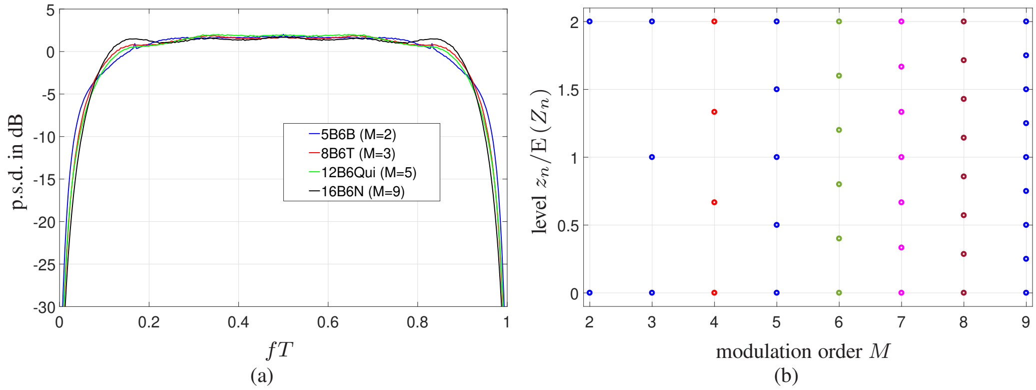

Figure 1. In (a) the normalized power spectral densities of the coded symbol streams for $M = 2$, 3, 5 and 9 are shown. Figure (b) shows the PAM symbol levels for a fixed average optical power. The peak-to-average power ratio (PAPR) of NRZ-PAM is therefore 2, independent of M.

ideally should be $6(M-1)/2$, varies by a maximum of ± 2. Thus the requirements according to (1) are met with ease even for $M > 2$.

4 HADAMARD-CODED MODULATION (HCM): A BETTER SOLUTION?

The line coding discussed in Section 3 increases the gross data rate by a factor of 6/5. This increases the required average optical power P at least by a factor of $\sqrt{6/5}$, about 0.4 dB. Are there better alternatives? In [10] an approach named HCM has been proposed, which should be characterized by a particularly *low* PAPR. It is inherently DC-balanced. We will briefly analyze the procedure where we use a slightly different notation than [10]. Our modeling quickly shows the core of the HCM approach and we can decide whether it is preferable to line coding.

Let $\boldsymbol{H}$ be a Hadamard matrix of size $N \times N$ with *bipolar* elements. The vector $\boldsymbol{X} = \begin{bmatrix} 0 & z_1^{\pm} & \ldots & z_{N-1}^{\pm} \end{bmatrix}^{\mathrm{T}}$ contains $(N-1)$ bipolar M-PAM symbols with $z_\mu^{\pm} \in \{\pm 1, \pm 3, \ldots, \pm(M-1)\}$. The transformation of $\boldsymbol{X}$ then results in a zero-mean vector $\boldsymbol{x}^{\pm} = \begin{bmatrix} x_0^{\pm} & \ldots & x_n^{\pm} & \ldots & x_{N-1}^{\pm} \end{bmatrix}^{\mathrm{T}}$ with

$$\boldsymbol{x}^{\pm} = \boldsymbol{H} \cdot \boldsymbol{X} = \begin{bmatrix} \boldsymbol{h}_0 & \boldsymbol{h}_1 & \ldots & \boldsymbol{h}_{N-1} \end{bmatrix} \cdot \boldsymbol{X} = \sum_{\mu=1}^{N-1} \boldsymbol{h}_\mu \cdot z_\mu^{\pm}.$$

Since all columns $\boldsymbol{h}_\mu$ of $\boldsymbol{H}$ are orthogonal in pairs, the HCM-scheme can also be considered as orthogonal code division multiplex, where each $\boldsymbol{h}_\mu$ represents one of N orthogonal basis vectors. The vector $\boldsymbol{h}_0$ is not used to ensure a zero mean $\boldsymbol{x}^{\pm}$. It is immediately clear that the discrete sequence $\{x_n^{\pm}\}$, i.e., the continuous repetition of vectors $\boldsymbol{x}^{\pm}$, no longer has much to do with an M-PAM sequence. As with any parallel transmission, the distribution of $x_n^{\pm}$ for big N tends towards a Gaussian distribution with a *large* PAPR. While a bipolar sequence $\{z_\mu^{\pm}\}$ has the maximum amplitudes $\pm(M-1)$, the sequence $\{x_n^{\pm}\}$ has the maximum values $\pm(N-1)(M-1)$. A big advantage of PAM is thus lost.

In addition, the procedure is very power inefficient if we consider that the transmitted signal $p(t)$ must be unipolar. In order to calculate the minimum Euclidean distance, cf. [3], we assign the continuous time unit-energy basis functions $\psi_\mu(t)$ to the basis vectors $\boldsymbol{h}_\mu$. For rectangular pulse shaping, the vector $\boldsymbol{h}_{N/2}$, for example, corresponds to the basis function

$$\psi_{N/2}(t) = \frac{1}{\sqrt{T_{\mathrm{hcm}}}} \left(\mathrm{rect}\left(\frac{t - T_{\mathrm{hcm}}/4}{T_{\mathrm{hcm}}/2} \right) - \mathrm{rect}\left(\frac{t - 3T_{\mathrm{hcm}}/4}{T_{\mathrm{hcm}}/2} \right) \right),$$

where $T_{\mathrm{hcm}} = \log_2(M)(N-1)/R_{\mathrm{b}}$. In order to assign a unipolar optical signal $p(t)$ with *constant* average power, a bias must be added which gives

$$p(t) = P_0 \cdot \sum_{\mu=1}^{N-1} z_\mu^{\pm} \cdot \psi_\mu(t) + \underbrace{P_0 \cdot (M-1)(N-1)}_{=P \text{ (bias)}}.$$

The minimum Euclidean distance at the photo diode output (responsivity R, measured in A/W) is thus

$$d_{\min} = 2\sqrt{T_{\mathrm{hcm}}} R P_0 = 2RP\sqrt{\log_2(M)/\left((N-1) \cdot R_{\mathrm{b}}\right)}/(M-1). \tag{2}$$

The equation shows immediately that for the same distance (and thus approximately the same BER) at the receiver we need a factor $\sqrt{N-1}$ of more average optical power compared to serial PAM transmission with $(N-1)=1$. Even for $N=16$ we almost lose a factor of 4 in power efficiency compared to uncoded PAM transmission, but the authors even suggest $N=128$ (10.5 dB loss).

HCM is very similar to OFDM, since its a parallel orthogonal transmission. But even if the scheme ensures better resistance to nonlinearities than OFDM [10], a serious disadvantage remains: the basis vectors $\boldsymbol{h}_\mu$ of the Hadamard matrix are no longer the eigenvectors of the system matrix, in contrast to the basis vectors of the DFT matrix. This eliminates the enormous advantage of 1-tap equalization in the frequency domain based on a DFT. However, multi-level PAM requires equalization. Due to the high PAPR and the poor power efficiency mentioned before, we will not further consider this approach although, of course, it could be combined with additional equalization [10].

5 LINE CODED PAM WITH FREQUENCY DOMAIN EQUALIZATION: PERFORMANCE

We assume a PAM system with zero-forcing frequency domain equalization to counteract the modulation distortions caused by the limited LED modulation bandwidth. For this reason we assume a block structure with periodic unique-word insertion [11]. This cyclic structure of the signal, which is very similar to that of a signal with cyclic prefix insertion, allows a signal processing of the received signal by means of a DFT/IDFT. We assume blocks of 512 symbols, whereby the unique word itself consists of 32 symbols. The message part of a block can therefore contain $(512-32)/6=80$ code words, each with 6 PAM symbols.

First we analyze the required average optical power $P=\overline{p(t)}$ at the receiver, if the transmission only suffers from additive noise with a power spectral density N_0 (in A^2/Hz) in the electrical domain. The target BER is 10^{-3}. The corresponding signal current is $R \cdot p(t)$.

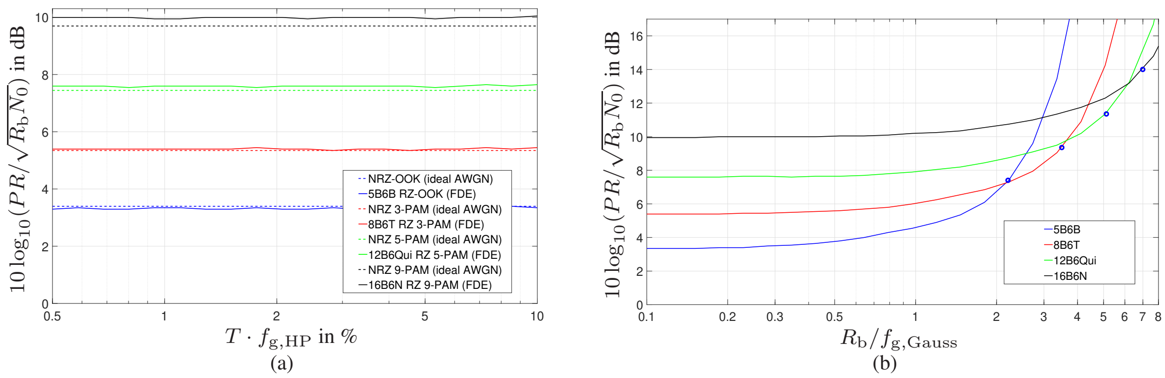

Figure 2. Normalized required average optical power at the receiver: figure (a) shows the dependency on the high pass filter cut-on frequency for AWGN. In figure (b) a Gaussian low pass is additionally considered, which limits the upper cut-off frequency of the system.

Fig. 2 (a) shows the influence of the high-pass cut-on frequency $f_{\mathrm{g,HP}}$, normalized with the PAM symbol rate $1/T$. The result shows that even for a product $f_{\mathrm{g,HP}} \cdot T = 0.1$, the high pass has no impact at all. This is not surprising. The high pass filter induced baseline wander, which remains for large products $f_{\mathrm{g,HP}} \cdot T$ despite the line coding, is equalized in the frequency domain. However, the noise at the demodulator input is not increased, since the high pass follows the noise sources in the transmission chain and not vice versa. Regarding the results, we have adopted the 50 % RZ pulse format, which always promises slightly better results than the NRZ format[2] [3]. The dashed lines show additionally the required normalized power in the ideal AWGN case with NRZ rectangular pulses. The values can be derived from (2) if $(N-1)$ is set to 1. A small disadvantage of the line coding approach becomes visible, especially with the 16B6N code (M=9). If an error occurs during transmission in one of the 6 PAM symbols of a code word, more than one error generally occurs in the decoded 16 bits. This causes an additional power penalty of about 0.3 dB for $M=9$.

Fig. 2 (b) shows the required power at the receiver as a function of R_b for $f_{\mathrm{g,HP}} \cdot T = 0.0125$, if the transmitted signal passes through a Gaussian low pass filter with with a 3 dB cut-off frequency $f_{\mathrm{g,Gauss}}$, measured in the optical domain. The results are very similar to those published in [4]. However, here the line codes are actually implemented, and the receiver contains the high pass filter. The picture clearly shows that using OOK, R_b cannot be increased arbitrarily, if the system bandwidth (here $f_{\mathrm{g,Gauss}}$) is limited.

The values, which are additionally marked with a blue dot, show which data rates can be achieved by the variation of M at a fixed symbol rate $1/T$ of about $2.2 f_{\mathrm{g,Gauss}}$. However, if the received optical power is not

[2] The spectral component at $f=0$ is not equalized but always set to zero. Each DFT block is thus zero mean in the time domain. This also simplifies the demodulation if the RZ format is used.

sufficient at this symbol rate even for $M = 2$, how can R_b then be further reduced?

A multiple repetition of the same code words within a DFT block at the same clock rate only makes sense, if the symbol clock rate is well below $f_{\mathrm{g,Gauss}}$, i.e., if the curves are flat. If the curve for $M = 2$ is not flat, ISI occurs before the equalization. This influence can only be completely removed by a longer symbol duration. Thus, if the system clock remains unchanged within the preamble and the unique word intervals, the symbols must be repeated within the codewords to increase the effective symbol interval T by a factor of 4, for example. From this point of view, the resistance to high-pass filtering even for large $T \cdot f_{\mathrm{g,Hp}}$ products shown in Fig. 2 (a) is extremely important for an adaptive system.

Of course, one could come up with the idea of setting the high-pass cut-off frequency very low in comparison to the system clock. But that would have at least two disadvantages. On the one hand, the settling time at the beginning of a transmission would be very long. During this time, the initially unipolar signal at the output of the high-pass filter is converted into a bipolar signal. However, a bipolar signal can be very important in the context of channel estimation and synchronization [11]. On the other hand, electronic-ballast fluorescent lamps, for example, can also cause interference in the hundreds of kHz range. A 500 kHz high-pass filter can attenuate these interferences by approximately 60 dB [12].

6 CONCLUSIONS

For systems with IM//DD, the use of serial PAM transmission is straightforward. From this point of view, for VLC systems, multi-level PAM is also a serious alternative to parallel transmission based on multiple electrical subcarriers (i.e., OFDM/DMT). For example, the peak-to-average power ratio for the NRZ pulse format is only 2, completely independent of the modulation order. The lighting requirements for good, flicker-free light are easily met by line codes that are an extension of the 5B6B code. These codes have a redundancy of approx. 20 %, but from the authors' point of view they represent a much better alternative than parallel transmission with orthogonal Walsh codes (Hadamard-coded modulation). Due to the short code words, coding and decoding can easily be done with the help of tables. Since a single symbol error can result in several bit errors at the decoder output, the application of a 16B9N code (N=novenary), for example, increases the required optical power by approx. 0.3 dB, assuming a BER of 10^{-3}. Of course, this penalty becomes smaller if channel coding with symbols from the GF(2^m) extension field is applied. The AC voltage coupling of the PAM receiver proves to be uncritical, if additional equalization is used. The high pass filter cut-on frequency can be as high as $0.1/T$ without additional loss. Without equalization, multi-level PAM does not work properly.

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