Blind Algorithms for Timing Acquisition of Impulse Radio UWB Communications

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Abstract—This paper deals with timing recovery for Impulse Radio (IR) ultra-wideband (UWB) communication systems, which is known to be a challenging task to accomplish. We focus on blind estimation approaches, with the aim of being able to operate with minimal constraints on the signal format (no use of pilot symbols). We consider also the blind timing estimation problem in the sense that no prior knowledge is needed on the modulation parameters. The proposed methods may contribute to develop flexible receivers, capable of synchronizing various IR-UWB signal formats from short data records, even if some key parameters are changed over time.


I. INTRODUCTION

Since its approval by the U.S. Federal Communications Commission (FCC) in 2002, Impulse Radio - UWB (IR-UWB) has given rise to considerable interest in wireless communications research community, due to its many attractive properties [1]. Today, IR is considered as a promising technique from an industrial perspective [2]; in particular, it is a main candidate solution for applications such as wireless sensor networks [3] due to its ability to provide joint data transmission and precise positioning [4] at short distances. However, a number of system design challenges remain to be solved to ensure a broader use of IR-UWB communications in practice [5], due to the very fine time resolution and low power of transmitted pulses: waveforms distortions (due to transmitting/receiving antenna), clock jitter and time-variant frequency-selective channel with rich multipath. Whatever the receiver architecture, a synchronization block must provide accurate information on the arrival times of the incoming pulses, which can be difficult to achieve due to the aforementioned characteristics of IR-UWB. As shown in [6], [7] timing errors as small as fractions of a nanosecond can seriously degrade the system performance. Synchronization of UWB systems has therefore received considerable attention over the past ten years; an overview of the various strategies is given in [8].

Timing recovery is typically performed in two stages: a coarse synchronization is first carried out to quickly identify the symbol starting frame (acquisition\textsuperscript{1} stage); then, a tracking stage aims at refining the initial estimate and maintaining the timing error below the chip duration. Various approaches for UWB signal acquisition can be categorized into two main groups: detection based methods and estimation based methods. The basic principle used in the first group is to perform a correlation between the received waveform and a locally generated template of the transmitted pulse, delayed by a candidate phase (evaluated in a serial, parallel, or hybrid manner), followed by a threshold comparison [9]. The requirement of synthesizing a template waveform for exploiting phase information is typical of coherent receivers. In some situations, it may be better to explore the timing uncertainty region following a “look-and-jump” principle rather than linear search [10]. Search space reduction techniques have also been proposed to quickly identify the subset of the search space in which the true phase of the received signal lies [11]. For the estimation based methods, the timing offset is typically obtained by maximizing a statistic over a set of candidate phases. This statistic is usually obtained from correlation of the received signal with a template signal and no threshold comparison is involved. A first way to proceed is to take advantage of the cyclostationarity associated to the frame repetition pattern of impulse radio signaling [12]. A Maximum Likelihood (ML) timing estimator is derived in [14] by considering a rake structure and assuming the pulse waveform to be known at the receiver, which may be unrealistic. The channel impulse response is actually required at the receiver to generate the optimal template. Therefore, some works have been devoted to joint channel estimation and synchronization based on least squares [15] or transform-domain techniques [16]. However, a precise characterization of the channel response may be difficult to achieve due to the large number of paths. Hence, other approaches have been developed with no particular assumption on the received template waveform (this is the fundamental characteristic of any non-coherent receiver): pieces of the received signal (dirty template) are used as correlation pattern in [17][19] and in [20] the timing estimation problem is analysed under the unconditional ML criterion. Acquisition can also be performed from energy measurements and a judiciously designed signal structure at the transmitter [21], [22] or by modifying the polarity of the pulses using carefully designed binary codes [23]. The unique features of IR-UWB signals make the energy detection based non-coherent receiver [24] the preferred solution for most applications, as coherent receivers lead to high computational

\textsuperscript{1}Two parts are usually distinguished in this process, the frame timing and the symbol timing.
Most of the synchronizations methods rely on pilot symbols or training sequences to overcome the rich multipath diversity in UWB systems; some of the methods can operate in blind mode, in the sense that they do not require any pilot symbol (NonData-Aided, NDA) to work hence enabling a maximization of the effective transmitted throughput and a minimization of the mean transmitted power. The term blind can translates into a slightly different problem statement depending on the application objectives. In the field of electronic warfare (EW), blind estimation is a key process for passive listening [26][27], with the objective of detecting various technical characteristics of the observed signal with no prior knowledge. Similar blind techniques are also developed for spectrum sensing in the field of cognitive radio, with the aim of opportunistically providing wireless links that best meet the user communications requirements [25].

A few recent works on blind UWB timing estimation co-authored by the present authors are reviewed in this paper. In a first part of the paper, we focus on the estimation of the chip length of a time-hopping IR-UWB signal from sparse, noisy timing measurements, which is closely related to the estimation of the pulse repetition interval of a radar pulse train [28] from time-of-arrival (ToA) data. We present a novel round-based nonlinear cost function combining various incomplete observation sets [29]; thanks to a highly oscillatory behavior of this function, it is shown how the chip duration can be estimated in a blind manner. This approach has the advantage of providing performance bounds for the chip time estimate; however, its main drawback is that it relies on an ideal model for timing data. Hence, the channel characteristics are not really taken into account. So, we then study the problem of joint ToA and chip length estimation owing to the quantification of temporal hops being controlled by the chip duration $T_c$ with a total number $N_f$ of frames per symbol; $p(t)$ denotes the elementary pulse (also known as the monocycle) of ultrashort duration $T_p < T_c$.

If PSM is employed, the transmitted signal is

$$s(t) = \sum_n p_{T,d(n)} (t - n T_s),$$

with each information-bearing symbol $d(n) \in \{0, 1, ..., M - 1\}$ being conveyed using one particular orthogonal pulse from the set $S = \{p_0(t), ..., p_{M-1}(t)\}$, as

$$p_{T,d(n)} = \sum_{j=0}^{N_f-1} p_d(n) (t - j T_f - c_j T_c)$$

The transmitted signal $s(t)$ propagates through a fading channel which can typically be described by a tapped-delay line model with an impulse response $h(t) = \sum_{\ell=0}^{\lambda_{l \delta}(t - \tau_{\ell})}$, where $\{\lambda_{l \delta}, \lambda_{\ell 1}\} \in [0, N_c - 1]$ and $\delta$ is the time-hopping sequence, the quantification of temporal hops being controlled by the chip duration $T_c$, with a total number $N_f$ of frames per symbol; $p(t)$ denotes the elementary pulse (also known as the monocycle) of ultrashort duration $T_p < T_c$.

The symbol-long waveform $p_T(t)$ taking the form

$$s(t) = \sum_{n} d(n) p_T(t - n T_s),$$

with $d(n) \in \{-1, 1\}$, $T_f$ stands for the frame time and $\{c_j\}_{j=0}^{N_f-1} \in [0, N_c - 1]$ represent the time-hopping sequence. Hence, in case of PAM, the received signal can be written as (a similar expression is obtained for PSM)

$$r(t) = \sum_{n} d(n) p_R(t - n T_s - \tau_0) + \eta(t)$$

where $p_R(t)$ stands for the received symbol waveform, i.e. $p_R(t) = p_T(t) * h(t) = \sum_{\ell=0}^{\lambda_{\ell 1}} \lambda_{\ell 1} p_T(t - \tau_{\ell 0})$ where $\tau_{\ell 0} = \tau_{\ell 1}$ being the relative path delay and $\eta(t)$ accounts for both ambient noise and multiple-access interference.

In the sequel, our aim is to estimate the timing offset $\tau_0$ with no prior knowledge on the transmitted symbols nor on the channel state. We also investigate the possibility to achieve timing acquisition in the absence of knowledge of some key parameters of the modulation scheme ($T_c, T_f, ...$).
III. TIMING ACQUISITION WITH NO PRIOR KNOWLEDGE

A. Chip time estimation via modeling of ToA

In this section, the issue of blind timing acquisition of an IR-UWB signal is treated in such a way that no prior knowledge is required. As pointed out in a recent paper [29], this problem can be solved in the same way as pulse repetition interval (PRI) estimation of a radar pulse train from ToA measurements. Many methods have been proposed in the literature on this subject based on histogramming, Kalman filtering, Euclidean algorithm, periodogram or function optimization. The standard model which is taken into consideration for a series of sparse and noisy discrete events, arising from a periodic process, relies on a set of random variables \( Y = \{y_j\}_{j=1,2,\ldots,n'} \), being expressed as \( y_j = k_j T + \tau_0 + \eta_j \), where \( T > 0 \) is the unknown period, indices \( k_j \in \mathbb{N}^* \) specify the events that have been observed, and the elements \( \eta_j \) characterize the measurement noise\(^3\), here being considered as identically distributed, zero-mean Gaussian random variables with standard deviation \( \sigma_\eta \). In case of a IR-UWB, \( T_c \) can be recovered from such model once pulses ToA have been collected, through energy measurements and thresholding. Period \( T \) then corresponds to the chip time \( T_c \) and the coefficients \( k_j \) are described through the generic model \( k_j = j N_c + c_j \), where \( c_j \) are the pseudo-random code elements taking integer values in \([0, N_c - 1]\). To better reflect the unknown transmitter parameters and the severe propagation conditions (background noise, multipath), the data set \( Y \) can be modified according to given proportion(s) of outliers or/and missing observations.

In absence of prior knowledge, the problem of timing acquisition can be decomposed in two steps, the first one focusing on the chip time estimation from ToA differences (hence the unknown phase \( \tau_0 \) is removed), the second one being devoted to the search of the initial time offset (frame/symbol timing).

We consider first step as being the main issue in a fully blind approach (for the cases of PAM or PSM modulations), as \( T_c \) is a fundamental parameter of the process controlling the position of the pulses in time and other parameters such as \( T_f \) or \( T_s \) being directly related to \( T_c \) (at least for short TH codes). As a result, a “conventional” acquisition method based on dirty templates [17] may yield to frame/symbol timing once an estimate of \( T_c \) is available. Therefore, we focus only on chip time estimation in this section.

We introduced a novel cost function defined from the set of time data \( t = \{t_j\}_{j \in \{1,2,\ldots,n\}} \) resulting from adjacent pair differencing \( t_j = y_{j+1} - y_j = (N_c + \Delta c_j) T_c + \delta_j \), \( j \in \{1,2,\ldots,n' - 1\} \); as a result \( \delta_j = \eta_j + 1 - \eta_j \) becomes a correlated random variable with distribution \( \mathcal{N}(0, 2\sigma_\eta^2) \). Our contribution can be viewed as an extension of a previous work of Sidiropolous et al. [35] who investigated the benefits of round-based processing of ToA data. Following this principle of quantization, we have shown that some key features of the modulation scheme (chip time, frame time,. . .) can be revealed from a mixing of \( n = n' - 1 \) partial functions operating on punctured observation sets; the corresponding cost function takes the form

\[
f(t, \tilde{T}) = \sum_{m=1}^{n} h_m(t, \tilde{T}),
\]

each partial function \( h_m \) being expressed as

\[
h_m(t, \tilde{T}) = \left\lfloor \sum_{j=1, j \neq m}^{n} \frac{t_j}{T} - \text{round} \left( \sum_{j=1, j \neq m}^{n} \frac{t_j}{T} \right) \right\rfloor,
\]

where \( \tilde{T} \) denotes a candidate value of the chip time.

Due to the operations involved, \( f(.) \) exhibits pseudo-periodical oscillations, with increasing frequency as the period hypothesis tends towards \( T_c \), which is a local minimum (Fig. 1); an approximate expression of the pseudo-frequency of oscillation is \( f_c \approx nN_c/T_c \).

The procedure proposed for chip time estimation combines multiple hops (MH) controlled by the pseudo-frequency estimate and a Golden Section Search (GSS). Owing to a limited complexity processing, the cost function envelope is precisely evaluated over the search space while avoiding the problem of staying in a particular local minimum. The oscillation pseudo-frequency depends upon the number of observations taken into account; hence, processing many data sets with different lengths \( n_k \) lead to various groups of local minima “matching only” in the nearby of \( T_c \) (Fig. 2). The proposed algorithm, denoted as Multiple Data Sets - Adjacent (MDS-ADJ), relies on a combination of various minima positions resulting from the processing of multiple data sets.

We proved that this algorithm has a loose performance bound expressed as \( \mathbb{E} \left( \left( \tilde{T} - T_c \right)^2 \right) \geq B_L \equiv (2\sigma_\eta^2) / (n_K N_c)^2 \), \( n_K \) being the longest set of ToA differences used for blind estimation. From Monte-Carlo simulations, we remarked that the proposed MDS-ADJ approach is a good

\( ^3 \)The percent jitter, defined as \( (3\sigma_\eta/T_c) \times 100 \), is considered in the sequel to reflect the measurement noise.
alternative to reference methods such as periodogram [36] or SLS2-ALL [37] (Fig. 3). Its statistical accuracy at low to moderate noise levels is rather close to the Cramér-Rao Bound $B_{T_x}$ (corresponding to ToA differences model), while avoiding both search space sampling and evaluation of trigonometric functions. By diversification of the observed time data, an efficient tradeoff can be achieved between estimation errors and computational complexity. However, the method has the drawback of requiring a relatively restricted search interval (a coarse estimate has to be available first, thanks to the SLS2-ADJ approach [35], for example). Also, no solution for an optimal setting of observation lengths $n = \{n_1, n_2, ..., n_K\}$ is available yet. The sensitivity of MDS-ADJ to outliers is another weak point, which can be slightly mitigated by a proper tuning at the ToA detection stage (increased threshold value to reduce the rate of erroneous data).

As we operate from the bandpass filtered analog signal at the output of the receiving antenna, we make the distinction between the clock-jitter $\zeta_t \sim N(0, \sigma_\zeta^2)$ and the channel noise $\eta(t) \sim N(0, \sigma_\eta^2)$. Again, term $\tau_0 \sim U(0, T_f)$ stands for the initial phase reflecting the transmitter-receiver time offset.

If the received signal is applied to a square law device and then integrated over a time window of duration $w$, energy samples are obtained:

$$z_n = \frac{1}{w} \int_{n w}^{(n+1)w} |s(t)|^2 \, dt, \quad n \in \mathbb{N}. \quad (9)$$

A first stage of the proposed scheme is devoted to ToA estimation from the energy samples, following a maximum selection principle. Then, a second stage yields the chip time estimate by iteratively changing the integration window, depending on the values of a nonlinear objective function $f(\hat{T}, \hat{w})$. The overall structure of the proposed timing scheme is depicted in Fig. 4.

To process the ToA estimation, we first select a reference vector $\mathbf{z}_0 = [z_0, z_1, ..., z_M - 1]$ of length $M$ which will be used to perform correlation:

$$c_k = \sum_{m=1}^{M} \mathbf{z}_0(m) \mathbf{z}_k(m), \quad (10)$$

$\mathbf{z}_k(m)$ denoting the $m$-th element of $\mathbf{z}_k = [z_k, z_{k+1}, ..., z_{k+M-1}]$ and $n_{\text{max}} \geq k + M$ being the total number of collected energy samples.

A maximum value of the temporal spread of the channel is considered, according to the radio environment setup, so that we can choose $M$ and $\mathbf{z}_0$ in such a way that it contains all the energy samples corresponding to first pulse. The relative arrival times $t_j - \tau_0 = \tau_0 + j T_f + c_j T_c + \zeta_j$ are detected by first identifying a confidence region resulting from a threshold crossing in the energy measurements and then performing local maximum selection after correlation over this region. Then, we get estimates $\{\hat{t}_j = t_j^{(1)} - t_j^{(2-1)}\}$ of differential ToA between adjacent pulses, $t_j^{(1)}$ denoting the first energy sample corresponding to $j$-th pulse. The estimation error is of
course directly related to the integration length \( w \); we proposed a bi-dimensional extension of the round-based cost function of Sidiropoulos et al. to evaluate this error:

\[
f(\widetilde{T}, w) = \left\| \frac{\hat{\theta}(w)}{T} - \text{round} \left( \frac{\hat{\theta}(w)}{T} \right) \right\|_1,
\]

where \( \hat{\theta}(w) = \left[ \hat{\theta}_j(w) \right]_{j \in \{2, \ldots, N\}} \) is a vector containing differential ToAs, estimated with an integration window of width \( w \), \( \widetilde{T} \) represents the chip duration hypothesis and \( \| \cdot \|_1 \) stands for \( \ell_1 \) matrix norm.

In the jitter free case, it can be shown that this function vanishes when \( \widetilde{T} = T_c \) and \( w = T_c/r \), \( r \in \mathbb{N} \). If jitter is present, error cannot be put to zero anymore, and a certain couple \( (\widetilde{T}, \widetilde{w}) \) will minimize (11), considered to be the estimate of \( (T_c, w_r) \):

\[
(\hat{T}, \hat{w}) = \arg\min_{\widetilde{T} \in D_T, \widetilde{w} \in D_w} f(\widetilde{T}, \widetilde{w})
\]

where \( D_T \) and \( D_w \) denote the search spaces for the chip time and integration length, respectively.

The abovementioned property of \( f(.) \) enables a very low computational complexity search procedure (very few values are tested for \( w \)) based on iterative one dimensional minimization followed by projection on straight line of equation \( y = \text{round} \cdot w \).

Figure 5 illustrates the results we obtained for 10000 Monte Carlo simulations for CM1 and CM2 channel models (LOS/NLOS) proposed in the frame of IEEE 802.15.4a. The reference observation \( z_0 \) was chosen over a time duration of 300 ns so as to be covering both CM1 and CM2 temporal spreading (hence, to avoid interframe interference), and the considered incertitude time interval for correlation peak search, corresponding to \( j \)-th ToA is \( t_j + [-T_c/2, T_c/2] \). As it can be observed, our approach yields a good chip time estimate for values of jitter below 30%, provided that the SNR is not too low (above –7 dB). Also, the results agree with a theoretical model (not detailed here) that we derived for the differential ToA error resulting from energy detection [30]. Finally, it should be mentioned that the results are strongly related to the number \( N \) of received pulses (\( N = 30 \) in this case).

IV. A CODE-ASSISTED ALGORITHM FOR NDA TIMING ACQUISITION OF PSM SIGNALS

In this section, we turn our attention on timing acquisition for the special case of orthogonal PSM signals, when no pilot symbols are available (NDA mode). In particular, we give an overview of a recent method relying on an original phase coding scheme enabling a synchronization through simple overlap and add operations followed by energy detection [33].

In the standard M-ary PSM modulation format (see equations 3-4), each information bearing symbol \( d(n) \) is conveyed using one pulse from the set \( \{p_0(t), p_1(t), \ldots, p_{M-1}(t)\} \). In the proposed approach, we introduce a binary antipodal pulse coding so that the sample mean estimator of the received signal, computed from consecutive symbol long segments, has its energy minimized when the candidate phase equals the unknown time-offset. With the modified PSM scheme, the symbol-long transmitted waveform takes the following expression (\( \lfloor . \rfloor \) denotes the integer ceil operation):

\[
p_T(t, d(n)) = p_0(t - c_0 T_c) + \gamma_n \sum_{j=1}^{N_f-1} p_{d(n)}(t - jT_f - c_j T_c)
\]

where

\[
\gamma_n = \sum_{m} \alpha_n^m \beta_n^m \in \{-1, +1\}, \quad m = 0, \ldots, M - 1
\]

with

\[
\beta_n^m = \beta_{n-1}^m [1 - 2\alpha_n^m] \in \{-1, +1\}
\]

\[
\alpha_n^m = \left\lfloor \frac{d(n) - m + M}{M} \right\rfloor \mod 2 \in \{0, 1\}
\]

As depicted by Fig. 6, two changes have been made: first, each frame starts with an information-free pulse \( p_0(t) \); second, pulses with alternate phases are used to represent a particular symbol.

![Figure 5. Tc estimation normalized MSE with regard to channel's type and white Gaussian noise (r = 20).](image)

![Figure 6. 4-ary PSM signal format with Nf = 2; (a) Original modulation scheme (b) Alternate signal inversion (changes in \( \gamma_n \)) (c) Information-free first frame.](image)
\[ x(t) = r(t + t_1) : \]

\[ \mathcal{P}(t) = \frac{1}{K} \sum_{k=0}^{K-1} x_k(t) \]

where \( x_k(t) = x(t + kT_s) \), \( k \in [0, K-1] \), \( t \in [0, T_s) \).

In the ISI/IFI free case, it can be shown that the resulting signal is only function of the received waveform associated to information-free pulse \( p_0(t) \):

\[ \mathcal{P}(t) = q_0(t - t_\phi) + q_0(t + T_s - t_\phi) \]

where \( q_0(t) = \sum_{l=0}^{L-1} \lambda_l p_0(t - \tau_l - \tau_0) \), with \( \{\lambda_l, \tau_l\} \) denoting the channel path gains and delays, respectively; \( t_\phi \) denotes the unknown time offset.

As a result, we have the key property that the sample mean takes non-zero values only in the vicinity of \( t_\phi \). The timing acquisition can then be formulated as a maximisation of an objective function measuring the energy of a delayed version of \( \mathcal{P}(t) \):

\[ \hat{t}_\phi = \arg\max_{t_\phi \in [0, T_s]} J(t_\phi), \quad J(t_\phi) = \int_{0}^{T_f} [\mathcal{P}(t + t_\phi) \bmod T_s]^2 dt \]

To limit the computational complexity, \( J(t_\phi) \) may be evaluated over a finite grid of values, at the expense of a reduced synchronization accuracy; in case of frame timing, we can take \( t_\phi = kT_f \), with integer \( k \in [0, N_f) \).

Many Monte Carlo simulations have been carried out to assess the performance of proposed algorithm in terms of normalized mean square error (NMSE), probability of acquisition (PA) and BER against \( E_s/N_0 \) (where \( E_s \) denotes the energy per pulse). A set of optimized pulses \( \{p_i(t)\}_{i=0,1,...,M-1} \) has been used to meet the orthogonality, spectral mask and spectral efficiency requirements [38]; the pulse duration is \( T_f = 1.28 \text{ ns} \). Each symbol consists of \( N_f = 10 \) frames with \( T_f = 12.8 \text{ ns} \), resulting in symbol duration \( T_s = 128 \text{ ns} \). No TH code has been used as we focused on the single user case. The multipath channel employed in simulations is CM1 indoor channel proposed by IEEE 802.15.3a working group [39] (channel impulse responses were first truncated to avoid IFI/ISI).

The results concerning frame acquisition probability, which is defined as the probability \( P_A \) that \( |\hat{t}_\phi - t_\phi| \leq T_f \), are illustrated in Fig. 7; a good overall performance can clearly be observed on this figure, as \( P_A \) is very close to 1 at \( E_s/N_0 = 0 \text{ dB} \) whatever the modulation order is, for \( K = 64 \). An almost perfect acquisition can still be performed at \( E_s/N_0 = -5 \text{ dB} \) for 8-ary modulation. The same approach has also proved to be robust against IFI/ISI over the CM1 channel.

### V. CONCLUSIONS

The issue of blind timing acquisition of IR-UWB signals has been considered in this paper, not only in the sense of NDA mode but also when no prior knowledge is available. In this latter case, it is pointed out that the chip time is a key parameter to estimate to get the unknown time-offset. Two approaches have been overviewed on this subject: the first method relies on a statistical model of the pulse arrival times whereas the second approach jointly estimates ToA and chip duration. Both methods achieve timing through minimization of round-based objective functions. A novel NDA acquisition specific to PSM signal format is also described; thanks to an antipodal pulse coding at the transmitter, it is shown that the synchronization can be performed via simple overlap-add operations and energy detection. This method still performs favorably in presence of IFI/ISI. Various approaches described in this paper may contribute to the development of flexible receivers, able to synchronize at any time despite some changes in the transmitted signal format to best meet spectral mask requirements.

## REFERENCES
