Low-Complexity Blind Timing Synchronization for ACO-OFDM-Based Optical Wireless Communications

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Abstract—In this paper, we present a novel technique for coarse timing synchronization for Asymmetrically-Clipped Optical Orthogonal Frequency Division Multiplexing (ACO-OFDM) systems. This technique is tailored specifically for the ACO-OFDM waveform, and performs coarse symbol timing synchronization without the need for training data. By using a specific symmetry property of the ACO-OFDM time-domain signal, the method can achieve accurate symbol timing synchronization with very low implementation complexity. Simulation results confirm this, and also show that the cyclic prefix-based correlation methods used commonly in traditional OFDM systems, such as 802.11x, cannot be applied to ACO-OFDM systems. The proposed method, therefore, provides an ideal replacement for the cyclic prefix-based method for future systems employing ACO-OFDM.

Keywords—ACO-OFDM, optical wireless communications, IM/DD, blind synchronization, channel impulse response.

I. INTRODUCTION

The growing demands of broadband applications and the current congestion in the RF spectrum are fueling interest in optical wireless communications (OWC). OWC is of particular interest because it offers the promise of higher data rates and license-free operation. Recent advances in optical device technologies have helped to reduce the costs and power demands of optical transceiver implementations, positioning OWC systems as complementary, and in some cases, viable alternatives to incumbent RF systems.

OFDM has emerged as the preferred modulation method for many broadband wireless communication systems. This is primarily due to its ability to combat narrowband interference and frequency selective fading, as well as its efficient hardware implementation [1]. OFDM is, therefore, a logical choice as the modulation scheme to use in OWC systems. However, differences with respect to RF devices and RF propagation characteristics prevent traditional OFDM techniques from being applied directly to OWC systems. In particular, in OWC systems, it is difficult to modulate and recover phase information [2]. Therefore, these systems typically use the intensity of the transmitted signal to communicate information. Systems employing this type of non-coherent modulation are referred to as intensity-modulated/direct detection (IM/DD) systems [3].

ACO-OFDM is a new modulation technique that adapts traditional OFDM for use in optical IM/DD systems. ACO-OFDM uses properties of the FFT to ensure a real and positive time domain signal. Despite the need to introduce redundancy in the frequency domain, ACO-OFDM has been shown to have a better power efficiency than any of the modulation schemes currently used for IM/DD systems, including on-off keying (OOK), pulse position modulation (PPM), and DC-offset optical OFDM (DCO-OFDM) [4]. This becomes an important factor in optical systems, where the average transmitted optical power is limited due to safety or other practical considerations. ACO-OFDM also retains the desirable properties of OFDM, such as resiliency to multipath fading, high spectral efficiency, and low implementation complexity [5].

It is well known that OFDM systems are highly sensitive to synchronization errors [6]. While techniques for synchronization in traditional OFDM systems are referenced extensively in literature, the topic of synchronization in ACO-OFDM is new. As was shown in [7], traditional synchronization techniques cannot be applied directly in ACO-OFDM due to the fundamental differences between the ACO-OFDM waveform and the traditional OFDM waveform. Also, while [7] presents a synchronization technique which is tailored to the ACO-OFDM system, it uses a training sequence to perform synchronization. A blind technique for synchronization is preferred, as it eliminates the bandwidth efficiency loss associated with a training sequence.

In this paper, we propose a new technique for synchronization that is tailored specifically for ACO-OFDM systems. The proposed synchronization scheme is a blind technique that exploits the redundancy inherent in the ACO-OFDM signal in order to achieve coarse symbol timing synchronization, while significantly reducing the overhead of a training sequence by replacing it with a single-symbol guard interval. Using this redundancy, a correlation function is defined whose minimum can be used, in conjunction with a specific refining technique, in order to locate the ACO-OFDM symbol timing at the receiver. The selected refining technique results in a non-data-aided (NDA) synchronization scheme which is shown to perform well for both additive white Gaussian noise (AWGN) and fading channel environments, even in low SNR conditions. The simulation results also show that the traditional blind synchronization method for OFDM [8]
cannot be applied in ACO-OFDM, making our method the only known viable blind approach for this OWC system. The remainder of this paper is organized as follows. In section II, the ACO-OFDM system is briefly described. Our novel technique is presented in section III, where the applicability of this technique to ACO-OFDM is further described. Section IV analyses the performance of the technique through simulation. Finally, some concluding remarks are made in section V.

II. ACO-OFDM SYSTEM

Fig. 1 shows a block-diagram representation of an ACO-OFDM system. The electrical domain of the ACO-OFDM system resembles that of a traditional OFDM system, except for the addition of certain key blocks (as highlighted in Fig. 1). The need for these blocks results from the fact that intensity modulation requires a positive and real-valued time-domain signal as an input. To ensure the signal is positive, the Clip operation clips the serialized IFFT output to zero. It can be shown that if prior to the IFFT, only the odd subcarriers are loaded with data and the even subcarriers are set to zero, the result of clipping is for the odd subcarrier amplitude to be reduced by half, and for the intermodulation distortion to fall only on the even subcarriers [5]. Therefore, to ensure that no distortion of the data subcarriers will be incurred as a result of clipping, the Even Sample Zero Insert block takes N/4 data subcarrier samples created by the QAM mapper, and inserts zero values at the even subcarrier positions. In order to ensure a real-valued time-domain signal is produced following the IFFT(N) operation, the frequency domain symbol is constrained to have Hermitian symmetry. This is accomplished by the Hermitian Symmetry block, which creates an N-point symbol from the N/2 samples output from the Even Sample Zero Insert block by inserting redundant samples such that \( X^H(N-K) = X^H(k) \). The 5th resulting time-domain symbol \( x^s(n) \) can be represented as:

\[
x^s(n) = \frac{1}{N} \sum_{k=0}^{N-1} X^H(k)e^{j2\pi n k / N}.
\]

Following the Clip operation, which ensures a positive time-domain signal, a cyclic prefix of length \( L \) is added to the time-domain symbol as in traditional OFDM. The sequence of \( L+N \) samples is then converted to an analog signal which is used to modulate the intensity of an optical signal.

The modulated optical signal is transmitted through an optical channel which can cause attenuation, introduction of background noise from other optical sources, dispersion from air particles and aerosols, and reflection from objects. A Direct Detection block at the receiver generally uses a photodetection device to convert the received light intensity into an electrical current. This photodetection operation also introduces a constant gain determined by the photodetector sensitivity. The received signal at the input of the Coarse Timing Synchronization Block is given by:

\[
y(m) = \sum_{j=0}^{L-1} h(j)x_c(m-j) + w(m)
\]

where \( x_c(m) \) is a unipolar signal due to the clipping operation, \( h(m) \) is the length \( v \) effective channel impulse response (CIR) which takes into account both the optical channel distortion and the photodetector gain, and \( w(m) \) is a real AWGN signal which is bipolar because it occurs in the electrical domain. It should be noted here that the variable \( m \) is introduced in the notation to indicate a signal with infinite time duration, in contrast to the variable \( n \) which can take values from 0 to \( N-1 \), and represents the sample index within an OFDM symbol. This notation also helps to distinguish a signal where the cyclic prefix has been inserted.

The task of the Course Timing Synchronization Block is to choose an FFT window which causes little to no inter-symbol interference (ISI). Namely, each of the samples \( y^m(n) \) within the chosen FFT window in (2) must have contributions from the transmitted signal that come only from the 5th OFDM symbol we would like to demodulate. From (2) it can be seen that the requirement on the coarse timing synchronization for no ISI is that [9]:

\[
L \geq v; \quad -(L-v) \leq \theta \leq 0
\]

where \( \theta \) is the offset in samples from the true FFT window, where a positive offset corresponds to a delay in time. Given the relationship in (3), the strategy often utilized in a coarse
timing synchronization technique is to estimate the timing of the FFT window with an algorithm which generates an error of, at most, $M$ samples, and then advance the timing by $M$ samples to ensure that (3) is satisfied. This assumes a channel impulse response that is limited in length so that $L \geq M + v$.

Assuming the ISI-free conditions of (3) are maintained, the samples $y_S(n)$ taken for the FFT window of the $S$th symbol can be expressed as the circular convolution of the transmitted symbol with a CIR:

$$y_S(n) = x_S^C(n) \otimes h(n) + w(n) \quad (4)$$

where the superscript $S$ denotes the samples obtained from the $S$th symbol (after discarding the cyclic prefix). As a result of (4), frequency domain equalization can be formed using a single tap representing the channel frequency response at each subcarrier frequency. This channel frequency response is estimated by the Channel Estimation and Compensation block, which accounts for the delay $\theta$ in the symbol timing as a complex phase term. Furthermore, if the coarse timing synchronization were to change the value of $\theta$ while maintaining ISI-free timing, a new channel estimate would be required due to the resulting change in the channel frequency response.

III. NOVEL BLIND TIMING SYNCHRONIZATION METHOD FOR ACO-OFDM

The timing synchronization method developed in this section makes use of time domain symmetry inherent in the ACO-OFDM signal. This symmetry results from the insertion of zeros at the even subcarriers, as discussed in section II. Due to this zeroing, (1) can be rewritten as:

$$x_S^C(n) = \frac{1}{N} \sum_{k=0}^{N-1} X_S^C(k) e^{j2\pi kn/N} \quad (5)$$

The contribution to $x_S^C(n)$ by the $k$th subcarrier observed at time $n+N/2$ is given by:

$$x_S^C(n + \frac{N}{2}, k) = \begin{cases} \frac{1}{N} X_S^C(k) e^{j2\pi kn/N}, & k, \text{odd} \\ 0, & k, \text{even} \end{cases} \quad (6)$$

If the contribution of each subcarrier is considered, then it can be seen that the time domain symbol $x_S^C(n)$ has a negative symmetry with respect to the midpoint of the symbol as shown in (7):

$$x_S^C(n + \frac{N}{2}) = -x_S^C(n) \quad (7)$$

Following the clipping operation in the transmitter, this symmetry property results in the following multiplicative rule, which holds for all values of $n$ for the $S$th symbol:

$$x_S^C(n + \frac{N}{2}) \cdot x_S^C(n) = 0; \forall n \in [0, N - 1] \quad (8)$$

The multiplicative property of (8) can be exploited to perform synchronization. At the receiver, a running correlation of length $N/2$ samples following the A/D block is maintained. Assuming an AWGN channel (the analysis for a fading channel is similar and yields the same conclusion), the correlation function can be expressed according to the offset $M$ from the actual frame boundary of the OFDM symbol as:

$$C(M) = \sum_{n=0}^{N/2-1} (x_c(n + M) + w(n + M))^* (x_c(n + M + \frac{N}{2}) + w(n + M + \frac{N}{2})) \quad (9)$$

If it is assumed that the OFDM symbol length $N$ is large enough, and that the noise samples $w(n)$ are independent zero-mean random variables, the correlation function can be approximated, using an expectation operator as follows:

$$C(M) \approx \sum_{n=0}^{N/2-1} \left( x_c(n + M) \cdot x_c(n + \frac{N}{2}) + \frac{N}{2} E\{x_c(n + M) \cdot w(n + \frac{N}{2}) + w(n + M) \cdot x_c(n + \frac{N}{2}) + w(n + M) \cdot w(n + \frac{N}{2})\} \right) \quad (10)$$

If we focus on the value of the approximate correlation function (denoted $\hat{C}$) at the actual symbol boundaries, i.e., $M = L + S\cdot N$ with $S$ being a non-negative integer, this function can be expressed in terms of the transmitted samples of the same OFDM symbol $S$:

$$\hat{C}(L + S\cdot N) = \sum_{n=0}^{N/2-1} x_c^S(n) \cdot x_c^S(n + \frac{N}{2}) = 0 \quad (11)$$

Because $x_c(n)$ is non-negative, $\hat{C}$ will have a local minimum at the actual symbol boundaries. The synchronization technique, therefore, searches for the minimum of the correlation function of (9) to establish the timing of the FFT window.

The ACO-OFDM Coarse Timing Synchronization block of Fig. 1 will, therefore, need to consist of two main stages: a running correlation function which correlates each block of $N/2$ samples of the received signal with the subsequent $N/2$ samples, and a minimization function that locates the minimum value of the running correlation function. This differs significantly from traditional blind timing synchronization techniques for traditional OFDM, where the block length is $L$.
samples, and where maximization of the correlation function is performed [8]. This illustrates a distinct advantage of the synchronization method proposed in this paper, in that the number of samples being used in the correlation function \((N/2)\) is significantly larger than that of the cyclic prefix-based method. As a result, the decision metric will be more robust to noise compared with the cyclic-prefix based method. Furthermore, the cyclic prefix-based method suffers from a severe limitation when applied directly to ACO-OFDM, since several of the samples of the cyclic prefix could theoretically be zeros. This results in a correlation function value at symbol boundary relative to the remaining symbol which can be significantly smaller than in the case of traditional OFDM, thus making timing synchronization more difficult. Maximization is, therefore, not a good strategy in the case of ACO-OFDM.

The local minimum value of \(\hat{C}\) will not only occur at the actual symbol boundary, but also at any point within the cyclic prefix of the ACO-OFDM symbol. This is due to the fact that the cyclic prefix is formed from the last \(L\) samples of the OFDM symbol, and the symmetry properties of the symbol that was shown by (8). Fig. 2 gives a simple example to illustrate this for an FFT size of \(N=8\) and cyclic prefix length of \(L=3\). Here, time \(t=0\) corresponds to the correlation function taken at the start of the cyclic prefix. Time \(t=3\) is the expected symbol timing location and corresponds to a local minimum of \(\hat{C}\) as concluded earlier. However, \(t = 0, 1, \) and \(2\) also yield local minima. In addition, this local minimum could be extended several samples after the symbol boundary, depending on the organization of positive and negative samples in the time domain data. In the example, if either \(x_C'(4)\) or \(x_C'(5)\) are zero, then the minimum extends for an extra sample beyond the symbol boundary. As we move further to the right of the symbol boundary, more terms will be added to the correlation sum, and the likelihood of having a minimum will sharply decrease.

Fig. 3 shows the simulated correlation function at the receiver for large SNR. Due to noise at the receiver and the approximation that was made in (10), there is a small variation in the correlation function at the points where a local

![Figure 2. Example Correlation Computation Demonstrating Duration of the Correlation Function Minimum](image)

minimum of \(\hat{C}\) is expected. Despite this, the actual correlation function shows a well-defined valley in the vicinity of the actual symbol boundary. This valley is composed of three regions: 1) the cyclic prefix, where the correlation is always zero in the absence of noise, 2) the leading zeros, which occur due to random zero elements in the correlation terms involving symbols \(S\) and \(S-1\), and 3) the trailing zeros, which occur due to random zero elements in the correlation terms involving symbols \(S\) and \(S+1\). Because this valley is greater in length than the cyclic prefix itself, a simple minimization function for the synchronization algorithm is not sufficient to reliably satisfy the requirements of (3) for the estimator.

In order to locate the actual symbol boundary within the correlation valley, a combination of two strategies is used. Firstly, a guard sample is inserted into the time domain signal. The guard consists of a single non-zero sample inserted at the beginning of the cyclic prefix in the time-domain signal. The value of this sample is chosen such that the average power of the transmitted signal is not affected. The presence of this non-zero sample is known at the receiver, where it is discarded along with the cyclic prefix. Since the timing at the receiver is assumed to fall within the cyclic prefix, insertion of this non-zero guard sample does not cause ISI. Furthermore, while the redundancy or loss in efficiency due to this sample is very small, it considerably reduces the likelihood and duration of the leading and trailing zeros in Fig. 3. In addition to insertion of a guard sample, a correlation function increase detection is used to detect a large increase in the correlation function. A threshold, \(\lambda\), is chosen based on the SNR observed. Once the correlation valley is detected, the change in the correlation function on a sample-by-sample basis is compared with the threshold \(\lambda\). If the correlation function increases by \(\lambda\) or more, the presence of a symbol boundary is assumed.

![Figure 3. General Behavior of the Correlation Function for the N/2 Sample Correlation Method](image)

The block diagram in Fig. 4 shows the Coarse Timing Synchro. Block of Fig. 1 as presented in this paper. It consists of an \(N/2\) Running Correlation Calculation, followed by a minimization of this function for locating the valley of the correlation function. The Minimum Detection function will observe the correlation over a period of \(N+L\) samples, and will locate the correlation function valley by finding the minimum point over that period. In the case of low SNR, the Minimum Detection function can use the correlation function.
over multiple symbol periods in order to ensure that a valley in the correlation is found correctly. Following this, a Correlation Function Increase Detection by $\lambda$ is used to select the estimated FFT window location. This estimate is then biased by a window advance of $M$ to ensure the estimated timing always falls within the cyclic prefix. The value of $M$ corresponds to the maximum error expected in determining the symbol timing.

### IV. SIMULATION RESULTS

Simulations were performed on an ACO-OFDM system with an FFT size of $N = 512$ and a cyclic prefix of length $L = 16$ samples. Each of the 128 data subcarriers was loaded with random BPSK data. In the time domain, a guard sample (as described in section III) was inserted just before the cyclic prefix of each symbol. Both AWGN and fading channels were used in the simulation. For the fading channel case, an exponentially decaying Rayleigh fading channel with 7 taps was used. The magnitude of the Rayleigh channel coefficients was used for the elements of the impulse response in order to obtain a real, positive impulse response. AWGN added at the receiver was also restricted to be real (positive or negative).

Fig. 5 and Fig. 6 show the output of the $N/2$ Correlation Computation block of Fig. 4 over several ACO-OFDM symbols with AWGN and Fading channels respectively. In each case, this output is compared with the correlation function that is computed in the Cyclic Prefix-Based Correlation Method [8], where a block of samples of length $L$ is correlated with an offset of $N$ samples in attempts to match the cyclic prefix with the time-domain data with which it is created. In both channel conditions, the $N/2$ correlation output generates a well-defined minimum in the vicinity of the actual symbol boundaries, thus demonstrating the robustness of this metric for determining symbol timing, and demonstrating its superiority compared to the traditional correlation maximization metric.

The Cyclic Prefix-Based Correlation Method, on the other hand, does not yield a well-defined peak at the actual symbol boundaries, as it does in traditional OFDM systems. Instead, several correlation peaks, randomly distributed within the symbol, can be observed, making timing synchronization using this metric impossible. This is due to the presence of zeros in the cyclic prefix that are inherent in the ACO-OFDM signal. Therefore, while the Cyclic Prefix-Based Correlation Method is adversely affected by these zeros, the $N/2$ Sample Correlation Method actually makes use of these zeros in performing symbol synchronization.

In Fig. 7, Fig. 8, and Fig. 9, the performance of the entire $N/2$ Sample Correlation Method illustrated in Fig. 4 is evaluated. A value of $\lambda = 0.001$ was used, as it showed optimal results in terms of the maximum observed timing error over the range of SNR (0dB to 30dB) considered. The value of $M$ was arbitrarily set to 0 in order to determine, through these same simulation results, what a practical value of $M$ would be in such a system.

Fig. 7 shows the probability of obtaining exact symbol timing for the $N/2$ Sample Correlation Method and for the Cyclic Prefix-based Method. The $N/2$ Sample Correlation Method again demonstrates much better performance than the traditional method.

In Fig. 8 and Fig. 9, the cumulative distribution function for the magnitude of the symbol timing error is shown for both channel conditions. The majority of the time, a very small error (2 samples or less) in timing is observed. Occasionally, larger errors up to 8 samples in magnitude are also observable. The observed errors are all in one direction (delay) with respect to the actual symbol timing. As a result, the figures below show that if $M=8$ is chosen as the final bias, the simulated ACO-OFDM system with a cyclic prefix of $L=16$ will adhere to (3) and, therefore, avoid ISI.
In this paper, a new method for blind timing synchronization was presented for ACO-OFDM systems. It was shown that traditional maximization of the cyclic-prefix-based correlation function is not suited for the ACO-OFDM signal structure. As a result, a method based on the minimization of the correlation between subsequent blocks of \(N/2\) samples in the received signal was derived. The proposed algorithm was simulated in a practical ACO-OFDM system. In addition to having considerably better performance compared with traditional cyclic prefixed-based blind synchronization methods [8], our NDA synchronization method benefits from relatively low implementation complexity and the absence of any training data in the transmitted symbols. As a result, the algorithm can be performed at any time during data transmission at the receiver in order to obtain proper timing, and it does not suffer from any loss in power efficiency due to training data. In addition, our proposed algorithm is capable of achieving reliable coarse timing synchronization at low SNR. These factors make this technique ideal for application in optical systems, where the transmitted power is limited by eye safety considerations. Finally, we demonstrated, for the simulated ACO-OFDM system, that a cyclic prefix of \(L=16\) is sufficient to avoid any ISI in the received signal caused by residual timing synchronization errors. The robustness of the algorithm was demonstrated for both AWGN and fading channel environments. Based on these promising results, a study of other aspects of an ACO-OFDM receiver, such as the channel estimation, would provide further insights into the overall system design.

**REFERENCES**


