BLIND SYNCHRONIZATION IN MULTIUSER TRANSMIT-REFERENCE UWB SYSTEMS

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Abstract

Ultrawideband (UWB) or impulse radio wireless communication systems are based on the transmission of extremely narrow pulses, with a duration inferior to a nanosecond. By design, Transmit-Reference (TR) UWB systems can avoid channel estimation at the receiver, while different users can share the same available bandwidth by using different spreading codes, similar to CDMA systems. This allows the receiver to separate different users, but, more crucially, to recover timing information of the transmitted packets and thus achieve synchronization within a short, burst-like packet transmission. By recognizing that a shift in time corresponds to a phase rotation in the frequency domain, a blind synchronization algorithm that takes advantage of the shift invariance structure in the frequency domain is proposed in this paper, allowing for a fast, high-resolution packet offset estimation.

1. INTRODUCTION

Ultrawideband or impulse radio is currently being proposed as a high data rate wireless communication system for short distances. While the potential data rates are extremely high due to a bandwidth usage of 500 MHz or more, a number of problems at signal transmission and reception must be addressed (see [1] for an overview of UWB signal processing and communications challenges). Much unlike conventional communication systems, impulse radio does not employ a carrier signal in order to convey the modulated data, but uses a carrier-less or baseband approach where data is modulated onto sub-nanosecond pulses. In order to allow coexistence with deployed narrowband systems such as GSM, GPS, and WLAN, transmitted data symbols need to be spread below noise level. Current technology does not allow sampling at the required giga-sample per second rates, and therefore classical signal processing techniques for channel estimation cannot be used without a considerable data rate or complexity penalty. To avoid channel estimation, the usage of a pulse as a reference or template to decode subsequent pulses leads to a simplified receiver, with a maximal decrease of data rate of 50 %, pioneered in early communication and radar systems. This Transmit-Reference (TR) transceiver scheme proposed by Hoctor and Tomlinson in [2], captures the energy of the channel by means of correlation and integration at the analog stage of the receiver side.

The problem faced in UWB communication systems is the burst-like nature of transmissions and the sharing of the spectrum among own-system interferers and narrowband interferers. In such ad-hoc communication systems, it is of great importance to estimate the beginning of the data packet of interest in order to subsequently estimate the data symbols. We assume that symbols are spread using amplitude codes and time-hopping codes, and are transmitted in individual data packets that have a random time offset. The proposed combined blind synchronization and detection algorithm will process blocks of received data samples, and compute a high resolution estimate of the starting point of the packet based on an efficient matching of the desired user code. This synchronization scheme allows for fast data exchange between users in an ad-hoc UWB network.

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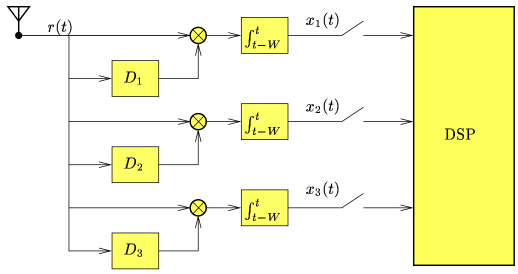

Figure 1: The structure of the autocorrelation receiver.

In this paper we present the data model for a multiuser Transmit-Reference Ultrawideband (TR-UWB) system, and a combined blind synchronization and detection scheme for TR-UWB transceivers, taking into account a channel with a long impulse response. The proposed algorithm processes a block of received data samples, takes advantage of a shift invariance structure in the frequency domain, and applies a MUSIC-like search to estimate the delay of the data packet.

2. ASYNCHRONOUS MULTI-USER DATA MODEL

2.1 TR-UWB system description

In this paper we consider a TR-UWB system that uses a stream of extremely short pulses (order of ns) to convey the information between users [2, 3]. First, we define a 'doublet' as a pair of pulses separated by a time interval D_i chosen from a set of predefined delays: $D_i \in \{D_1, \cdots, D_M\}$. The fixed time interval that contains a doublet is called a 'frame'. Both pulses propagate through the same channel. Combining them constructively at the receiver side, enables the energy smeared by the multipath channel to be recollected. Several doublets employing the same delay D_i form a CDMA kind of chip $\tilde{c}_j = \{\pm 1\}$ of duration T_c. One data symbol $s_k \in \{\pm 1\}$ comprises several chips, *i.e.*, $\tilde{c}_j = c_{j \, mod \, N_c} \cdot s_{\lfloor j/N_c \rfloor}$, with $c_j \in \{\pm 1\}$ for $j \in \{0, 1, \cdots, N_c - 1\}$, where N_c is the length of the polarization (CDMA kind of) code. Within a single chip, the delay D_i and the polarization of the information carrier pulse $c_j = \{\pm 1\}$ are kept constant according to the 'time-hopping' and polarization code, respectively.

As in [2] we implement the autocorrelation receiver presented in figure 1 in order to reconstruct the data. The signal from the antenna output $r(t)$ is correlated (multiplied) by delayed versions of itself for all possible delays $D_1, \cdots, D_M$. Subsequently, signals are integrated using a sliding window of size $W = T_c$ and afterwards sampled P times per chip duration T_c, where P is the oversampling rate. A data model for the synchronous single symbol, single user case under the assumption that the channel is shorter than the frame duration is derived in [3].

An extension to the asynchronous multiple symbol, single user case is presented in [4] and is extended below to the multi-user case. This is not trivial, since next to the autocorrelation terms of the different users, there are also crosscorrelation terms, due to the use of the autocorrelation receiver. However, since different users employ distinct time-hopping and polarization codes, propagate through different channels, and arrive at the receiver at random time instants, we can treat these cross terms as white noise. In this manner, the algorithm used to synchronize to a single user [4] can be extended for synchronization of multiple users, as explained in section 3.

2.2 Analog received data model

Similar to [3] but now for the asynchronous multiple symbol, multi-user case the output at the mth ($m \in \{1, 2, \cdots, M\}$) integrator for the qth ($q = \{1, 2, \cdots, Q\}$) user becomes

$$x_m^{(q)}(t) = \sum_{i=1}^{M} \sum_{j=kN_c}^{(k+1)N_c-1} p(t - jT_c - \tau^{(q)})(\alpha_{mi}^{(q)} J_{ij}^{(q)} \tilde{c}_j^{(q)} + \beta_{mi}^{(q)} J_{ij}^{(q)}) \quad (1)$$

Henceforth q represents the user index. The pulse $p(t)$ has a staircase 'tent' shape due to the integration of multiple frames per chip. $p(t)$ is of duration $2T_c$, height one [3] and is assumed to be known. All users employ short CDMA codes. Therefore we define $\tau^{(q)}$ as the offset of a data packet with respect to the beginning of the received data block as in figure 2. Without loss of generality and for simplicity reasons in the derivation of the data model we assume an integer packet offset $\tau^{(q)}$. In [3] it is shown that even in cases where the transmit delay $D_i^{(q)}$ and the receive delay $D_m^{(q)}$ do not match, some residual useful information remains. This effect is represented by unknown channel parameters $\alpha_{mi}^{(q)}$ and $\beta_{mi}^{(q)}$ where the first is a scaling and the latter depicts a bias. $J_{ij}^{(q)}$ determines which time interval $D_i^{(q)}$ is used between the pulses of a doublet within the jth chip. Specifically, $J_{ij}^{(q)} = 1$ if delay $D_i^{(q)}$ is used for chip j and $J_{ij}^{(q)} = 0$ otherwise.

2.3 Matrix model

The sequence $x_m^{(q)}(t)$ is subsequently sampled at rate P/T_c where typically $P = 2$. The sampled output of the m-th integrator in the kth symbol period is given by

$$\mathbf{x}_{m,k}^{(q)} = \begin{bmatrix} x_{m,kN}^{(q)} \\ \vdots \\ x_{m,(k+1)N-1}^{(q)} \end{bmatrix}$$

where we define $N = N_c P$. Stacking horizontally $\mathbf{x}_{m,k}^{(q)}$ for all integrator outputs $m = 1, \cdots, M$ in the kth sampling period produces $\mathbf{X}_k^{(q)} = [\mathbf{x}_{1,k}^{(q)}, \cdots, \mathbf{x}_{M,k}^{(q)}]$, an $N \times M$ matrix.

When transmission of multiple data symbols is considered, inter-symbol interference arises and therefore at most two data symbols of user q affect a single block of received data $\mathbf{X}_k^{(q)}$. Stacking $\mathbf{X}_k^{(q)}$ and $\mathbf{X}_{k+1}^{(q)}$ vertically yields the block data model of the received data samples

$$\begin{bmatrix} \mathbf{X}_k^{(q)} \\ \mathbf{X}_{k+1}^{(q)} \end{bmatrix} = [\underline{\mathbf{Z}}_1^{(q)} \;\; \underline{\mathbf{Z}}_2^{(q)} \;\; \underline{\mathbf{Z}}_3^{(q)} \;\; \mathbf{1}] \begin{bmatrix} \mathbf{A}^{(q)} s_k^{(q)} \\ \mathbf{A}^{(q)} s_{k+1}^{(q)} \\ \mathbf{A}^{(q)} s_{k+2}^{(q)} \\ \mathbf{b}^{(q)T} \end{bmatrix} . \quad (2)$$

We define $[\mathbf{A}^{(q)}]_{ij} = \alpha_{ij}^{(q)}{:}M \times M$ and $[\mathbf{B}^{(q)}]_{ij} = \beta_{ij}^{(q)}{:}M \times M$ as the matrices that collect all channel scaling- ($\alpha_{ij}^{(q)}$) and bias- ($\beta_{ij}^{(q)}$) coefficients, respectively. It can be shown that $\mathbf{A}^{(q)}$ is a symmetric ($\mathbf{A} = \mathbf{A}^{\mathrm{T}}$) Toeplitz matrix with dominant positive diagonal elements, while $\mathbf{B}^{(q)}$ is a rank one matrix *i.e.,* $\mathbf{B}^{(q)} = \mathbf{b}^{(q)} \mathbf{1}^{\mathrm{T}}$ where $\mathbf{b}^{(q)}$ and $\mathbf{1}$ explicitly appear in (2). The k-th data symbol of user q is represented by $s_k^{(q)}$. The block columns $\underline{\mathbf{Z}}_i^{(q)}$ of size $2N \times M$ comprise the effects of the time-hopping and polarization codes and the effect of the pulse shape $p(t)$. We begin with defining the second block column $\underline{\mathbf{Z}}_2^{(q)} = [\mathbf{0}_{\tau^{(q)}P}^{\mathrm{T}} \; \mathbf{Z}^{(q)T} \; \mathbf{0}_{N-(\tau^{(q)}+1)P}^{\mathrm{T}}]^{\mathrm{T}}$ that comprises the complete version of a known, user specific block code $\mathbf{Z}^{(q)T}$ shifted by an integer delay $\tau^{(q)}$ where $\mathbf{Z}^{(q)T} = \mathbf{P}\mathrm{diag}(\mathbf{c}^{(q)})\mathbf{J}^{(q)}$. $[\mathbf{P}]_{ij} = p(i-1)\frac{T_c}{P} - (j-1)T_c{:}N \times N_c$ contains one sampled version of the pulse shape $p(t)$ per column. Note that inter-chip interference (ICI) of P samples arises as the effect of the sliding window integration. $\mathbf{0}_{\tau^{(q)}P}{:}\tau^{(q)}P \times M$ and $\mathbf{0}_{N-(\tau^{(q)}+1)P}{:}N - (\tau^{(q)}+1)P \times M$ are matrices of all zero elements. $\mathbf{c}^{(q)} = [c_1^{(q)}, \cdots, c_{N_c-1}^{(q)}]^{\mathrm{T}}{:}N_c \times 1$ depicts the outcome of the polarization (CDMA kind) spreading code. Furthermore, $[\mathbf{J}^{(q)}]_{ij} = J_{ij}^{(q)}$ is of size $N_c \times M$ and has a single non-zero element per column, which determines what delay $D_i^{(q)}$ is used for the j-th chip thus, $\mathbf{J}^{(q)}$ determines the time-hopping scheme of user q.

Now we define $\underline{\mathbf{Z}}_1^{(q)} = [\mathbf{Z}'^{(q)T} \; \mathbf{0}_{2N-(\tau^{(q)}+1)P}^{\mathrm{T}}]^{\mathrm{T}}$ and $\underline{\mathbf{Z}}_3^{(q)} = [\mathbf{0}_{N+\tau P}^{\mathrm{T}} \; \mathbf{Z}''^{(q)T}]^{\mathrm{T}}$, which contain only part of the user block code $\mathbf{Z}^{(q)}$. $\mathbf{Z}'^{(q)T}{:}(\tau^{(q)}+1)P \times M$ and $\mathbf{Z}''^{(q)}{:}N - \tau P \times M$ depict the effect of a 'previous' and a 'subsequent' data symbol, respectively.

Equation (2) can be written more compact as

$$\begin{bmatrix} \mathbf{X}_k^{(q)} \\ \mathbf{X}_{k+1}^{(q)} \end{bmatrix} = [\mathbf{G}'^{(q)} \; \mathbf{1}] \begin{bmatrix} \mathbf{S}_k^{(q)} \\ \mathbf{b}^{(q)T} \end{bmatrix} . \quad (3)$$

Here $\mathbf{S}_k^{(q)}$ contains the effect of three consecutive data symbols and the channel coefficient matrix *i.e.,* $\mathbf{S}_k^{(q)T} = [\mathbf{A}^{(q)} s_k \, | \mathbf{A}^{(q)} s_{k+1} \, | \mathbf{A}^{(q)} s_{k+2}]^{\mathrm{T}}$. Note that $\mathbf{A}^{(q)} = \mathbf{A}^{(q)T}$. In the proposed data model, the unknowns are the data packet offsets of all existing users $\tau^{(q)}$, the channel parameters $\mathbf{A}^{(q)}$ and $\mathbf{b}^{(q)}$ and the data symbols $s_k^{(q)}$.

2.4 Resulting asynchronous multi-user model

Define

$$\mathbf{X}^{(q)} = \begin{bmatrix} \mathbf{X}_1^{(q)} & \mathbf{X}_2^{(q)} & \cdots & \mathbf{X}_n^{(q)} \\ \mathbf{X}_2^{(q)} & \mathbf{X}_3^{(q)} & \cdots & \mathbf{X}_{n+1}^{(q)} \end{bmatrix},$$

as the stack of received data blocks $\mathbf{X}_k^{(q)}$ collected over $n+1$ consecutive analysis windows of user q. Now we can extend the asynchronous single user data model (3) to the case multiple users are considered. Define $\mathbf{X}$ as the matrix that

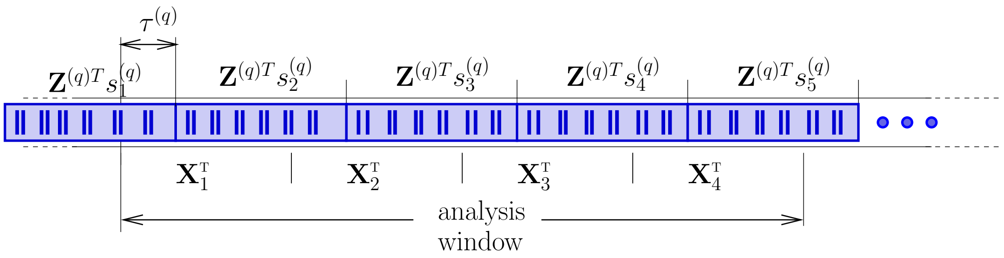

Figure 2: The structure of analysis window for the asynchronous TR-UWB scheme.

collects the contribution of all Q users at the receiver side:

$$\mathbf{X} = \sum_{q=1}^{Q} \mathbf{X}^{(q)} = [\mathbf{G}'^{(1)}|\cdots|\mathbf{G}'^{(Q)}|\mathbf{1}] \begin{bmatrix} \mathbf{S}^{(1)} \\ \vdots \\ \mathbf{S}^{(Q)} \\ \underline{\mathbf{b}}^{\mathrm{T}} \end{bmatrix} . \qquad (4)$$

Here $\mathbf{S}^{(q)} = [\mathbf{S}_1^{(q)}, \cdots, \mathbf{S}_n^{(q)}]$ is the collection of n consecutive $3M \times M$ data blocks defined as in (3). Note that the $\mathbf{b}^{(q)}$'s for all $q = 1, \cdots, Q$ will add up to produce $\mathbf{b}^{\mathrm{T}} = \sum_{q=1}^{Q} \mathbf{b}^{(q)T}$. Stacking horizontally n copies of $\mathbf{b}^{\mathrm{T}}$ we obtain $\underline{\mathbf{b}}^{\mathrm{T}}$ of size $1 \times nM$.

Note that in the case some users are not active for the duration of the whole analysis window $\mathbf{X}$, several $\mathbf{S}_k^{(q)}$ matrices will be zero and some small changes in the structure of $\underline{\mathbf{b}}$ will occur. Consequently, a few additional vectors with low energy may emerge in the the left signal subspace.

3. BLIND PACKET OFFSET ESTIMATION

The goal of the synchronization scheme we propose is to estimate the packet offset of each user with respect to the beginning of a received data block $\mathbf{X}_k = \sum_{q=1}^{Q} \mathbf{X}_k^{(q)}$. We propose a deterministic scheme that works on the block of received data $\mathbf{X}$ and exploits subspace techniques.

From (4) we observe that $\mathbf{G}'^{(q)}$ of each user is in the left signal subspace of $\mathbf{X}$. Additionally, the packet offset $\tau^{(q)}$ appears as the unknown in each block column $\underline{\mathbf{Z}}_1^{(q)}$, $\underline{\mathbf{Z}}_2^{(q)}$ and $\underline{\mathbf{Z}}_3^{(q)}$. By combining these two properties we can estimate the packet delay exploiting the fact that $\mathbf{G}'^{(q)}$ is orthogonal to the left nullspace of $\mathbf{X}$ *i.e.*, $\mathbf{U}_0^{\mathrm{H}}\mathbf{G}'^{(q)} = \mathbf{0}$. In order to estimate $\tau^{(q)}$ we minimize

$$\begin{gathered} \operatorname{argmin}_\tau \parallel \mathbf{U}_0^{\mathrm{H}}\mathbf{G}'^{(q)} \parallel^2 = \\ \operatorname{argmin}_\tau \sum_i \parallel \begin{bmatrix} \mathbf{u}_{1,i} \\ \mathbf{u}_{2,i} \end{bmatrix}^{\mathrm{H}} \begin{bmatrix} \mathbf{Z}_2^{(q)} & \mathbf{Z}_1^{(q)} & \mathbf{0} \\ \mathbf{0} & \mathbf{Z}_2^{(q)} & \mathbf{Z}_1^{(q)} \end{bmatrix} \parallel^2 , \end{gathered} \qquad (5)$$

where $\mathbf{u}_{1,i}$ and $\mathbf{u}_{2,i}$ are both of size $N \times 1$ and depict the first and the second half of the i-th column of the zero subspace $\mathbf{U}_0$, respectively. $\mathbf{Z}_1^{(q)}$ and $\mathbf{Z}_2^{(q)}$ are of size $[N \times M]$ and represent the upper and lower half of $\underline{\mathbf{Z}}_2^{(q)}$. We define $\mathbf{Z}_\tau^{(q)} = \underline{\mathbf{Z}}_2^{(q)} = [\mathbf{Z}_1^{(q)T} \ \mathbf{Z}_2^{(q)T}]^{\mathrm{T}}$. Note that $\mathbf{Z}_\tau^{(q)}$ contains the known block code $\mathbf{Z}^{(q)}$ shifted over $\tau^{(q)}$. Restacking (5) similar as in [5] yields

$$\begin{gathered} \operatorname{argmin}_\tau \sum_i \parallel \mathbf{Z}_\tau^{(q)H} \begin{bmatrix} \mathbf{0} & \mathbf{u}_{1,i} & \mathbf{u}_{2,i} \\ \mathbf{u}_{1,i} & \mathbf{u}_{2,i} & \mathbf{0} \end{bmatrix} \parallel^2 = \\ \operatorname{argmin}_\tau \sum_i \parallel \mathbf{Z}_\tau^{(q)H} \mathcal{U}_i \parallel^2 \quad . \end{gathered} \qquad (6)$$

Here, i sweeps all the vectors from the left null space of $\mathbf{X}$. By stacking horizontally $\mathcal{U}_i$ for all possible i's we get the matrix $\mathcal{U}_0$. Now (6) can be written as:

$$\underset{\tau}{\operatorname{argmin}} \parallel \mathbf{Z}_\tau^{(q)H}\mathcal{U}_0 \parallel^2 = \underset{\tau}{\operatorname{argmin}} \parallel \mathbf{z}_{1,\tau}^{(q)H}\mathcal{U}_0|\cdots|\mathbf{z}_{M,\tau}^{(q)H}\mathcal{U}_0 \parallel^2 , \qquad (7)$$

where $\mathbf{z}_{l,\tau}^{(q)H}$ represents the l-th row of $\mathbf{Z}_\tau^{(q)H}$ and subscript τ is a real number indicating the time offset of the data packet and takes any value from the interval $\tau \in [0\ ,\ N)$. In order to obtain a high resolution packet offset estimate, at this point we make use of the property that a shift in the time domain (under a narrowband assumption) corresponds to a phase rotation in the frequency domain as presented in (8). Since fast algorithms are developed for the Fourier transform (*e.g.*, FFT) we obtain a fast, high resolution, low complexity synchronization scheme. More specifically, we can write

$$\mathbf{F}\mathbf{z}_{l,\tau}^{(q)} \quad = \quad \mathbf{D}_\tau \mathbf{F}\mathbf{z}_{l,0}^{(q)} \ , \qquad (8)$$

where $\mathbf{F}$ stands for the discrete Fourier transform matrix, $\mathbf{z}_{l,0}^{(q)} := \mathbf{z}_{l,\tau=0}^{(q)}$ and $\mathbf{D}_\tau = \mathrm{diag}([1, \ldots, e^{-j2\pi\tau/(2N)}])$, where $\mathrm{diag}(\cdot)$ converts a vector to a diagonal matrix and vice versa. Note that $\mathbf{z}_{l,0}^{(q)}$ is known for all $l \in \{1, \cdots, M\}$. We also define $\tilde{\mathbf{z}}_l^{(q)} := \mathbf{F}\mathbf{z}_{l,0}^{(q)}$, $\tilde{\mathcal{U}}_0 := \mathbf{F}\mathcal{U}_0$ and $\boldsymbol{\phi}_\tau = \mathrm{diag}(\mathbf{D}_\tau)$. Applying these definitions to (8) yields $\mathbf{F}\mathbf{z}_{l,\tau}^{(q)} = \mathbf{D}_\tau\tilde{\mathbf{z}}_l^{(q)}$ or equivalently $\mathbf{F}\mathbf{z}_{l,\tau}^{(q)} = \mathrm{diag}(\tilde{\mathbf{z}}_l^{(q)})\boldsymbol{\phi}_\tau$. Using (8), we can rewrite (7) as

$$\begin{aligned} &\operatorname{argmin}_\tau \parallel \mathbf{z}_{l,\tau}^{(q)H}\mathcal{U}_0|\cdots|\mathbf{z}_{M,\tau}^{(q)H}\mathcal{U}_0 \parallel^2 \\ &= \operatorname{argmin}_\tau \parallel \mathbf{z}_{1,\tau}^{(q)H}\mathbf{F}^{\mathrm{H}}\mathbf{F}\mathcal{U}_0|\cdots|\mathbf{z}_{M,\tau}^{(q)H}\mathbf{F}^{\mathrm{H}}\mathbf{F}\mathcal{U}_0 \parallel^2 \\ &= \operatorname{argmin}_\tau \parallel \mathbf{z}_{1,0}^{(q)H}\mathbf{F}^{\mathrm{H}}\mathbf{D}_\tau^*\tilde{\mathcal{U}}_0|\cdots|\mathbf{z}_{M,0}^{(q)H}\mathbf{F}^{\mathrm{H}}\mathbf{D}_\tau^*\tilde{\mathcal{U}}_0 \parallel^2 \\ &= \operatorname{argmin}_\tau \parallel \tilde{\mathbf{z}}_1^{(q)H}\mathbf{D}_\tau^*\tilde{\mathcal{U}}_0|\cdots|\tilde{\mathbf{z}}_M^{(q)H}\mathbf{D}_\tau^*\tilde{\mathcal{U}}_0 \parallel^2 \\ &= \operatorname{argmin}_\tau \textstyle\sum_{l=1}^{M} \parallel \boldsymbol{\phi}_\tau^{\mathrm{H}}\mathrm{diag}(\tilde{\mathbf{z}}_l^{(q)H})\tilde{\mathcal{U}}_0 \parallel^2 \\ &= \operatorname{argmin}_\tau \parallel \boldsymbol{\phi}_\tau^{\mathrm{H}}[\mathrm{diag}(\tilde{\mathbf{z}}_1^{(q)H})\tilde{\mathcal{U}}_0|\cdots|\mathrm{diag}(\tilde{\mathbf{z}}_M^{(q)H})\tilde{\mathcal{U}}_0] \parallel^2 \\ &= \operatorname{argmin}_\tau \parallel \boldsymbol{\phi}_\tau^{\mathrm{H}}\mathcal{K} \parallel^2 \ , \end{aligned} \qquad (9)$$

where * denotes the complex conjugate. Due to the structure of $\boldsymbol{\phi}_\tau$, searching for the $\boldsymbol{\phi}_\tau$ that minimizes the last expression is equivalent to performing an inverse Fourier transform (IFFT) on the matrix $\mathcal{K}$ and searching for the row of the resulting matrix that has the lowest norm. The index k of the row with the lowest norm determines the part of the delay offset that corresponds to an integer multiple, *i.e.*, $\hat{\tau}_{int} = k$. An additional fine grid MUSIC-kind search $\operatorname{argmin}_\tau \parallel \boldsymbol{\phi}_{\hat{\tau}_{int}+\tau}^{\mathrm{H}}\mathcal{K} \parallel^2$ performed around $\hat{\tau}_{int}$ provides a non-integer delay $\hat{\tau}_{frac}$ that takes a value in the interval $[-1/2, 1/2)$. The overall delay estimate is thus $\hat{\tau}^{(q)} = \hat{\tau}_{int} + \hat{\tau}_{frac}$.

4. SIMULATIONS

To evaluate the performance of the proposed algorithm a series of the computer simulations were conducted. The simulation scenario consists of multiple asynchronous users. We

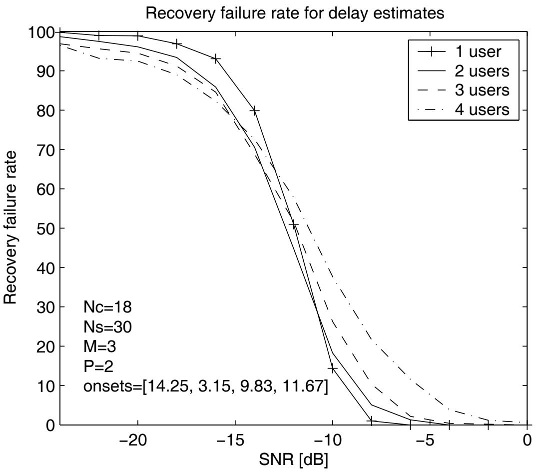

Figure 3: The percentage of incorrectly estimated packet offsets of the user of interest.

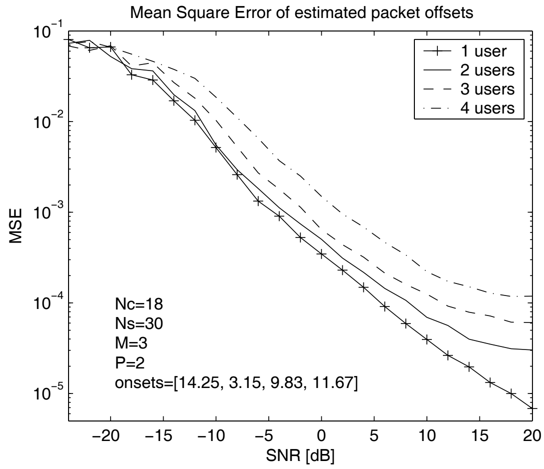

Figure 4: Mean square error of the correctly estimated packet offset delays.

assume $M = 3$ delay positions, polarization codes of length $N_c = 18$ chips, and data packets of $N_s = 30$ data symbols. We perform 1000 Monte Carlo runs, changing the user code, data symbols and noise samples in each run. To avoid extremely time-consuming simulations that would occur when 'real' data is generated we take advantage of the assumptions made in section 2 and generate the data as they appear at the integrator *outputs*. Function $p(t)$ is generated as a triangular pulse shape. Note that the sampling of a triangular pulse shape is the same as the sampling of the 'tent' pulse shape $p(t)$ in the case the oversampling factor $P > 1$ and the number of doublets per chip is an integer multiple of P. The oversampling rate $P = 2$ is lower than the Nyquist rate for the triangular shape used, and introduces some aliasing due to spectral folding, similar as in [6]. Nevertheless, the bias introduced in that manner is insignificant. For the purpose of this simulation, the SNR is defined as the ratio of the signal power of a single user over the AWGN noise power both measured at the integrator *outputs*. (A more accurate but very slow simulation would consider the complete transmission system where the AWGN noise is present at the *input* of the integrators. However, according to [7] the noise at the output can be considered to be AWGN). The packet offsets of multiple users are fixed and are randomly chosen as $[\tau^{(1)}\ \tau^{(2)}\ \tau^{(3)}\ \tau^{(4)}] = [14.25T_c\ 3.15T_c\ 9.83T_c\ 11.67T_c]$, where T_c is the chip duration normalized to $T_c = 1$. We define the signal to noise ratio as $SNR = 10\log(P_s^{(1)}/P_n)$ where $P_s^{(1)} = \| (\mathbf{A}^{(1)}\mathbf{J}^{(1)} \circ \mathbf{P})\mathbf{c}^{(1)} \|^2/(MN)$ is the energy of a single data symbol of the user of interest *i.e.*, user 1, and $P_n = \sigma^2$ stands for the power of the AWGN. The SNR is changed in steps of 2dB. All users are assumed to have equal power at the receiver.

Figure 3 shows the percentage of cases where the packet offset estimate for user 1 is incorrect. An estimate is considered to be incorrect if it does not fall into the interval $\tau - T_c/2 \leq \hat{\tau} < \tau + T_c/2$. We observe the algorithm's performance in the cases different number of users are active simultaneously. Increasing the number of active users results in a slight performance drop, because the total number of vectors in the left nullspace is decreasing and because of higher multi-user interference.

Figure 4 shows the mean square error of the 'good' estimates of τ for the user of interest for different numbers of interfering users. In this figure, we perceive that the reduction of the number of vectors in the zero subspace that is used for the delay estimation affects the variance of the packet offset estimates.

5. CONCLUSION

In this paper we presented the algorithm that provides fast, low complexity, blind packet synchronization in multiuser TR-UWB systems. Therefore it can be used for fast initial code exchange in multiuser asynchronous UWB ad hoc networks.

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