Joint synchronization and demodulation for IR-UWB

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Abstract—This paper addresses the problem of synchronisation and data demodulation for Impulse Radio Ultrawideband systems. A novel non-coherent non-data aided joint synchronization and data demodulation algorithm is proposed. The scheme consists of a coarse estimation phase which is performed over a single symbol period, followed by a fine synchronization stage that estimates the timing with high accuracy based on a frequency domain ToA estimation algorithm. In turn, the demodulator exploits the high time resolution characteristics of UWB signals by taking decisions directly over the timing estimates.

I. INTRODUCTION

UWB communication technology is motivated for some attractive features: robustness to multipath propagation, low probability of interception, high resolution in location applications, high rate in short range, low power and low complexity. Nevertheless, to take advantage of these features, accurate synchronization and channel estimation are needed, being the two main challenges in UWB systems. The Rake receiver [1] exploits the rich multipath diversity of UWB channel at the expense of high complexity due to the large number of fingers required to capture a significant portion of the signal energy. The estimation of the amplitude and delay of the channel paths requires a very high computational load. Furthermore, timing jitter and imperfect channel estimation considerably degrade the performance of the Rake receiver [2]. Non-coherent receivers that do not require channel estimation have been proposed in the literature. In [3] the authors propose a Transmitted Reference (TR) algorithm based on the correlation of the received signal with a reference pulse sent prior to each data pulse. The main drawback of this approach is the energy inefficiency and rate reduction. The Differential Demodulator (DD) propose in [4] correlates adjacent received information waveforms of differentially encoded symbols, avoiding the energy expense in the reference pulse used in TR schemes. Both approaches, Transmitted Reference and Differential Demodulator, avoid channel estimation but require timing synchronization and yield significant performance degradation when timing is imperfect. A different non-coherent demodulator based on dirty templates has been proposed in [5]. In this case, both timing and channel are known. However, the algorithm needs an initialization phase that assumes the knowledge of the two first symbols, that means that the demodulation can not be performed blindly at any time. Recently, several different approaches for UWB timing synchronization have appeared in the literature to solve synchronization. The Maximum Likelihood synchronizer in [2] involves high sampling rate and assumes the knowledge of the received waveform at the receiver. This assumption may not be realistic because the transmitted pulse can be seriously distorted by the antenna and by the channel. Data Aided (DA) and Non Data Aided (NDA) timing estimation based on dirty templates [5] do not require waveform knowledge. However, the proposed NDA algorithm requires high number of symbols for an accurate estimation of the time delay even for high signal to noise ratio. Two DA strategies based on Least Squares methods have been presented in [6] that jointly estimate the channel and the timing delay requiring a high rate sampling. In [7] the timing estimation problem is analysed under the unconditional maximum likelihood criterion. This method does not require any prior knowledge of neither the transmitted symbols (NDA) nor the received waveform. The main drawbacks of this proposal are the high complexity and the considerable length of the observation interval required for the acquisition. This paper addresses a non-coherent receiver that jointly performs NDA timing synchronization and demodulation. The time delay estimation is based on the frequency domain Time Of Arrival (TOA) estimator previously presented by the authors [8]. The estimation is performed in a symbol period and does not need any training data sequence. Furthermore, the frequency domain processing allows the reduction of the sampling rate requirements if filter-bank receivers are used in the receiver front-end [8, 9].

The rest of the paper is organized as follows: Section II introduces the IR-UWB signal model, the timing estimation and demodulation algorithm based on Time Of Arrival estimation is described in Section III, performance evaluation is given in Section IV and conclusions are drawn in V.

II. SYSTEM MODEL

We consider an IR-UWB system where transmission of an information symbol is typically implemented by the repetition of \( N_f \) pulses of very short duration. The transmitted signal is expressed as,

\[
s(t) = \sum_{k=-\infty}^{N_f-1} \sum_{j=0}^{N_f-1} a_{jk} p(t - (kN_f + j)T_f - c_j T_c - b_k T_b) \quad (1)
\]

where Pulse Position Modulation (PPM) is assumed with \( \{b_k\} \) being the information symbols taking values \{0, 1\} with equal
probability. $p(t)$ refers to the single pulse waveform, being typically a Gaussian monocyte or one of its derivatives of duration $T_p$. $T_{sym} = N_fT_f$ is the symbol duration, where $T_f \gg T_p$ is the repetition pulse period also referred to as frame period, and $N_f$ is the number of frames per symbol, $T_c$ is the chip period, $T_s$ is the PPM modulation interval, $N_c$ is the number of chips per frame and $\{c_j\}$ is the time hopping sequence which takes integer values in $\{0, 1, \ldots, N_c-1\}$ and $a_j = \pm 1$ denotes a polarization sequence typically used for spectrum shaping. Without loss of generality we assume in the sequel $a_j = 1 \forall j$.

The channel model considered is given by the general expression for the multipath fading propagation channel as follows,

$$h(t) = \sum_{m=0}^{M-1} h_m \delta(t - \tau_m)$$

(2)

With no loss of generality we assume $\tau_0 < \tau_1 < \ldots < \tau_{M-1}$, being $\tau_0$ the TOA that is to be estimated.

The received signal is then the summation of multiple delayed and attenuated replicas of the received pulse waveform $\hat{p}(t)$ which includes the antenna and filters distortion,

$$y(t) = \sum_{m=0}^{M-1} \sum_{k=-\infty}^{\infty} \sum_{j=0}^{N_f-1} h_m \hat{p}(t - (kN_f + j)T_f - c_jT_c - b_kT_s - \tau_m) + v(t)$$

(3)

We assume the received pulse from each $m$-th path exhibits the same waveform but experiences a different fading coefficient, $h_m$, and time delay, $\tau_m$. The additive noise $v(t) \sim \mathcal{N}(0, N_0)$ is modeled as Gaussian circularly symmetric. Given the low duty cycle of UWB signals we assume the received signal is free of intersymbol interference (ISI).

The signal associated to the $j$-th transmitted pulse corresponding to the $k$-th symbol, in the frequency domain yields:

$$Y_j^k(w) = \sum_{m=0}^{M-1} h_m S_j^k(w) e^{-jw\tau_m} + V(w)$$

(4)

with

$$S_j^k(w) = \hat{P}(w) e^{-jw((kN_f + j)T_f + c_jT_c + b_kT_s)}$$

(5)

$\hat{P}(w)$ and $V(w)$ are the Fourier transform of $\hat{p}(t)$ and $v(t)$ denoting by $\mathcal{F}\{\cdot\}$ the Fourier transform. Sampling (4) at $w_n = \omega_0 n$ for $n = 0, 1, \ldots, N - 1$ being $\omega_0 = 2\pi/N$ and rearranging the frequency domain samples $Y_j^k[n]$ into the vector $Y_j^k \in \mathbb{C}^{N \times 1}$ yields,

$$Y_j^k = S_j^k E_r h + V$$

(6)

where the matrix $S_j^k \in \mathbb{C}^{N \times N}$ is a diagonal matrix which components are the frequency samples of $S_j^k(w)$ and the matrix $E_r \in \mathbb{C}^{N \times M}$ contains the delay-signature vectors (harmonic components) associated to each arriving delayed signal (paths),

$$E_r = [e_{\tau_0} \ldots e_{\tau_j} \ldots e_{\tau_{M-1}}]$$

(7)

with $e_{\tau_j} = [1 e^{-j\omega_0\tau_j} \ldots e^{-j\omega_0(N-1)\tau_j}]^T$.

The channel fading coefficients are arranged in the vector $h \in \mathbb{R}^{M \times 1}$ and the noise samples in vector $V \in \mathbb{C}^{N \times 1}$.

### III. Joint Synchronization and Demodulation Algorithm Based on TOA Estimation

The proposed algorithm resolves synchronization and data demodulation blindly, without the need for channel estimation or transmitted training sequence. The strategy is based on a first block that performs energy estimation which only requires the knowledge by the receiver of the time-hopping sequence, followed by a low complexity high resolution Time Of Arrival estimation algorithm that provides fine timing synchronization, from which data demodulation can be achieved by a direct decision approach. A general block diagram of IR-UWB transmission scheme and the proposed receiver structure is depicted in Fig. 1. For the purpose of describing the joint synchronization and demodulation algorithm the receiver assumes an ideal Nyquist sampling rate, followed by a DFT module. However there exist implementations that can reduce the high sampling rate requirements associated with UWB signals, by using for instance a filter-bank receiver [9].

![Block diagram of the IR-UWB transmission scheme.](image)

**A. Coarse estimation**

The coarse estimation block identifies the beginning of the symbol. Let’s define the time hopping sequence vector as

$$c = [c_0 \ldots c_{N_f-1}]$$

Previous to signal energy estimation we shall define a matrix which elements are the relative chip delay between monocytes in each frame within one symbol. Lets denote $\rho_{c}(n) = N_c - c_{n-1} + c_n$, the number of chips between two consecutive monocytes. Then, we can define the circulant matrix $\Delta_{\rho_c}$ as,

$$\Delta_{\rho_c} = \begin{bmatrix}
\rho_c(1) & \rho_c(2) & \ldots & \rho_c(N_f-1) & \rho_c(N_f) \\
\rho_c(2) & \rho_c(3) & \ldots & \rho_c(N_f) & \rho_c(1) \\
\vdots & \ddots & \ddots & \ddots & \vdots \\
\rho_c(N_f) & \rho_c(1) & \ldots & \rho_c(N_f-2) & \rho_c(N_f-1)
\end{bmatrix}$$

The first frame in the acquisition interval is denoted as $v$ and $\tau_0$ indicates the beginning of the first complete symbol in
the observation interval, being $0 \leq \tau_0 \leq T_{sym}$. Let denote $y[m] = y(mT_s)$ the discrete-time received signal, where $T_s$ is the sampling period.

The frequency domain samples of the received signal in the $i$-th PPM modulation $T_\delta$ interval is defined as,

$$Y_{\delta-i}[n] = \sum_{m=1}^{K_\delta} y[(i-1)K_\delta+m]e^{-j\frac{2\pi}{T_\delta}mn} \quad \text{for } n = 1, \ldots, K_\delta$$

Rearranging the samples $Y_{\delta-i}[n]$ in the vector $Y_{\delta-i} \in \mathbb{C}^{K_\delta \times 1}$, being $K_\delta = \lfloor T_\delta/T_s \rfloor$, the energy at each PPM modulation interval is obtained as,

$$E_{\delta-i} = \|Y_{\delta-i}\|^2 \quad \text{for } i = 1, \ldots, N_\delta N_c N_f$$

being $N_\delta = T_\delta/T_\delta$ the number of PPM modulation intervals per chip. The algorithm then finds the position of the $N_f$ maximum values of the signal energy, $\alpha = [\alpha_1 \ldots \alpha_{N_f}]$, which correspond to the position of the monocyte pulses in the symbol with a temporal resolution equal to the PPM time shift $T_\delta$. Hence, the relative distance between the $N_f$ peaks of the estimated maximum energy values conforms the following vector,

$$\Delta \alpha = [\alpha_2 - \alpha_1 \ldots \alpha_j - \alpha_{j-1} \ldots \alpha_{N_f} - \alpha_{N_f-1}]$$

The estimation of the frame number $\nu$ which corresponds to the first detected pulse is carried out by finding the closest shifted time-hopping sequence to the estimated $\Delta \alpha$. More specifically, the algorithm finds which $j$-th row of the circulant matrix $\Delta_{\rho_c}$, denoted as $\Delta_{\rho_c}[j]$, minimises,

$$\nu = \arg \min_{j=1,\ldots,N_f} \|\Delta \alpha - N_\delta \Delta_{\rho_c}[j]\|^2$$

From the estimated $\nu$, the TOA coarse estimation can be identified as,

$$\hat{\tau}_0^c = T_\delta (\alpha_1 - 1 + N_c - \nu (N_f - \nu) N_c)$$

**B. Fine synchronization and demodulation**

Once the beginning of the symbol is coarsely estimated, joint fine estimation of the time delay and non-coherent blind demodulation is performed on a symbol by symbol basis. The fine time synchronization is based on the frequency domain ToA estimation algorithm previously introduced in [8], which most significant features are reviewed here.

The fine time delay estimation and demodulation comprises the following steps:

a) **Estimation of the correlation matrix** from the frequency domain samples corresponding to the $k$-th symbol, averaging over the properly arranged frame signals in the symbol,

$$R_k = \frac{1}{N_f} Y_k Y_k^H$$

where matrix $Y_k = [Y_k^1 \ldots Y_k^j \ldots Y_k^{N_f}]$ contains column vectors which elements are the DFT samples of the observation signal in a frame period,

$$Y_k^j[n] = \sum_{m=1}^{K_j} y[\tilde{n}_j^k + m]e^{-j\frac{2\pi}{T_\delta}nm} \quad \text{for } n = 1, \ldots, K_j$$

where $K_j = \lfloor T_f/T_s \rfloor$ is the number of samples in the frame period and $\tilde{n}_j^k$ is the first sample associated to the $j$-th transmitted pulse corresponding to the $k$-th symbol,

$$\tilde{n}_j^k = \lfloor \hat{\tau}_0^c - T_\delta - (kN_f + j)T_f - c_j T_c \rfloor/T_s$$

The PPM modulation time shift $T_\delta$ is subtracted because the beginning of the symbol, $\mu$, has been indistinctly estimated for either $b_k = 0$ or $b_k = 1$. **Calculation of the power delay profile** defined as the signal energy distribution with respect to propagation delays, from $R_k$. In particular, the power delay profile is obtained by computing the following quadratic form for different values of $\tau$,

$$P_k(\tau) = e^H \mathbf{R}_k e$$

The so called pseudo-periodogram, $P_k(\tau)$, allows for a low complexity implementation by means of a Fast Fourier Transform (FFT) applied to the following coefficients,

$$\hat{R}_n = \left\{ \sum_{j=1}^{N} R_k(j-n,j) \quad 0 \leq n \leq N - 1 \\right\} \sum_{j=1}^{N+n} R_k(j-n,j) \quad -N + 1 \leq n < 0$$

where $\hat{R}_n$ is the sum of the $n$-th diagonal elements of the correlation matrix $R_k$ and $R_k(i,j)$ denotes the $i$-th row, $j$-th column element of $R_k$. Then $P_k(\tau) = \text{FFT}_L \{ \hat{R}_{-N+1} \ldots \hat{R}_0 \ldots \hat{R}_{N-1} \}$ where FFT$_L(.)$ denotes an FFT operation of length $L$.

c) **Estimate the Time Of Arrival** by searching for the first "peak" (a peak is defined as a relative maximum in the power delay profile) value that exceeds a given threshold, $P_{th}$, in the pseudo-periodogram,

$$\hat{\tau}_k = \min_{\nu} \{ P_k(\tau) > P_{th} \}$$

where $P_{th}$ can be defined as the noise power obtained from the energy calculated in the coarse estimation process.

d) **Symbol demodulation**. The $k$-th symbol can be jointly synchronised and demodulated directly from the time delay estimate $\hat{\tau}_k$, without any knowledge of the channel impulse response and without the need of a training data sequence. The procedure is as follows: depending on the value of the first symbol $b_0$, used to estimate the beginning of the symbol $\hat{\tau}_0^c$, the estimated delays will take values,

$$b_0 = 0 \rightarrow \left\{ \begin{array}{l} b_k = 0 \Rightarrow \hat{\tau}_k > T_\delta \\ b_k = 1 \Rightarrow \hat{\tau}_k > 2T_\delta \end{array} \right.$$
\[ b_0 = 1 \rightarrow \begin{cases} b_k = 0 & \tilde{\tau}_k < T_\delta \\ b_k = 1 & \tilde{\tau}_k > T_\delta \end{cases} \]

Thus, it is needed to solve the ambiguity caused by the lack of knowledge of the data value. The proposed algorithm defines a differential demodulated symbol \( \hat{a}_k \) as,

\[
\hat{a}_k = \begin{cases} 0 & \tilde{\tau}_k < T_\delta \\ 1 & T_\delta < \tilde{\tau}_k < 2T_\delta \\ 2 & \tilde{\tau}_k > 2T_\delta \end{cases}
\]

The demodulated symbols are then,

\[
\hat{i}_k = \begin{cases} \hat{a}_k & \text{if } \hat{a}_k = 0, 1 \\ \hat{a}_k - 1 & \text{if } \hat{a}_k = 1, 2 \end{cases}
\]

(e) **Fine synchronization.** Fine time delay estimation is directly obtained from the application of a direct decision to the demodulated data. More specifically,

\[
\tilde{\tau}_k = \tilde{\tau}_0 + \hat{i}_k - b_k T_\delta - T_\delta
\]

IV. **NUMERICAL RESULTS**

For numerical evaluation of the algorithms we consider the IEEE 802.15.4a channel models.

The performance of the timing estimator and demodulator has been evaluated over 2000 channel realizations of the channel type 3. The channel is considered fixed during one data block duration. The data block has a length of 25 bits, modulated to 2-PPM symbols. The most relevant parameters considered in the simulation are shown in Table I.

Fig. 2 shows the bit error rate of the proposed demodulator when perfect timing synchronization is assumed (dashed line) and for the proposed non-coherent demodulator (solid line).

In Fig. 3 the normalised root mean squared (RMSE) timing estimation error is depicted for a completely blind estimation. The RMSE has been normalised to the symbol duration.

<table>
<thead>
<tr>
<th>PARAMETER</th>
<th>VALUE</th>
</tr>
</thead>
<tbody>
<tr>
<td>Pulse duration, ( T_p )</td>
<td>0.4 ns</td>
</tr>
<tr>
<td>Frame duration, ( T_f )</td>
<td>400 ns</td>
</tr>
<tr>
<td>Chip duration, ( T_c )</td>
<td>20 ns</td>
</tr>
<tr>
<td>Number of chips, ( N_c )</td>
<td>20</td>
</tr>
<tr>
<td>Number of frames in one symbol, ( N_f )</td>
<td>20</td>
</tr>
<tr>
<td>2-PPM modulation, ( T_\delta )</td>
<td>( T_c/2 )</td>
</tr>
<tr>
<td>Pulse waveform ( p(t) )</td>
<td>Gaussian monocycle</td>
</tr>
</tbody>
</table>

V. **CONCLUSION**

A novel joint NDA timing synchronisation estimator and noncoherent demodulator has been proposed for IR-UWB systems. Time delay estimation is performed in frequency domain yielding accurate synchronisation. The joint approach avoids channel estimation, allows acquisition in a symbol period and presents robustness to imperfect timing.

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