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Block synchronization algorithms for UWB-OFDM systems

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ABSTRACT

Orthogonal Frequency-Division Multiplexing (OFDM) is a modulation known by its high spectral efficiency, high tolerance to multipath delay spread in frequency selective channels, and low cost of implementation. However, OFDM systems are extremely sensitive to synchronizations errors. In this paper, we propose an unbiased block synchronization algorithm that uses the structure of the cyclic prefix, the presence of pilot tones in the OFDM block, and an estimation of the channel time impulse response. This algorithm is able to achieve fast synchronization as required, for instance, on Ultra-Wide Band (UWB) multipath fading channels. To quantify the performance of the block synchronization algorithm, we use the start of frame mean square error. Through numerical simulations, we show that our algorithm outperforms previously published synchronization algorithms.

1. Introduction

Orthogonal Frequency-Division Multiplexing (OFDM) has been largely recognized as a communication system with high spectral efficiency and low algorithmic cost. As a consequence, OFDM is the preferred choice for communication systems that operate over fading channels subject to numerous disturbance sources. In particular, OFDM is a candidate for Ultra-Wide Band (UWB) wireless communications.

On the other hand, a successful implementation of an OFDM system requires a tight solution to different synchronization problems. In particular, local clock drift may generate a phase difference between transmitter and receiver. When this difference is an integer number of sampling intervals, the starting point of each OFDM block gets shifted. Erroneous block starting point may lead to inter-channel interference (ICI) or even inter-block interference (IBI) [1–4]. Therefore, the starting point for each OFDM block needs to be determined before processing it. This problem is particularly acute on wireless transmissions where the transmission channel varies over time.

In [5], a maximum likelihood problem is set to estimate the initial sample corresponding to the OFDM block. Even though the solution presented is very accurate, the numerical algorithm is complex and it leads to costly receiver implementations. An alternative solution for AWGN channels has been proposed in [6]. In particular, the authors formulate a simplified likelihood problem considering the structure of the cyclic prefix and the pilot tones embedded on the OFDM symbol. However, they disregard the effect of multipath channel, and hence the overall performance degrades noticeably when frequency selective channels are considered. In general, time dispersion due to multipath effects is a severe limitation for UWB systems. In [7], a novel OFDM system is designed for UWB communications to mitigate the multipath effect. On a different venue, the authors of [8] propose a low-complexity equalization technique for UWB systems to compensate for the channel distortion. The inter-symbol interference problem due to frequency selective dynamics has already been recognized by the IEEE standard committee when proposing the UWB standard IEEE 802.15 [9]. In particular, a cyclic prefix length of

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60.6 ns has been chosen to accommodate channels with delay spreads as long as several tens of nanoseconds. Clearly, this type of channel has a dynamic behavior that depends on the frequency.

Distortion properties of UWB propagation in real environments have been largely studied in the past. For instance, in [10], indoor UWB propagation channels with line of sight (LOS channels) and without it (NLOS channels) were modeled using experimental observations. Moreover, the results obtained after deep analysis of several measurement campaigns of UWB communication systems, show channel impulse responses (CIR) lasting between 200 ns and 400 ns due to multipath effects. A more recent study [11] has confirmed these observations for outdoor channels. In addition, the authors in [12] have proposed a statistical model considering the varying behavior when transmitting in different environments, such as residential indoor, office indoor, industrial indoor, farm environments, and body area networks. The results of the paper show that the effect of frequency dependence of channel gains is severe when communication along realistic environments, instead of free-space propagation, is considered.

Building upon all these observations and the fact that previous block synchronization techniques did not considered frequency selective effects, we propose to modify the likelihood function to include knowledge of the time impulse response of the transmission channel. The robustness of our algorithm with respect to estimation errors is tested through numerical simulations for several signal to noise ratios.

It is shown that the estimation problem is solved by maximizing the convolution of two autocorrelation functions. Following this remark, we suggest the best pilot tone locations to improve the estimation performance. To quantify our strategy, we compute the mean square error of the estimation for different signal to noise ratios. Finally, we compare the performance of the newly proposed algorithm with other algorithms, previously proposed in the literature [6,13].

The paper is organized as follows: in Section 2, we present the block synchronization problem; later, in Section 3 we develop the start of frame estimation algorithms and we propose the pilot tone location strategy; finally, in Section 4, we discuss the performance of algorithms using numerical simulations of a OFDM-UWB channel.

2. Block synchronization problem

Consider an OFDM system transmitting N independent sub-carriers over a finite impulse response (FIR) channel. The impulse response duration is τ_d seconds. Then, $N_h = \tau_d/T$ is the impulse response length, measured in sampling intervals. We assume that N_h is an integer number.

OFDM systems use a redundancy of ν samples in the form of a cyclic prefix (CP), i.e., the first ν samples of the symbol are a copy of the last ν samples of the same symbol. Thus, the OFDM symbol length is $N + \nu$. At the receiver, the first ν samples are discarded and an N-point FFT is performed with the remaining samples. When the transmission channel satisfies $N_h \leqslant \nu + 1$, and the phase error is null, then it is well known that OFDM communications are IBI and ICI free. However, due to channel variations and drifts in the local clocks this may not be always true and phase estimation should be performed regularly to avoid phase errors. When the phase difference accumulates to an integer number of samples, the starting point of each OFDM block gets shifted. If the FFT is performed without correcting the shift, it is possible to prove that on the worst case, the symbol received on the k-th sub-carrier results as [14]

$$Y_k = \alpha_{e_\theta} X_k H_k e^{\jmath \frac{2\pi}{N} k e_\theta} + W_{e_\theta,k} + W_k,$$

where X_k are the transmitted symbols, H_k is the complex channel gain, W_k is complex Gaussian white noise, and e_θ is the phase block error in samples. It is clear that the information transmitted through the k-th sub-carrier suffers a phase shift proportional to e_θ, it is attenuated by α_{e_θ}, and it is perturbed by a crossover term $W_{e_\theta,k}$ due to ICI.

Moreover, we are interested in solving this problem for UWB systems that run with high clock rates and therefore, require fast synchronization algorithms. UWB wireless channels are time-varying in nature. Different measurement campaigns have shown that coherence time for UWB systems is about $T_c \sim 200$ µs. Thus, the start of frame estimation must be performed on a time interval much smaller than T_c because after this time the channel state, and the channel delay, may be different and it is necessary to refresh the synchronization algorithm. These are the restrictions that we consider in the analysis of the block synchronization problem performed in the following sections.

3. Estimation algorithms

Most estimation algorithms proposed in the literature have been developed for ideal AWGN channels. However, the performance of these algorithms decay noticeably when they are used on frequency selective channels [4,6,13,15,16]. On the other hand, the algorithms that consider the effect of a real channel [5] are computationally expensive and they cannot be employed on UWB channels.

Consider the following base-band model for the AWGN channel

$$y_n = s_{n-\theta} + p_{n-\theta} + w_n, \tag{1}$$

where y_n is the received signal, s_n and p_n are the data-bearing and the pilot signals respectively, θ is the channel delay and w_n is the noise term. We will consider that the noise has a zero-mean circular Gaussian distribution with variance σ_w^2.

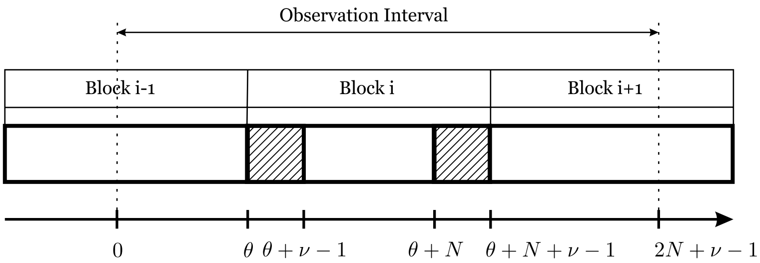


Fig. 1. Block synchronization problem for OFDM systems.

We assume that N_p sub-carriers out of the N OFDM sub-carriers are pilot tones. Let $\mathcal{P}$ be the set formed by the pilot tone indexes, and P_k be the known symbol sent through the k-th pilot tone. Then, the pilot signal is as follows:

$$p_n = \frac{1}{\sqrt{N}} \sum_{k \in \mathcal{P}} P_k e^{\jmath \frac{2\pi}{N} kn}, \quad n = 0, \ldots, N-1. \tag{2}$$

Now, let us consider the data-bearing signal. The symbol sent through the k-th sub-channel is X_k. Then,

$$s_n = \frac{1}{\sqrt{N}} \sum_{k=\{0,\ldots,N-1\},\, k \notin \mathcal{P}} X_k e^{\jmath \frac{2\pi}{N} kn}, \quad n = 0, \ldots, N-1. \tag{3}$$

Since s_n is the sum of independent random variables X_k of similar energy $\mathcal{E}_x(k)$, it may be approximated by a Gaussian process. We assume that the transmitted energy $\mathcal{E}_x(k)$ can be different across sub-carriers because a loading algorithm could be used. The variance of s_n is $\mathcal{E}_s = \alpha\sigma_x^2$, where $\alpha = (N - N_p)/N$ and σ_x^2 is the mean transmitted energy, i.e.,

$$\sigma_x^2 = \frac{1}{N - N_p} \sum_{k=\{0,\ldots,N-1\},\, k \notin \mathcal{P}} \mathcal{E}_x(k).$$

We assume that the pilot symbols have the same average energy as the data-bearing ones, i.e., $E[|P_k|^2] = \sigma_x^2$. Finally, y_n in (1) is approximated by a Gaussian process with time varying mean given by the sequence p_n.

Now, consider a block that leaves the transmitter at $n = 0$ and it arrives at the receiver θ samples later. We define two set of indexes:

$$\mathcal{I} = \{\theta, \ldots, \theta + \nu - 1\}, \qquad \mathcal{I}' = \{\theta + N, \ldots, \theta + N + \nu - 1\}. \tag{4}$$

We see that $\mathcal{I}$ contains the indexes corresponding to the cyclic prefix and $\mathcal{I}'$ contains the indexes of the last ν samples of the OFDM symbol. These are the samples that are copied onto the cyclic prefix. We will consider a block of $2N + \nu$ samples to estimate the starting point of the OFDM block, as shown in Fig. 1.

Following [6], we compute the likelihood function of θ given the sequence y_n, $n \in [0, 2N + \nu - 1]$. Let $\Lambda_1(\theta)$ be such function. Then, the estimation is computed as

$$\hat{\theta}_{\mathrm{MLE1}} = \arg\max_{\theta} \Lambda_1(\theta), \tag{5}$$

where

$$\Lambda_1(\theta) = \rho\Lambda_{cp}(\theta) + (1-\rho)\Lambda_p(\theta), \tag{6}$$

$$\Lambda_{cp}(\theta) = \Re\left\{ \sum_{n=\theta}^{\theta+\nu-1} y_n y_{n+N}^* \right\} - \frac{\rho}{2} \sum_{n=\theta}^{\theta+\nu-1} |y_n|^2 + |y_{n+N}|^2, \tag{7}$$

$$\Lambda_p(\theta) = (1+\rho)\Re\left\{ \sum_{n=0}^{2N+\nu-1} y_n p_{n-\theta}^* \right\} - \rho\Re\left\{ \sum_{n=\theta}^{\theta+\nu-1} (y_n + y_{n+N}) p_{n-\theta}^* \right\}, \tag{8}$$

and

$$\rho = \frac{\alpha \mathrm{SNR}}{\alpha \mathrm{SNR} + 1}, \qquad \mathrm{SNR} = \sigma_x^2/\sigma_w^2. \tag{9}$$

The first term in (6), $\Lambda_{cp}(\theta)$, exploits the correlation among the samples of the OFDM block due to the cyclic prefix [13]. For high SNR, it also takes into account the energy of the received signal. In the limit when $\mathrm{SNR} \to \infty$ then $\rho = 1$, the estimator MLE1 simply minimizes the square of the absolute value of the difference between two complex samples, N instants apart

$$\hat{\theta}_{\text{MLE1}} = \arg\min_{\theta} \sum_{n=\theta}^{\theta+\nu-1} |y_n - y_{n+N}|^2. \tag{10}$$

When SNR is small, $\rho \to 0$, then the estimation $\hat{\theta}_{\text{MLE1}}$ is performed with the pilot tones. In particular, $\Lambda_p(\theta)$ behaves like $\Re\{\sum_{n=0}^{2N+\nu-1} y_n p_{n-\theta}^*\}$. Notice that this term represents the output of the matched filter for p_n when the input is y_n.

Although ρ was defined by (9), it may also be considered as a design parameter. Then, its use will determine the designer confidence on Λ_{cp} and Λ_p. On the sequel, we use this approach and we consider ρ as a fixed parameter that has no direct dependency with SNR.

The estimator (5) has a good performance on AWGN channels. However, frequency selective channels introduce a correlation in the signal p_n that is not taken into account in $\Lambda_p(\theta)$. In this case, the first term of $\Lambda_p(\theta)$ is no longer a matched filter and the performance of the estimator is deteriorated. Now, the base-band model for dispersive channel is,

$$y_n = (s_{n-\theta} + p_{n-\theta}) * h_n + w_n, \tag{11}$$

where h_n is the time impulse response of the channel. Suppose that an estimation of the transmission channel, is available. Let $\hat{h}_n$ be such estimation, and define the following sequence

$$v_n = p_n * \hat{h}_n. \tag{12}$$

Now, by replacing p_n in (8) by v_n, we define a new estimator, MLE2

$$\hat{\theta}_{\text{MLE2}} = \arg\max_{\theta} \Lambda_2(\theta), \tag{13}$$

where

$$\Lambda_2(\theta) = \rho\Lambda_{cp}(\theta) + (1-\rho)\bar{\Lambda}_p(\theta), \tag{14}$$

and $\bar{\Lambda}_p(\theta)$ is given by

$$\bar{\Lambda}_p(\theta) = (1+\rho)\Re\left\{\sum_{n=0}^{2N+\nu-1} y_n v_{n-\theta}^*\right\} - \rho\left\{\sum_{n=\theta}^{\theta+\nu-1} (y_n + y_{n+N}) v_{n-\theta}^*\right\}. \tag{15}$$

Alternatively, we define a computationally efficient estimator as follows,

$$\hat{\theta}_{\text{MLE3}} = \arg\max_{\theta} \Lambda_3(\theta), \tag{16}$$

where

$$\Lambda_3(\theta) = \Re\left\{\sum_{n=0}^{2N+\nu-1} y_n v_{n-\theta}^*\right\}. \tag{17}$$

To obtain (17), we have considered the term related to the matched filter only and discarded all other terms in (14). We differ the analysis of performance of (13) and (16) for next section.

Now, consider the expected value of (17) when $\hat{\theta}$ is the estimation of θ. Assume that h_n is perfectly known. Then, v_n is a deterministic signal equal to $p_n * h_n$. Since the data X_k and the noise w_k are zero-mean random variables, we obtain

$$\begin{aligned} E\left[\sum_n y_n v_{n-\hat{\theta}}^*\right] &= \sum_n v_{n-\theta} v_{n-\hat{\theta}}^* \\ &= R_v(\theta - \hat{\theta}) \\ &= R_p(\theta - \hat{\theta}) * Q_h(\theta - \hat{\theta}), \end{aligned} \tag{18}$$

where $R_v(\tau) = v_\tau * v_{-\tau}^*$, $R_p(\tau) = p_\tau * p_{-\tau}^*$ and $Q_h(\tau) = h_\tau * h_{-\tau}^*$ are the deterministic autocorrelations functions of the matched filter impulse response, the pilot sequence and the channel respectively. Ideally, we would like $R_v(\tau)$ to have a clear maximum at the origin to help the numerical evaluation of (13) or (16). We can contribute to this goal by shaping $R_v(\tau)$ through a careful selection of the pilot tone sequence. In particular, we want to select a sequence whose autocorrelation function enhances the maximum of (18).

Now suppose that the pilot tones are equidistant, i.e., there is a pilot tone every L sub-carriers. Then $\mathcal{P}$ is the set

$$\mathcal{P} = \{k = k_0 + mL,\ m = 0, \ldots, N_p - 1\}, \tag{19}$$

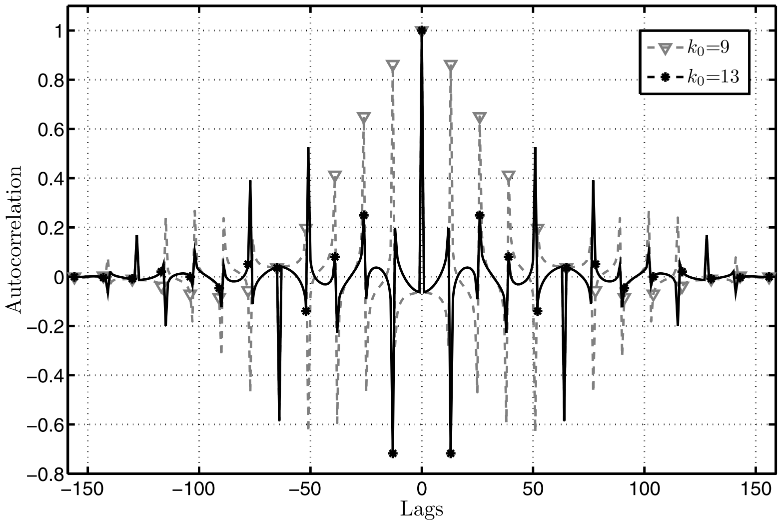


Fig. 2. Pilot signal autocorrelation for $k_0 = 9$ (UWB standard) and $k_0 = 13$ (best).

where k_0 is the index of the first pilot tone. On a first approximation, we will consider the pilot signal as the sum of sinusoidal signals. Then, the autocorrelation of p_n is

$$R_p(\tau) = e^{\jmath \frac{2\pi}{N}(k_0 + L\frac{N_p-1}{2})\tau} \frac{\sin(\frac{\pi}{N} L N_p \tau)}{\sin(\frac{\pi}{N} L \tau)}. \tag{20}$$

The periodic sinc function in (20) has its peaks spaced by N/L samples. The UWB standard fixes N and N_p, so N/L is determined and the sinc peaks cannot be separated more than that N/L. However, it is possible to select k_0 properly to use the phase factor in (20) to reduce or to cancel some of these peaks in the correlation function. In this way, we improve our block synchronization algorithm in (15) and (17) avoiding peaks in the likelihood function that may induce incorrect channel delay estimations.

4. Numerical results

In this section, we compare the performance of the block synchronization algorithms MLE1, MLE2 and MLE3 through numerical simulations over frequency selective channels using the base-band model.

In particular, we use a UWB channel that satisfies the Wireless Personal Area (WPAN) standard UWB [9]. Such system uses $N = 128$ carriers, $N_p = 12$ pilots, a cyclic prefix with $T_g = 60.6$ ns (32 samples) and the symbol period is $T_s = 303$ ns (160 samples). The OFDM signal bandwidth is 528 MHz, which implies that each individual sub-channel has a bandwidth of $1/T_u = 4.125$ MHz and the sampling period is $T_m = T_u/N = 1.894$ ns. The frequency band used in this case, which has its central frequency at $f_c = 3432$ MHz, is the first one out of the 14 proposed in the standard.

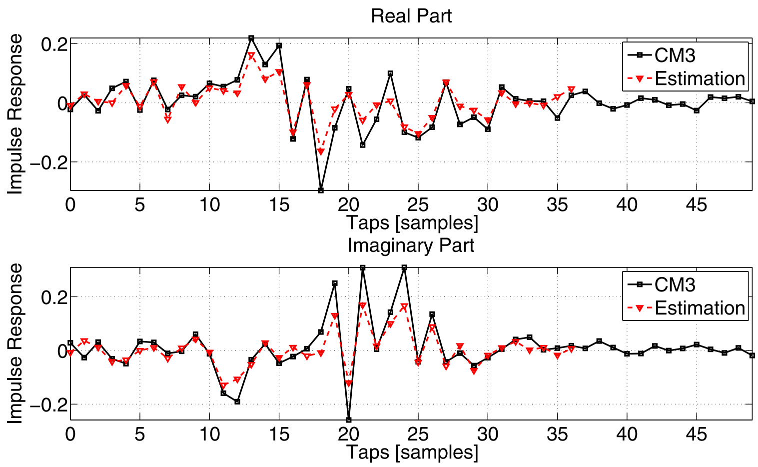
Taking into account the analysis respect to (20), we have found that $k_0 = 13$ minimizes the real part of the correlation peaks (here we have assumed that the first OFDM subcarrier has index 0). Fig. 2 shows the pilot signal autocorrelation function for distinct values of k_0. As we can see, a proper choice of the pilot signal decreases noticeable the peaks away from lag 0 of the autocorrelation function, therefore minimizing the start of frame miss-detection.

The UWB channel is modified version of Saleh–Valenzuela model [17] proposed by the IEEE 802.15.SG3 channel modeling sub-committee [18]. We have generated several realizations of the channel and normalized the channel time impulse responses so that the 2 norm of the channel is 1 on average. We have used a NLOS channel with r.m.s. delay spread $\tau_{rms} = 14.3$ ns that corresponds to the channel mode CM3.

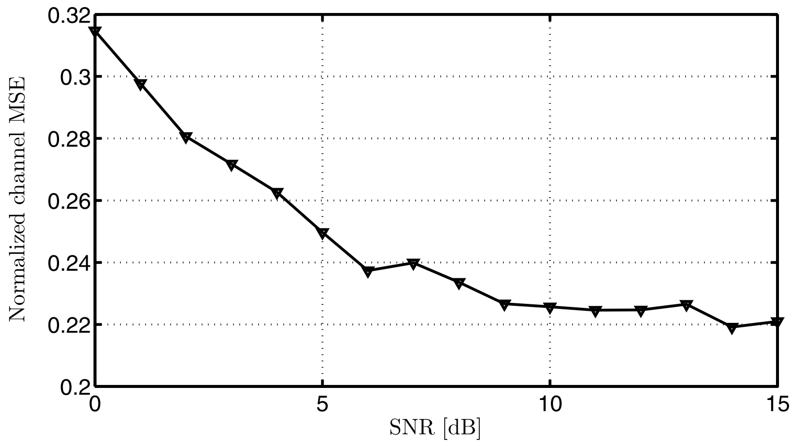
For obtaining $\hat{h}_n$, we have used a simple algorithm assuming one known OFDM symbol and computing the cross-correlation function between the transmitted symbol and the received sequence. Fig. 3 shows the realization of a channel impulse response, its estimation, and the normalized mean square identification error.

The parameter ρ weights the likelihood functions Λ_{cp} and Λ_p. In this paper, we selected ρ in order to minimize the MSE_θ for a fixed SNR (5 dB). We have found that this choice also works properly with high SNR as well.

We will consider the performance of the synchronization algorithm over $\gamma = 10^4$ different realizations of the transmission channel. For each realization, $\xi = 20$ independent synchronization processes are performed. For simplicity, we assume that the actual beginning of the OFDM block θ_0 remains unchanged. Let MSE_θ be the mean squared error between the θ_0 and $\hat{\theta}$ defined as



(a)



(b)

Fig. 3. (a) A UWB channel impulse response and its estimation for SNR = 5 dB. (b) Channel estimation mean square error as a function of SNR.

$$\mathrm{MSE}_\theta = \frac{1}{\gamma\xi}\sum_{i=1}^{\gamma\xi}(\hat{\theta}_i - \theta_0)^2. \tag{21}$$

Fig. 4 shows the curves MSE_θ vs SNR obtained with the three estimators MLE1, MLE2, and MLE3, using the CM3 UWB channel for $k_0 = 9, 13$. The performance of estimation MLE1 is not affected by the position of the pilot sequence. On the other hand, MLE2 and MLE3 have similar performance when $k_0 = 9$. However, by setting $k_0 = 13$, we improve significantly the performance of both MLE2 and MLE3. The performance of MLE3 is worse than MLE2 for high SNR, but its implementation is simpler, so it might be preferred when computational cost is at consideration.

To assess the impact of the identification errors on the performance studied above, we compare the curves of MSE_θ versus SNR using perfect knowledge of h_n and using $\hat{h}_n$. This is shown in Fig. 5.

As expected, we obtain the best performance when h_n is perfectly known. However, using a rough estimation $\hat{h}_n$ as the one obtained in Fig. 3, we are still able to achieve acceptable performance. In the worst case, for low signal to noise ratios, the maximum synchronization error using $\hat{h}_n$ is only 2.7 times the error using perfect channel knowledge.

Fig. 6 shows the normalized histogram obtained for each start of frame estimation $\hat{\theta}$ in 2×10^5 data symbols transmitted. Notice that MLE2 and MLE3 have sharp peaks at the correct position, while MLE1 has a broad bell near θ_0. It is also shown on Fig. 6 that the MLE1 estimation is slightly biased.

In order to quantify the histograms we define the probability of error for an incorrect estimation of the start of frame, i.e.,

$$P_e = P(\hat{\theta} \neq \theta_0).$$

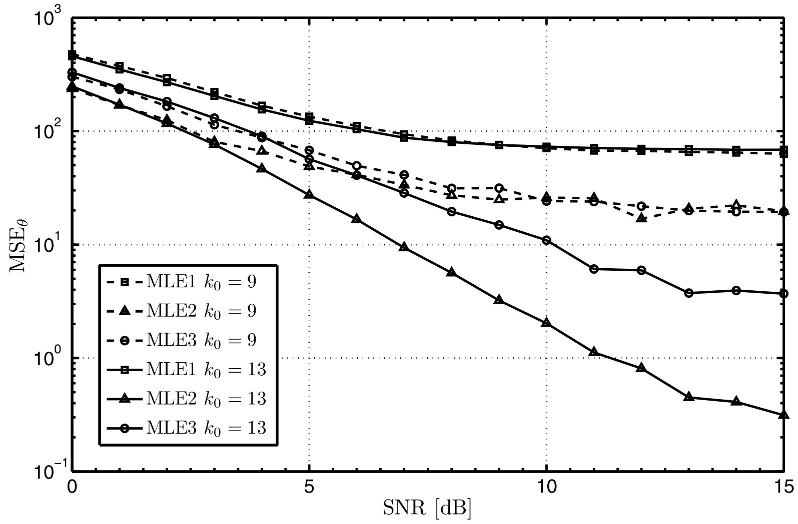


Fig. 4. Start of frame mean squared error as a function of the signal to noise ratio.

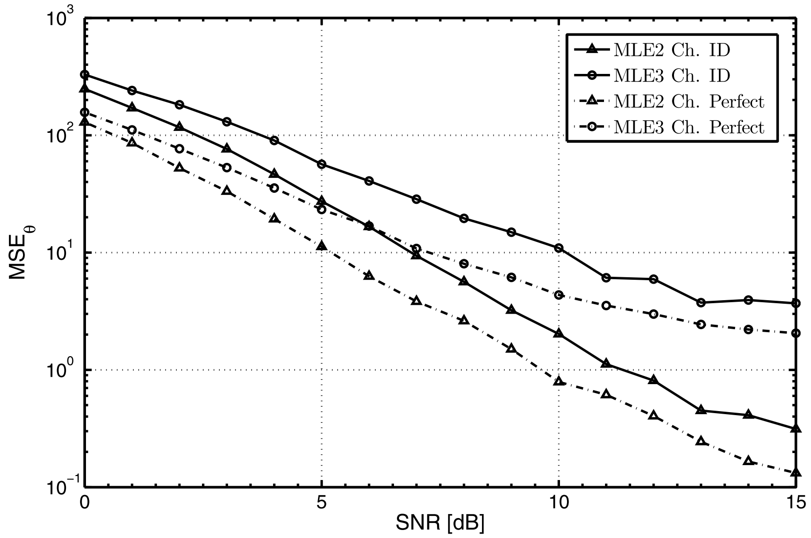


Fig. 5. Start of frame mean squared error as a function of the signal to noise ratio. Comparison when h_n is perfectly known and when it is estimated by $\hat{h}_n$ for $k_0 = 13$.

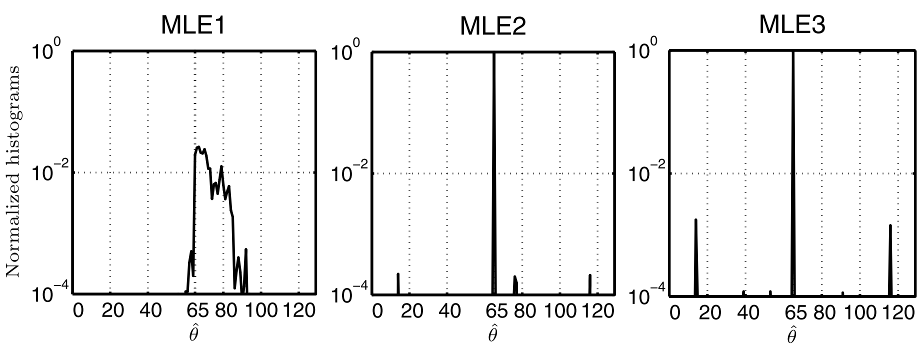


Fig. 6. Comparison of the histograms obtained with each estimation method when SNR = 10 dB. The correct start of frame position is $\theta_0 = 65$.

Table 1
Start of frame error probability P_e for SNR = 10 dB and $k_0 = 13$.

Estimator	MLE1	MLE2	MLE3
P_e	0.98	1.9×10^{-3}	5.2×10^{-3}

In Table 1 we show those quantities for the three estimators when $k_0 = 13$. Again, the best performance is achieved with the MLE2 estimator.

5. Conclusions

We have proposed two OFDM block synchronization algorithms. Both use the cyclic prefix structure, the pilot tones included into the block and an estimated time impulse response for the transmission channel. The improvement observed on the estimation performance of both schemes is due to the incorporation of the channel impulse response into the correlation of the pilot sequence.

Although our approach requires a first channel identification step, we have also shown that a rough channel estimation achieves a performance degradation of less than 50%. It is clear that the channel identification error is improved with proper phase synchronization. Therefore, this approach may be used in an alternate scheme where channel identification and synchronization are combined to obtain good start of block estimations.

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