BLIND SYNCHRONIZATION IN ASYNCHRONOUS UWB NETWORKS BASED ON THE TRANSMIT-REFERENCE SCHEME

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Abstract

Ultra wideband (UWB) wireless communication systems are based on the transmission of a stream of narrow pulses (each shorter than a ns). In such a system it is of great importance to estimate the beginning of the data packet of interest in order to subsequently estimate the data symbols. In this paper we present a combined blind synchronization and detection scheme based on the transmit-reference ultra wideband (TR-UWB) transceiver model 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.

1. Introduction

Ultra wideband systems use sub-nanosecond pulses as information carriers. The state of technology does not allow sampling at GHz rate. A solution to this problem is the transmit-reference (TR) transceiver scheme proposed by Hoctor and Tomlinson [1], which captures the energy of the channel by means of correlation and integration at the receiver side. There is no need for channel estimation and the sampling rate is highly reduced facilitating synchronization and symbol estimation. This paper introduces a blind synchronization scheme for a TR-UWB communication system, now in a rich multipath environment. The 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.

The design of UWB systems based on impulse radio is complicated by the fact that the channel length is relatively long compared to the pulse duration [2], and therefore requires equalization. Even if channel estimation would be possible, adaptive digital filtering would require sampling rates of several GHz and is currently unpractical. A system which avoids this and yet requires only data-independent analog processing is offered by the transmit-reference (TR) scheme, as proposed in [1, 3] for UWB. This system is based on the transmission of pairs of pulses (doublets), with a controlled delay between the two pulses. Several doublets form a chip and, as in CDMA, several chips form a symbol. The code consists of an amplitude and a time-hopping code, and allows to distinguish multiple users. Because the UWB signal is generally transmitted below the noise level, further performance improvements to the proposed TR scheme are possible as proposed in [4]. In [5], a data model for a specific TR-UWB receiver was derived, taking into account that the channel has a long impulse response. This is the starting point for the present paper. In contrast to [5], we consider finite data packets with an unknown time offset. The blind synchronization problem is to find the known user code at an unknown offset, which is an extension of a similar problem considered in CDMA, now for a more complicated data model. In particular, we propose an extension of the blind channel estimation algorithm for CDMA proposed by Torlak and Xu [6]. The received data samples are stacked in a matrix such that a shifted version of the user specific block code is in its column span. Subsequently, we exploit the fact that a shift in time domain corresponds to a phase rotation in the frequency domain. Finally, a MUSIC-like search for a shift invariant vector in the signal subspace provides a high resolution delay estimate.

2. Data model

2.1. Receiver structure

In the TR-UWB scheme [1,5], pulses are transmitted in pairs (doublets) which are mutually separated by a delay $D_i \in \{D_1, D_2, \ldots, D_M\}$, where M is the number of possible delays, see figure 1. Both pulses will undergo the same channel, therefore one can be used as a "matched filter" for the other one at the receiver. The first pulse is the reference, the second is modulated by the data. N_t doublets represent a CDMA chip $\tilde{c}_j \in \pm 1$, which has a duration T_c. The value of the delay D_i is constant within a chip and is defined by the so-called time-hopping code. As in CDMA, one data symbol $s_k \in \{-1, +1\}$ comprises several chips, *i.e.*, $\tilde{c}_j = c_{j \bmod N_c} \cdot s_{\lfloor j/N_c \rfloor}$, with $c_j \in \{-1, +1\}$ for $j \in \{0, 1, \cdots, N_c - 1\}$.

A possible receiver structure is depicted in figure 3 [1]. The received data $r(t)$ is delayed over all possible M delays $D_1, \cdots, D_M$, correlated with the non-delayed signal, and subsequently integrated over a sliding window $W = T_c$. For a single symbol transmission s_k, the

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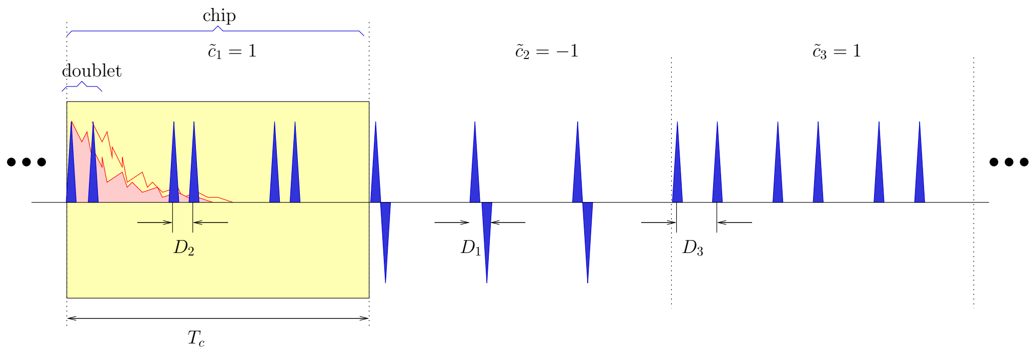


Figure 1. The structure of the transmitted data burst.

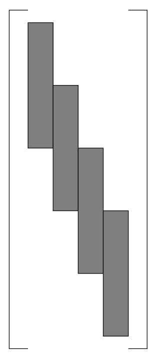
$\mathbf{P} =$ 

Figure 2. The structure of the $\mathbf{P}$ matrix.

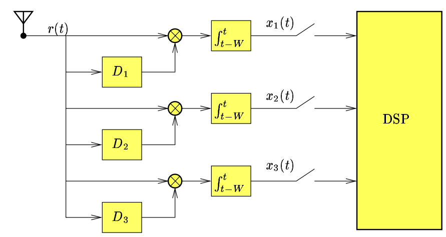


Figure 3. The structure of the autocorrelation receiver.

received signal at lag D_m is modeled as

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

where $m = 1, \cdots, M$. The response to a single symbol will generally have the shape of a "tent", $p(t)$, which is considered to be known. The effect of the channel will be a scaling α_{mi} and voltage offset β_{mi} of this response, which depends on the correlation properties of the channel, the transmitted delay D_i, and the receiver delay D_m as introduced in [5]. J_{ij} determines which transmit delay D_i is used for the j-th chip c_j. $J_{ij} = 1$ if the chip j is transmitted at delay D_i, elsewhere $J_{ij} = 0$. The sequences $x_m(t)$ are subsequently sampled at rate P/T_c, where P is the oversampling factor (typically $P = 2$).

2.2. Single transmitted symbol data model

Let $\mathbf{x}_k$ be a vector of samples of $x_m(t)$ corresponding to the k-th symbol period, of length $MN = MN_cP$ see [5]. Because of the sliding window integration, a single transmitted symbol s_k will generally be spread over two consecutive symbol periods, $\mathbf{x}_k$ and $\mathbf{x}_{k+1}$. Based on (1) and similar as in [5], a data model for an asynchronous single transmitted symbol now becomes

$$\begin{bmatrix} \mathbf{x}_k \\ \mathbf{x}_{k+1} \end{bmatrix} = \begin{bmatrix} \mathbf{0}_\tau \\ \mathbf{AJ} \circ \mathbf{P} \\ \mathbf{0} \end{bmatrix} \mathbf{c} s_k + \begin{bmatrix} \mathbf{0}_\tau \\ \mathbf{BJ} \circ \mathbf{P} \\ \mathbf{0} \end{bmatrix} \mathbf{1} \,.$$

Here, $\circ$ denotes the Khatri-Rao product (column-wise Kronecker product), $\mathbf{A} = [\alpha_{mi}]$ and $\mathbf{B} = [\beta_{mi}]$ are $[M \times M]$ matrices which depend on the channel correlation coefficients, $\mathbf{c} = [c_0 \,, \cdots \,, c_{N_c-1}]^T$ is the known spreading code vector of length N_c chips, $\mathbf{P}$ is a known $[(N_c{+}1)P \times N_c]$ block-Sylvester matrix whose columns are shifts of $p(t)$ as in figure 2. $\mathbf{AJ} \circ \mathbf{P}$ and $\mathbf{BJ} \circ \mathbf{P}$ are both of size $[M(N_c + 1)P \times N_c]$ while zero matrices $\mathbf{0}_\tau$ and $\mathbf{0}$ have τMP and $MP(N_c - 1 - \tau)$ rows, respectively. We define $\tau \in \{0, \ldots, N_c - 1\}$ as an integer data packet delay with respect to the beginning of the received data block. Moreover, $\mathbf{J}$ of size $[M \times N_c]$ is a known large selector matrix which has a single unit element per column that determines the time-hopping pattern for the chip that corresponds to that particular column. Note that column vectors $(\mathbf{AJ} \circ \mathbf{P})\mathbf{c}s_k$ and $(\mathbf{BJ} \circ \mathbf{P})\mathbf{1}$ are both of size $M(N_c + 1)P$ and are MP samples longer than a symbol period $\mathbf{x}_k$. For this reason inter symbol interference (ISI) arises in the case of multiple transmitted symbols.

$\mathbf{A}$ and $\mathbf{B}$, both of size $[M \times M]$, are unknown matrices which depend on the correlation properties of the multipath channel. It can be shown that $\mathbf{A}$ is symmetric, approximately Toeplitz, and diagonally dominant with positive entries on its main diagonal. $\mathbf{B}$ is rank one, $\mathbf{B} = \mathbf{b}\mathbf{1}^{\mathrm{T}}$ where $\mathbf{1}$ is a vector with all entries equal to 1.

To separate the known matrices from the unknown, we restack $\mathbf{x}_k$ into an $[N_cP \times M]$ matrix such that $vec(\mathbf{X}_k) = \mathbf{x}_k$. This yields

$$\begin{bmatrix} \mathbf{X}_k \\ \mathbf{X}_{k+1} \end{bmatrix} = \tilde{\mathbf{P}} \mathrm{diag}(\mathbf{c}) \mathbf{J}^{\mathrm{T}} \mathbf{A}^{\mathrm{T}} s_k + \tilde{\mathbf{P}} \mathbf{J}^{\mathrm{T}} \mathbf{B}^{\mathrm{T}}$$

where

$$\tilde{\mathbf{P}} = \begin{bmatrix} \mathbf{0}_\tau \\ \mathbf{P} \\ \mathbf{0} \end{bmatrix} . \tag{2}$$

This can be written as

$$\begin{aligned}\begin{bmatrix}\mathbf{X}_k \\ \mathbf{X}_{k+1}\end{bmatrix} &= \tilde{\mathbf{P}}\mathrm{diag}(\mathbf{c})\mathbf{J}^{\mathrm{T}}\mathbf{A}^{\mathrm{T}}s_k + \tilde{\mathbf{P}}\mathbf{J}^{\mathrm{T}}\mathbf{B}^{\mathrm{T}} \\ &= \tilde{\mathbf{P}}[\mathrm{diag}(\mathbf{c})\mathbf{J}^{\mathrm{T}} \;\; \mathbf{J}^{\mathrm{T}}][\mathbf{A}s_k \;\; \mathbf{B}]^{\mathrm{T}} \\ &= \tilde{\mathbf{P}}[\mathrm{diag}(\mathbf{c})\mathbf{J}^{\mathrm{T}} \;\; \mathbf{J}^{\mathrm{T}}\mathbf{1}][\mathbf{A}s_k \;\; \mathbf{b}]^{\mathrm{T}} \\ &= \begin{bmatrix}\mathbf{0}_\tau & \mathbf{0}_\tau \\ \mathbf{Z} & \mathbf{q} \\ \mathbf{0} & \mathbf{0}\end{bmatrix}[\mathbf{A}s_k \;\; \mathbf{b}]^{\mathrm{T}}\end{aligned} \tag{3}$$

where $\mathbf{Z} := \mathbf{P}\mathrm{diag}(\mathbf{c})\mathbf{J}^{\mathrm{T}}$ is a known $[(N_c+1)P \times M]$ "code matrix" while $\mathbf{q} := \mathbf{P}\mathbf{J}^{\mathrm{T}}\mathbf{1}$ is a known column vector. Note that $\mathbf{X}_k$ and $\mathbf{X}_{k+1}$ are both of size $[N_cP \times M]$ while matrix $\mathbf{0}_\tau$ and vector $\mathbf{0}$ have τP and $P(N_c - 1 - \tau)$ rows, respectively, and an appropriate number of columns. The delay takes any integer value in the interval $\tau \in \{0\,, \ldots,\; N_cP-1\}$. The channel parameters $\mathbf{A}$, $\mathbf{b}$ and the data symbol s_k are unknown. In this form, the model is very similar to a CDMA data model, with an additional term $\mathbf{q}\mathbf{b}^{\mathrm{T}}$.

2.3. Asynchronous single user model

We now consider a model for an asynchronous single user and several transmitted symbols. In this case, the spread of a symbol over two adjacent symbol periods introduces inter-symbol interference (ISI). Define a received data matrix $\mathbf{X}$ as

$$\mathbf{X} = \begin{bmatrix}\mathbf{X}_1 \; \mathbf{X}_2 \cdots \;\; \mathbf{X}_n \\ \mathbf{X}_2 \; \mathbf{X}_3 \cdots \mathbf{X}_{n+1}\end{bmatrix},$$

where n is the length of the analysis window over which data is collected. We can write the data model as $\mathbf{X} = \mathbf{G}\mathbf{S}$, where the matrices are defined in figure 4. Matrix $\mathbf{G}$ has four block columns. The first three blocks correspond to the "code matrix" and the shifts thereof. In particular, the second block column contains $\mathbf{Z}$ shifted over an unknown delay τ. Its tail $\mathbf{Z}'$ and its head $\mathbf{Z}''$ correspond to the effect of 'previous' and 'next' symbols. Note that $\mathbf{Z} = [\mathbf{Z}''^{\mathrm{T}} \; \mathbf{Z}'^{\mathrm{T}}]^{\mathrm{T}}$. The same partitioning holds for the vector $\mathbf{q}$. The remaining parts of the matrix $\mathbf{G}$ are zero.

In the case where the analysis window is not within the transmitted packet, we can use the same model but allow some of the symbols s_k to be zero.

3. Blind synchronization algorithm

We now describe the synchronization algorithm. In figure 5 the relation between the received data at the integrator outputs $\mathbf{X}_k^{\mathrm{T}}$ and the transmitted symbols is presented. We describe a block algorithm that provides an estimate of the packet offset delay τ.

The aim is to estimate τ and the data $\{s_k\}$. The algorithm is an extension of the algorithm of Torlak and Xu [6], who considered blind channel estimation for CDMA using subspace techniques.

3.1. Integer packet offset

Let $\mathbf{G}'$ represent the first $3M$ columns of the matrix $\mathbf{G}$ as presented in figure 4. The synchronization algorithm uses the property that the matrix $\mathbf{G}'$ is orthogonal to the left nullspace of the matrix $\mathbf{X}$, i.e., $\mathbf{U}_0^{\mathrm{H}}\mathbf{G}' = \mathbf{0}$. We can use this relationship in order to find an estimate of τ. More specifically, we solve

$$\begin{aligned}&\mathrm{argmin}_\tau \parallel \mathbf{U}_0^{\mathrm{H}}\mathbf{G}' \parallel^2 = \\ &\mathrm{argmin}_\tau \textstyle\sum_i \parallel \begin{bmatrix}\mathbf{u}_1^{(i)} \\ \mathbf{u}_2^{(i)}\end{bmatrix}^{\mathrm{H}} \begin{bmatrix}\mathbf{Z}_2 & \mathbf{Z}_1 & \mathbf{0} \\ \mathbf{0} & \mathbf{Z}_2 & \mathbf{Z}_1\end{bmatrix} \parallel^2,\end{aligned} \tag{4}$$

where $\mathbf{u}_1^{(i)}$ and $\mathbf{u}_2^{(i)}$ are both of size $N_cP \times 1$ and depict the first and the second half of the i-th column of $\mathbf{U}_0$, respectively. $\mathbf{Z}_1$ and $\mathbf{Z}_2$ are of size $[N_cP \times M]$ and represent the upper and lower half of the middle block column of $\mathbf{G}'$. Let $\mathbf{Z}_\tau = [\mathbf{Z}_1^{\mathrm{T}} \; \mathbf{Z}_2^{\mathrm{T}}]^{\mathrm{T}}$ be the second block column of the matrix $\mathbf{G}'$. Note that $\mathbf{Z}_\tau$ is known up to the integer packet offset delay τ. Restacking the last equation [6] yields

$$\begin{aligned}&\mathrm{argmin}_\tau \textstyle\sum_i \parallel \mathbf{Z}_\tau^{\mathrm{H}} \begin{bmatrix}\mathbf{0} & \mathbf{u}_1^{(i)} & \mathbf{u}_2^{(i)} \\ \mathbf{u}_1^{(i)} & \mathbf{u}_2^{(i)} & \mathbf{0}\end{bmatrix} \parallel^2 = \\ &\mathrm{argmin}_\tau \textstyle\sum_i \parallel \mathbf{Z}_\tau^{\mathrm{H}}\mathcal{U}^{(i)} \parallel^2 \quad .\end{aligned} \tag{5}$$

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 (5) can be written as:

$$\underset{\tau}{\mathrm{argmin}} \parallel \mathbf{Z}_\tau^{\mathrm{H}}\mathcal{U}_0 \parallel^2 \quad . \tag{6}$$

This can be solved by performing the enumeration over $\tau \in \{0, \cdots, PN_c - 1\}$, since we know $\mathbf{Z}_\tau$ up to the integer packet offset delay τ. The resolution of this algorithm is limited by the oversampling rate $1/P$.

3.2. Non-integer packet offset

The idea presented in section 3.1 can be made computationally more efficient and can also be implemented for the estimation of non-integer data packet offsets by use of the Fourier transformation (FFT). The FFT transforms a delay into a phase progression, and it will be shown that the problem transforms into a MUSIC-like estimation problem. We start by rewriting (5) as

$$\begin{aligned}&\mathrm{argmin}_\tau \parallel \mathbf{Z}_\tau^{\mathrm{H}}\mathcal{U}_0 \parallel^2 = \\ &\mathrm{argmin}_\tau \parallel \mathbf{z}_\tau^{(1)H}\mathcal{U}_0 | \mathbf{z}_\tau^{(2)H}\mathcal{U}_0 | \cdots | \mathbf{z}_\tau^{(M)H}\mathcal{U}_0 \parallel^2,\end{aligned} \tag{7}$$

here $\mathbf{z}_\tau^{(l)H}$ represents the l-th row of $\mathbf{Z}_\tau^{\mathrm{H}}$ and subscript τ is a real number indicating the time offset of the data packet and takes any value from the interval $\tau \in [0\;,\; PN_c)$. Recognizing the property that a shift (under a narrowband assumption) in the time domain corresponds to a phase shift in the frequency domain, we have

$$\mathbf{F}\mathbf{z}_\tau^{(l)} = \mathbf{D}_\tau\mathbf{F}\mathbf{z}_0^{(l)} \quad . \tag{8}$$

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

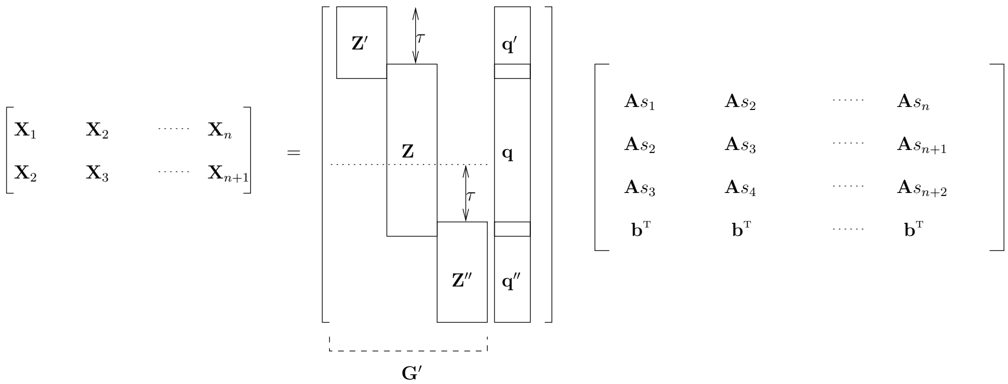


Figure 4. Block data model $\mathbf{X} = \mathbf{GS}$ for the asynchronous single user case using a TR UWB scheme.

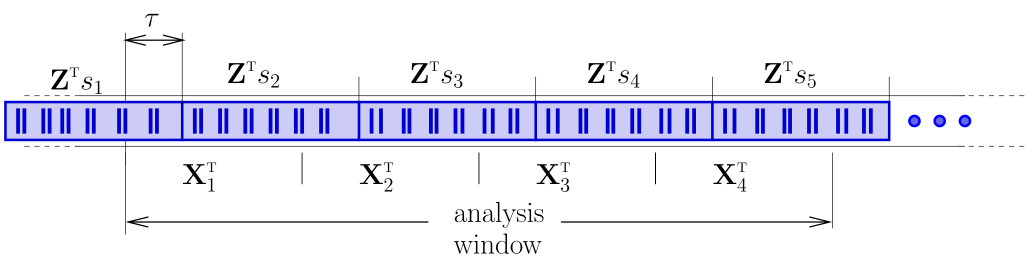


Figure 5. The structure of analysis window for the asynchronous TR-UWB scheme.

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

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 search 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 $\mathrm{argmin}_\tau \parallel \boldsymbol{\phi}_{\hat{\tau}_{int}+\tau}^{\mathrm{H}}\mathcal{K} \parallel^2$ 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 $\hat{\tau} = \hat{\tau}_{int} + \hat{\tau}_{frac}$.

3.3. Symbol estimation

After estimating the packet offset τ, we can reconstruct the complete $\mathbf{G}$ matrix. Estimation of the transmitted data symbols is now possible by performing a deconvolution of the matrix $\mathbf{X}$ using the known user code, i.e., we compute $\mathbf{S} = \mathbf{G}^\dagger\mathbf{X}$ where $\dagger$ denotes the pseudo-inverse. Considering $\mathbf{b}^{\mathrm{T}}$ as a nuisance part of $\mathbf{S}$ we simply chop it obtaining $\mathbf{S}'$. We now stick to the middle block row of $\mathbf{S}'$, name it $\tilde{\mathbf{S}}'$ as the part that carries most of the energy. The data symbols at this point can be estimated from $\tilde{\mathbf{S}}'$ in two different ways: (i) by computation of the trace of the $[M \times M]$ data blocks that have the $\mathbf{A}s_k$ structure, or (ii) by vectorizing the $[M \times M]$ blocks of the matrix $\tilde{\mathbf{S}}'$ such that we get a rank one matrix whose row span corresponds to the data symbols. Vectorizing is performed over subsequent estimated $\mathbf{A}s_k$ blocks. Vectors are stacked in a new matrix whose row span represents a scaled version of the data symbols s_k. The estimates can be further refined by iteration.

4. Simulations

We test the algorithm on simulated data involving a single user. We consider code length $N_c = 18$, $M = 3$ possible delay lags, $N_s = 30$ data symbols and oversampling rate $P = 2$. We perform 1000 Monte-Carlo runs changing the user code, data symbols, and noise samples in each run. The packet offset is $12.25T_c$, where $T_c = 1$ is the chip period normalized to 1. Generating the 'real' data is too time-consuming, therefore we create the signals as they appear at the output of the integrators in accordance with the results obtained in [5]. Function $p(t)$ is generated as a triangular pulse shape of duration $2T_c$. The oversampling rate $P = 2$ is still lower than the Nyquist rate for a triangular pulse shape but is sufficient as the distortions in the estimates of τ are negligible compared to sampling with the Nyquist rate. For the purpose of this simulation, the SNR is defined as the ratio of the signal power 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. Nevertheless, according to [7] the noise at the output can be considered to

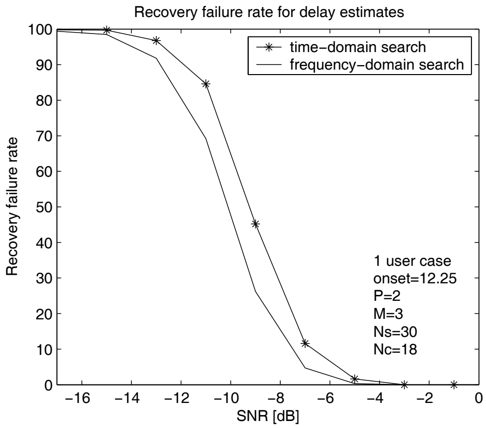


Figure 6. The percentage of incorrectly estimated packet offsets using the time- and frequency-domain approach.

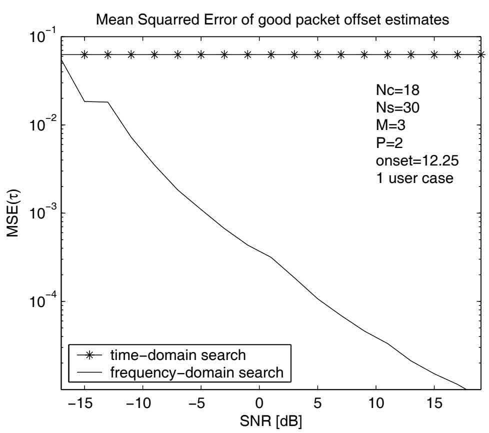


Figure 7. Mean square error of the correctly estimated packet offset delays.

be AWGN). Therefore, we use $SNR = 10\log(P_s/P_n)$ where $P_s = \parallel (\mathbf{AJ} \circ \mathbf{P})\mathbf{c} \parallel^2/(MN)$ is the energy of a single data symbol while $P_n = \sigma^2$ stands for the power of the AWGN. The SNR is changed in steps of 2dB.

Figure 6 shows the percentage of cases where the packet offset is estimated wrongly. An estimate is considered to be wrong if it does not fall into the interval $\tau - T_c/2 \leq \hat{\tau} < \tau + T_c/2$. The solid line depicts the performance of the delay estimation based on the frequency-domain search (as presented in section 3.2), while the solid-asterix line shows the performance of the time-domain estimator (see section 3.1). Figure 7 shows the mean square error of the 'good' estimates of τ for both time- (solid-asterix) and frequency-domain search (solid). The precision of the time domain search is limited by the oversampling rate $1/P$. In figure 8 the BER of the symbol estimates is shown for the τ estimated in the time- and frequency- domain, where the symbol estimates are obtained by use of both the diagonal dominant structure of $\mathbf{A}s_k$ and the vectorization approach (section 3.3).

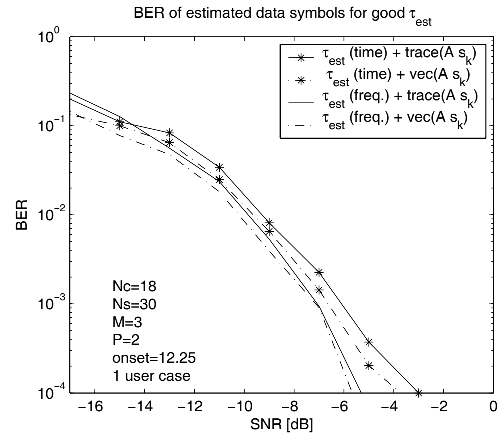


Figure 8. BER of the estimated data symbols computed for 'good' estimates of τ (time and frequency search). Symbols are estimated using both the dominant diagonal structure of $\mathbf{A}s_k$ and the vectorization approach.

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