Synchronization and channel estimation for optical block-transmission systems with IM/DD

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ABSTRACT
The timing synchronization of OFDM/DMT systems is often based on auto-correlation algorithms. However, these algorithms do not provide a sharp correlation peak. In this work, we present a novel timing synchronization and channel estimation concept for optical block transmission systems with intensity modulation and direct detection. This concept is based on Golay complementary sequences. As these sequences are binary, multiplier-less matched filters can be implemented in order to use the cross-correlation function for the timing synchronization. Since Golay complementary sequences have ideal correlation properties, they enable an estimation of the channel impulse response directly in the time domain. We present an appropriate preamble structure and receiver concept for PAM-block transmission systems with frequency domain equalization and demonstrate the performance exemplary for 7B8B-coded binary PAM.

Keywords: visible light communication (VLC), wireless optical transmission, Li-Fi, polymer optical fiber (POF) communication, orthogonal frequency division multiplex (OFDM), discrete multitone transmission (DMT), block transmission, frequency domain equalization (FDE), channel estimation, synchronization, Schmidl-Cox algorithm

1 INTRODUCTION
For the block-wise transmission in dispersive optical channels with intensity modulation and direct detection (IM/DD), two basic concepts can be distinguished. While the first approach is based on discrete multitone transmission (DMT) [1], the second approach uses PAM or single-subcarrier transmission combined with frequency domain equalization [2]. Both concepts employ fast Fourier transform (FFT) based signal processing at the receiver. The precision of the FFT-window placement at the receiver and subsequently the quality of the channel estimation required to equalize the channel in the frequency domain have a significant impact on the overall system performance.

In this paper, we present a very efficient ‘single-shot’ block synchronization and channel estimation concept, where single-shot means that the synchronization is based on a short preamble (training sequence). Specifically, we discuss this synchronization and channel estimation procedure for the example of PAM block transmission based on a rectangular non-return-to-zero (NRZ) pulse shape. However, it should be noted that the same concept is equally suitable for DMT systems or single-subcarrier block-transmission including carrier less amplitude and phase modulation (CAP) [3].

An important application scenario for such optical systems is visible-light communication (VLC) for Li-Fi systems, where Li-Fi is used in analogy to Wi-Fi to address wireless optical systems. For such VLC systems, we favor PAM block transmission over DMT for several reasons [4]. Firstly, although classical orthogonal frequency division multiplexing (OFDM) is known as a very bandwidth efficient modulation scheme, it does not provide a bandwidth efficiency advantage over PAM in the IM/DD context, since the time domain signal needs to be real-valued. In other words, for non-coherent sources no quadrature up-conversion to a pair of sinusoidal optical carriers is possible. Therefore, an OFDM spectrum with a complex conjugated symmetry is required1. Secondly, PAM offers a much lower peak-to-average power ratio, which makes the LED driver considerably easier to realize. In particular, the M different power levels of a multilevel PAM signal can be realized by switching different numbers of LEDs on and off, since typical LED lamps consist of several LEDs. Similar statements are also valid for plastic optical fiber (POF) communications [5, 6].

The timing synchronization of block transmission systems, which employ a periodic cyclic prefix or unique word insertion, is often based on the auto-correlation of two (or more) consecutive identical parts of the preamble, where the majority of the approaches are based on the Schmidl-Cox algorithm [7]. Such auto-correlation based approaches can offer two main advantages. Since the auto-correlation can be computed iteratively, only one multiplication is required per clock cycle, whereas a cross-correlation over a length of \( N \) samples may require \( N \) multiplications per clock cycle. Furthermore, in RF-systems or coherent optical systems, the frequency offset of the down-converted complex signal may affect the cross-correlation process negatively [7].

However, optical systems with IM/DD do not suffer from such a carrier frequency offset, since only the intensity of a signal is information bearing. Moreover, if binary sequences are used for the preamble, the correlation process requires only additions or subtractions. This means that the corresponding circuit is multiplier-less and thus it is both implementation and energy efficient. Well known binary sequences with good correlation properties are Golay complementary sequences (CS), which are also used for the 60 GHz WLAN PHY [8], have even better correlation properties.

1 For such multiple subcarrier signals, we use the term DMT instead of OFDM.
In this work we will show that these Golay sequences are also ideal candidates for the block synchronization and channel estimation in optical systems with IM/DD. We will show that even short sequences ensure a reliable placement of the FFT-window. Based on this precise estimation of the cyclic prefix (or unique word) interval, which extends over $L$ symbols, the cross-correlation process provides a direct estimate of the channel impulse response of length $L$. This impulse response estimate, appended by a number of zeros, i.e., a completely noise free part, is then used to obtain an estimated channel transfer function via the FFT.

In following section, the preamble structure is discussed. Section 3 describes the correlation process, the timing synchronization and the channel estimation. In Section 4, the performance is analyzed, while Section 5 concludes the work.

2 COMPLEMENTARY SEQUENCES AND PREAMBLE STRUCTURE

CS have unique auto-correlation properties. The out-of-phase aperiodic and periodic auto-correlation coefficients of a CS pair sum up to zero.

Golay CS are binary and as such ideal candidates for the preamble of optical block transmission schemes. Golay CS with a length of a power of 2 can be encoded as Shapiro-Rudin polynomials [9]. According to the Shapiro-Rudin polynomials of degree 3, a Golay CS pair of length $N_{CS}$ is given as

$$A^\pm_4 = [a^\pm_1 \ldots a^\pm_4] = [1, 1 , 1 , -1]$$
$$B^\pm_4 = [b^\pm_1 \ldots b^\pm_4] = [1, 1 , -1 , 1],$$

where the CS pairs of a length $2N_{CS}$ can be obtained iteratively according to

$$A^\pm_{2N_{CS}} = [A^\pm_{N_{CS}} B^\pm_{N_{CS}}]$$
$$B^\pm_{2N_{CS}} = [A^\pm_{N_{CS}} -B^\pm_{N_{CS}}].$$

The index $\pm$ is used to emphasize that the elements are $\pm 1$. Clearly, for intensity modulation, the preamble comprises the unipolar equivalents $\bar{A} = [a_1 \ldots a_n \ldots a_{N_{CS}}]$ and $\bar{B} = [b_1 \ldots b_n \ldots b_{N_{CS}}]$ with $a_n \in \{0, 1\}$ and $b_n \in \{0, 1\}$.

For PAM-block transmission, we suggest a preamble structure as shown in Fig. 1. In order to imitate a periodic repetition of $\bar{A}$ ($\bar{B}$), we use a prefix containing the last $L$ elements of $\bar{A}$ ($\bar{B}$) and a suffix with the first $L$ samples of $\bar{A}$ ($\bar{B}$). This approach ensures that the sum of both correlation coefficients exhibits a main value which is framed by $L$ zeros in the noiseless case, where $L$ is the length of the cyclic prefix. Fig. 2(a) shows the basic principle and demonstrates that an ideal test signal for an impulse response measurement is available, if the length of the channel impulse response is not longer than $L$ pulses.

For a better DC-balance we suggest a further, simple manipulation of the preamble, which does not effect the desired correlation properties\(^2\), but which reduces the baseline wandering effect during the preamble. Firstly, instead of $\bar{B}$, the (logical) inverted sequence $\bar{B}$ is used, where the prefix and suffix sequences are still $\bar{B}_P$ and $\bar{B}_S$. Secondly, we invert the suffixes and prefixes of $\bar{A}$. So the prefix sequence for $\bar{A}$ is $\overline{\bar{A}}_P = [\bar{a}_NCS-L+1 \ldots \bar{a}_{NCS}]$, and the suffix sequence is $\overline{\bar{A}}_S = [\bar{a}_1 \ldots \bar{a}_L]$.

3 SYNCHRONIZATION AND CHANNEL ESTIMATION

Fig. 3 shows a block diagram of the receiver, where the components used for equalization and demodulation are not shown. With respect to the receiver, AC-coupling is assumed. We suppose that this coupling is dominated by single first order high-pass filter. This high-pass filter maps the unipolar input signal into a bipolar signal, which is required in order to exploit the correlation properties of the CS. To address the settling time of the high-pass, an additional training part (here: 1010...) is inserted at the beginning of the preamble, see Figure 1. This part can also be used for automatic gain control and packet detection. A 5th order Bessel filter is used for noise rejection, where the 3 dB cut-off frequency is chosen according to $1/(2T)$, where $T$ is the width of the transmitted rectangular NRZ pulses.

It is assumed that the ADC is clocked with a frequency $2/T$, thus fractional sampling is used for synchronization and channel estimation. However, it is assumed that this clock may exhibit a random (but constant) time offset $\delta T$, where $-1 \leq \delta \leq 1$.

\(^2\)Eq. (2) defines one method to obtain Golay CS. In [8], a different rule is described which may lead to an even better DC balance and further manipulations are not required.
The core of the receiver is the CS cross-correlator block, which consists of two multiplier-less matched filters. According to the clock, the FIR filter orders are $2N_{CS}$. The transfer functions are given as

$$
G_A(z) = a_{N_{CS}}^\pm z^0 + a_{N_{CS}-1}^\pm z^{-2} + \cdots + a_1^\pm z^{-2N_{CS}-1} \quad \text{and} \quad G_B(z) = b_{N_{CS}}^\pm z^0 + b_{N_{CS}-1}^\pm z^{-2} + \cdots + b_1^\pm z^{-2N_{CS}-1},
$$

where every second filter coefficient is equal to zero. Fig. 2(b) shows an example for the time discrete filter output signal. In this example (noiseless case), the correlator leads to a perfect estimation of the channel impulse response, even though the maximum of the impulse response is not sampled.

For the time synchronization, the correlator output signal is squared and further processed with a sliding window of length $2L$. This ensures that the part of the channel impulse response which exhibits the maximum energy is captured and makes the synchronization independent of the concrete shape of the channel impulse response.\(^3\)

Once the time synchronization is obtained, an estimation $\hat{g}[n]$ of the channel impulse with $2L$ fractionally spaced samples is available. For fractionally spaced frequency domain equalization with an FFT length $N_{FFT}$, an estimation $\hat{G}[\mu]$ of the transfer function will be obtained via a FFT of $\hat{g}[n]$, appended by a number of $N_{FFT} - 2L$ zeros. Since these zeros are not effected by noise, this leads to a strong noise reduction in the estimate $\hat{G}[\mu]$.\(^{10}\)

It is also possible to down-sample $\hat{g}[n]$ by a factor of 2 in order to use symbol spaced equalization. However, the fractionally spaced processing described above still ensures that the synchronization is accurate as $T/4$.

4 PERFORMANCE

Compared to a perfect channel knowledge, a channel estimation in AWGN (additive white Gaussian noise) may increase the bit error ratio (BER) at a given power level. Thus, in order to ensure that a certain BER limit $p_b$ will not be exceeded, an increased minimum power level at the receiver input is required. The receiver sensitivity is reduced. Fig. 4 shows the normalized required optical power for NRZ OOK-transmission, i.e., NRZ PAM with a modulation order of 2, and zero forcing frequency domain equalization. 7B8B coding is used to ensure a DC-balance at the

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\(^3\)In multipath channel, the first tap may not be strongest. So this solution is preferable compared to a fixed threshold.

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Fig. 3. Receiver block diagram (HP: high-pass, BF: Bessel filter)
transmitter\textsuperscript{4}. It is assumed that the photo diode output current is superimposed by signal independent AWGN with a power spectral density $N_0$.

With respect to the optical channel between the transmitter and receiver, we have distinguished between the ideal case with $G_{ch}(f) = 1$ and a channel with a Gaussian low-pass characteristic, where the transfer function is given as

$$G_{ch}(f) = e^{-\ln(2)(\frac{f}{f_{3dB}})^2}.$$\label{eq:gaussian_low_pass}

The parameter $f_{3dB}$ is the 3 dB cut-off frequency in the optical domain. The power levels shown in Fig. 4 are normalized with respect to the minimum required optical power

$$P_{ref} = \sqrt{\frac{N_0 R_b}{R}} \cdot \text{erfc}^{-1}(2p_b)$$

of uncoded NRZ-OOK transmission in a flat AWGN channel at the same values of $R_b$ (bit rate), $R$ (photo diode responsivity), $N_0$ and $p_b$, cf. [4].

The green bars correspond to perfect channel knowledge. For $G_{ch}(f) = 1$, the penalty of $\approx 1.3$ dB compared to the reference scheme is determined by three factors. Approximately 0.55 dB are caused by the redundancy of the 7B8B code and the cyclic prefix (here: 30 symbols for a block with 256 data symbols). Additional 0.25 dB are introduced by the Bessel filter, see Section 3. At $p_b = 10^{-3}$, the line code causes another 0.5 dB loss, since the decoding process leads to an error multiplication: on average, one bit error at the decoder input induces 3.6 bit errors at the output.

In order to evaluate the proposed synchronization and channel estimation method, a number of 200 independent realizations have been considered, where the sampling phase $\delta$, see Fig. 3, has been randomly chosen between -1 and 1. For the histograms, a CS with a length of $N_{CS} = 256$ has been used. The figure shows that the proposed concept leads to an excellent performance. For fractional equalization shown in Fig. 4(a), the maximum additional penalty is almost 0.5 dB, where the average additional loss does not exceed $0.3$ dB.

For symbol spaced equalization (see Section 3), a larger penalty compared to perfect channel knowledge may occur. However, the additional and still acceptable penalty is not a result of the algorithm — it is basically a result of a sampling theorem violation. The actual value depends mainly on the value of $\delta$. For a certain delta, the sampling time error may be as large as $\pm T/4$, if the timing circuit operates at a clock $2/T$. The corresponding energy loss cannot fully compensated, since the next samples used for equalization are separated by a time $T$. Clearly, if the timing circuit operates at a high clock, almost the same performance as for fractional equalization can be obtained.

5 CONCLUSION

In this work, we have shown that Golay complementary sequences are ideal candidates for the synchronization and channel estimation of optical block transmission systems with intensity modulation and direct detection. As these

\textsuperscript{4}Among all unipolar code words, all 8 bit code words with a weight of 4 are used. All code words with a weight of 5 are used, where the corresponding inverted codewords (code weight 3) are assigned according to the running disparity. 2 code words with a weight of 6 and the inverted versions are also used.
sequences are binary, multiplier-less matched filters / cross-correlators can be used at the receiver. Compared to auto-correlation based concepts, cross-correlation leads to a much more precise timing synchronization. Moreover, the ideal correlation properties of Golay complementary sequences also enable a direct estimation of the channel impulse response in the time domain, which in return offers a strong noise reduction in the estimated channel transfer function. We have demonstrated the performance of this Golay sequence based channel estimation concept for the example of binary PAM transmission with NRZ rectangular pulses, however, the basic concept is also applicable for DMT or single-subcarrier transmission. At the receiver, the transmitted unipolar binary preamble sequence is mapped into a bipolar sequence by means of a high-pass filter. To address the settling time of this filter, an additional training sequence part is required which is inserted prior to the Golay sequence pair. However, this part of the training sequence can also be used for automatic gain control and packet detection. For the timing synchronization in NRZ PAM systems, we have presented an approach with $T/2$ spaced sampling. If $T/2$ spaced equalization is additionally used at the receiver, the power penalty compared to perfect channel knowledge does not exceed 0.5 dB for Golay sequences with a length of 256. For symbol spaced equalization, the additional loss is larger, but can be reduced by a higher sampling rate of the timing circuit.

REFERENCES


