STATISTICAL ANALYSIS OF A TRANSMIT-REFERENCE UWB WIRELESS COMMUNICATION SYSTEM

Quang Hieu Dang, Alle-Jan van der Veen, António Trindade

Circuits & Systems, Department of Electrical Engineering
Delft University of Technology, The Netherlands

The delay-hopped transmit-reference ultra wideband communication system introduced by Hoctor and Tomlinson results in a receiver based on a bank of correlators and a sliding window integrator. In this paper, we derive a complete signal processing model for this receiver structure. In particular, we take the effects of the radio propagation channel on the correlations into account, as well as the effects of the additive noise. Blind and semi-blind receiver algorithms are proposed to estimate the effective channel coefficients and the data, and the performance of the algorithms is tested in simulations.

1. INTRODUCTION

In the context of ultrawideband (UWB) communications, a variety of transceiver schemes have been proposed. While the multi-band (OFDM) solution has been most widely adopted as a natural evolution of existing technology, a potentially simpler solution is offered by Transmit Reference (TR) impulse-based systems, studied here. As initially proposed by Hoctor and Tomlinson in [1, 2], TR-UWB systems have the advantage of not requiring channel parameter estimation at the nanosecond level. Instead the detection of the transmitted data relies on the autocorrelation with a reference pulse, while multi-user access capabilities are provided by a delay-hopping (DH) code in addition to a code modulation (CDMA) scheme.

The receivers proposed in [1, 2] did not take the delay spread of more realistic FIR channels into account. This extension was made in our preceding paper [3], where a data model for a multi-user delay-hopped transmit-reference (DH-TR) UWB system over a dispersive channel was formulated. These effects were further quantized in [4] for Rayleigh-fading channels with an exponential path delay profile. Furthermore, a similar autocorrelation system was studied in [5, 6] for the case of differential transmissions (at symbol and frame level), with a detailed analysis of inter-frame and inter-symbol interference. Work has started to build implementations of these transceivers, based on CMOS technology [7].

In this paper, we extend this work, and derive a detailed data model for a DH-TR UWB system, including the effect of dispersive channels and additive noise, as processed by the correlator (in previous research, noise was ignored). Based on this, we propose several receiver algorithms, first for the noiseless single-user case (matched filter, blind multi-symbol, and iterative extensions of both), and subsequently for the complete data model. The proposed algorithms are blind or semi-blind: the channel parameters (in this case correlations) are estimated along with the data. Section 5 shows the simulated performance of the algorithms.

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2. TRANSMIT-REFERENCE DATA MODEL

2.1. Analog receiver model

We consider a single-user DH TR system as in [2] where narrow pulses $g(t)$ are transmitted in pairs (doublets) $d(t)$, spaced by varying time intervals of duration D_i. In a doublet, the first pulse is fixed, the second is modulated by the chip value $c \in \{+1, -1\}$, $d(t) = g(t) + c \cdot g(t - D_i)$. A chip consists of N_d identical doublets, spaced by T_d. A sequence of N_c chips $[c_1, \cdots, c_{N_c}]$ forms a code vector for a symbol of duration $T_s = N_c T_c$, where $T_c = N_d T_d$ is the chip duration, as depicted in figure $1(a)$.

The signal propagates through a radio channel with impulse response $h_p(t)$, and at the receiver it is passed through a bank of M correlators, each correlating the signal with a delayed version of itself at lags D_m, $m = 1, \cdots, M$. Subsequently, the outputs of the correlators are integrated over a sliding window of duration $W = T_c$, as in figure $1(b)$.

Consider the output signal of an integrator due to a single doublet at transmitted lag D_i and received lag D_m. Assuming $W \gg T_h$, where T_h is the effective length of the impulse reponse, it can be modeled as [3]

$$x_m(t) = b(t)(\alpha_{mi} c + \beta_{mi}) \quad (1)$$

where $b(t)$ is a "brick" function (equal to 1 between 0 and W, and zero elsewhere), and α_{mi}, β_{mi} are unknown channel parameters. Since α_{mi} is associated with the modulated data c, we may regard it as a gain, while β_{mi} is an offset. For a stationary channel, we can derive that

$$\alpha_{mi} = \rho(D_i - D_m) + \rho(D_i + D_m) \quad (2)$$
$$\beta_{mi} = 2\rho(D_m) \quad (3)$$

where $\rho(\Delta)$ is the channel auto-correlation function,

$$\rho(\Delta) = \int_{-\infty}^{\infty} h(t)h(t - \Delta)dt \quad (4)$$

and $h(t) = g(t) * h_p(t)$. For a chip consisting of N_d identical doublets spaced by T_d, the model becomes

$$x_m(t) = p(t)(\alpha_{mi} c + \beta_{mi}) \quad (5)$$

where the "tent" function $p(t)$ has a staircase triangular shape with support on $0 \le t \le 2T_c$:

$$p(t) = \sum_{k=0}^{N_d - 1} b(t - kT_d)$$

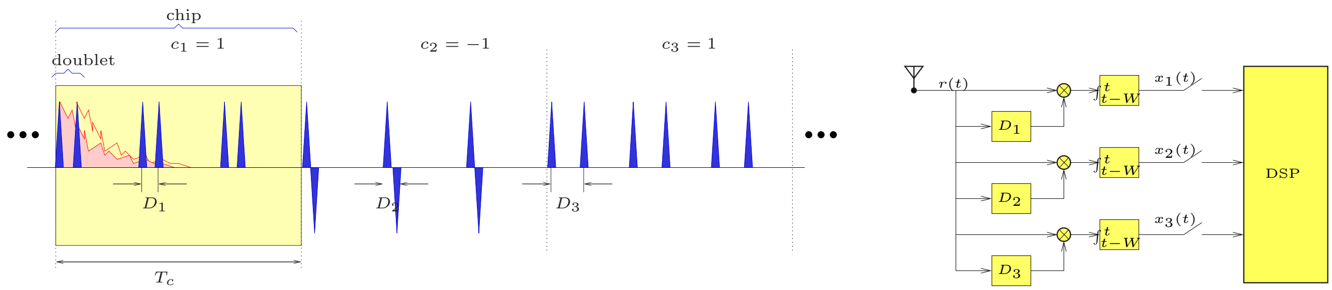

Figure 1. (*a*) Structure of the transmitted data burst, (*b*) Structure of the auto-correlation receiver.

2.2. Matrix formulation

Consider the transmission of N_c consecutive chips $\mathbf{c} = [c_1 \cdots c_{N_c}]^T$ for a single symbol s. Each chip is transmitted using one of the delays $D_1, \cdots, D_M$ and is received using a bank of M correlators at delays $D_1, \cdots, D_M$.

Based on the channel coefficients α_{mi} and β_{mi}, we define the channel matrices $\mathbf{A} = [\alpha_{mi}]$ and $\mathbf{B} = [\beta_{mi}]$ of size $M \times M$. From (3), we know that the offset coefficients β_{mi} are independent of transmitter delay, i.e., the i index. Therefore, all columns of $\mathbf{B}$ are identical. Similarly, it is observed that $\mathbf{A}$ is symmetric, thus

$$\mathbf{B} = \mathbf{b}\mathbf{1}_M^T \,, \qquad \mathbf{A} = \mathbf{A}^T \,. \tag{6}$$

Additionally, a "code delay" or selector matrix $\mathbf{J} = [J_{ij}]$ of size $M \times N_c$ is defined as

$$J_{ij} = \begin{cases} 1, & \text{if chip } j \text{ is transmitted at delay } D_i \\ 0, & \text{elsewhere} \end{cases} \tag{7}$$

Note that matrix $\mathbf{J}$ has for each column only one nonzero entry, corresponding to the transmitted delay index. Therefore, $\mathbf{J}^T\mathbf{1}_M = \mathbf{1}_{N_c}$. From (5), the received signal for a single symbol can be written as

$$x_m(t) = \sum_{i=1}^{M}\sum_{j=1}^{N_c} p(t - jT_c)(\alpha_{mi}J_{ij}c_j + \beta_{mi}J_{ij}) \,. \tag{8}$$

Assume that the outputs of the integrators are sampled at P times the chip rate, where P is the oversampling rate (typically $P = 2$). The sampled data at the instances $t = k\frac{T_c}{P}$ is given by

$$x_{mk} = \sum_{i=1}^{M}\sum_{j=1}^{N_c} p_{kj}(\alpha_{mi}J_{ij}c_j + \beta_{mi}J_{ij})$$

where $p_{kj} = p(t - jT_c)\,|_{t=kT_c/P}$. Define $\mathbf{P} = [p_{kj}]$, where $k = 0, \cdots N-1$ and $j = 1, \cdots, N_c$. Collecting the temporal samples into a vector, we obtain the model

$$\begin{aligned} \mathbf{x}_m &= \sum_{i=1}^{M}\sum_{j=1}^{N_c} \mathbf{p}_j[\alpha_{mi}J_{ij}c_j + \beta_{mi}J_{ij}] \\ &= \sum_{j=1}^{N_c} \mathbf{p}_j[\mathbf{a}_m^T\mathbf{j}_jc_j + \mathbf{b}_m^T\mathbf{j}_j] \\ &= \sum_{j=1}^{N_c}[\mathbf{p}_jc_j\mathbf{j}_j^T\mathbf{a}_m + \mathbf{p}_j\mathbf{j}_j^T\mathbf{b}_m] \\ &= [\mathbf{p}_1c_1, \cdots, \mathbf{p}_{N_c}c_{N_c}]\mathbf{J}^T\mathbf{a}_m + [\mathbf{p}_1, \cdots, \mathbf{p}_{N_c}]\mathbf{J}^T\mathbf{b}_m \\ &= \mathbf{P}\text{diag}(\mathbf{c})\mathbf{J}^T\mathbf{a}_m + \mathbf{P}\mathbf{J}^T\mathbf{b}_m \end{aligned} \tag{9}$$

where $\mathbf{a}_m^T$ and $\mathbf{b}_m^T$ are the m-th rows of $\mathbf{A}$ and $\mathbf{B}$ matrices, and $\mathbf{p}_j$ and $\mathbf{j}_j$ are the j-th columns of $\mathbf{P}$ and $\mathbf{J}$, respectively.

Collecting all vectors $\mathbf{x}_m$ into a matrix $\mathbf{X} = [\mathbf{x}_1, \cdots, \mathbf{x}_M]$

$$\mathbf{X} = \mathbf{P}\text{diag}(\mathbf{c})\mathbf{J}^T\mathbf{A}^T + \mathbf{P}\mathbf{J}^T\mathbf{B}^T \,.$$

Finally, if we transmit multiple symbols $\mathbf{s} = [s_1, \cdots, s_{N_s}]^T$, and ignore the overlaps between consecutive symbols due to ISI (can be up to 1 chip duration), we have for the n-th symbol

$$\begin{aligned} \mathbf{X}_n &= \mathbf{P}\text{diag}(\mathbf{c})\mathbf{J}^T\mathbf{A}^Ts_n + \mathbf{P}\mathbf{J}^T\mathbf{B}^T \\ &= \mathbf{P}[\text{diag}(\mathbf{c})\mathbf{J}^T \quad \mathbf{1}_{N_c}][\mathbf{A}s_n \quad \mathbf{b}]^T \end{aligned} \tag{10}$$

The absence of ISI is a reasonable assumption in case we are synchronized at the chip level: in that case the rows of $\mathbf{X}_n$ affected by ISI are simply dropped. The more general case (no synchronization, and ISI taken into account) is studied in [8].

3. RECEIVER ALGORITHMS IGNORING NOISE TERMS

In the data model (10), $\mathbf{J}$, $\mathbf{P}$ and $\mathbf{c}$ are known, where $\mathbf{J}$ is the transmitted "code delay" matrix at chip level, $\mathbf{P}$ depends on the integration method and the distribution of doublets in each chip, and $\mathbf{c}$ is the user's modulation code. The problem is, given the received signal $\mathbf{X}_n$, to estimate data symbol s_n along with the unknown "channel" matrices $\mathbf{A}$ and $\mathbf{B} = \mathbf{b}\mathbf{1}_M^T$.

3.1. Matched filter receiver

A simple receiver can be derived if we assume that the channel does not have temporal correlations. The channel matrices, thus, will be

$$\mathbf{A} = \alpha\mathbf{I} \,, \qquad \mathbf{B} = \mathbf{0}$$

where $\alpha > 0$ is the only unknown constant (the channel power). The simplified data model is

$$\mathbf{X}_n = \mathbf{P}\text{diag}(\mathbf{c})\mathbf{J}^T\alpha s_n \,, \tag{11}$$

and the matched filter receiver is

$$\alpha\hat{s}_n = \text{tr}[\mathbf{J}\text{diag}(\mathbf{c})\mathbf{P}^T\mathbf{X}_n] \,, \tag{12}$$

where tr is the trace operator. Since α is always positive, it does not change the result for BPSK modulations or differential modulations, and, thus, does not need to be estimated.

3.2. Blind multiple symbol receiver

If $\mathbf{A}$ and $\mathbf{B}$ are unknown, they can be estimated along with the data $\mathbf{s} = [s_1, \cdots, s_{N_s}]^T$ in a blind scheme as follows. Write the model as

$$[\mathbf{X}_1\; \mathbf{X}_2 \cdots \mathbf{X}_{N_s}] = \mathbf{P}[\text{diag}(\mathbf{c})\mathbf{J}^T\; \mathbf{1}] \begin{bmatrix} \mathbf{A}^Ts_1 & \mathbf{A}^Ts_2 & \cdots & \mathbf{A}^Ts_{N_s} \\ \mathbf{b}^T & \mathbf{b}^T & \cdots & \mathbf{b}^T \end{bmatrix}$$

or, with simplified notation,

$$\mathbf{X} = \mathbf{Q}\begin{bmatrix} \mathbf{A}^T s_1 \, \mathbf{A}^T s_2 \cdots \, \mathbf{A}^T s_{N_s} \\ \mathbf{b}^T \quad\; \mathbf{b}^T \quad \cdots \, \mathbf{b}^T \end{bmatrix} \tag{13}$$

where $\mathbf{Q}$ is known, and $\mathbf{A}$, $\mathbf{b}$, $\mathbf{s}$ are unknown. We can solve this equation by first pre-multiplying both sides with the left pseudo-inverse of $\mathbf{Q}$,

$$\mathbf{Y} := \mathbf{Q}^{\dagger}\mathbf{X} = \begin{bmatrix} \mathbf{A}^T s_1 \, \mathbf{A}^T s_2 \cdots \, \mathbf{A}^T s_{N_s} \\ \mathbf{b}^T \quad\; \mathbf{b}^T \quad \cdots \, \mathbf{b}^T \end{bmatrix} \tag{14}$$

Partition matrix $\mathbf{Y}$ into n sub-matrices $\mathbf{Y}_i$ of size $(M+1)\times M$. The offset $\mathbf{b}$ can be estimated from the last row of $\mathbf{Y}$ as

$$\mathbf{b}^T = \frac{1}{n}\sum_{i=1}^{n} \mathbf{Y}_i(M+1,:)$$

where $\mathbf{Y}_i(M+1,:)$ is the $(M+1)$-th row of the matrix $\mathbf{Y}_i$.

To estimate $\mathbf{A}$ and $\mathbf{s}$, we unstack all the matrices $\mathbf{Y}_i(1:M,:)$ into vectors $\mathbf{y}_i$, and define $\mathbf{Y}' = [\mathbf{y}_1, \cdots, \mathbf{y}_{N_s}]$. This matrix has the model

$$\mathbf{Y}' = \mathrm{vec}(\mathbf{A}^T)\mathbf{s}^T\,. \tag{15}$$

Hence, the source symbol vector $\mathbf{s}$ and channel matrix $\mathbf{A}$ can be estimated up to a scaling by computing a rank-1 decomposition (SVD) of $\mathbf{Y}'$.

3.3. Iterative estimation receiver

In the preceding receiver algorithm, the inversion of $\mathbf{Q}$ may be undesirable (e.g., it may color the noise). Improved performance can be obtained by a two-step iterative receiver which is initialized by the preceding one: (1) assume $\mathbf{s}$ is known, and estimate $\mathbf{A}$, $\mathbf{b}$; (2) assume $\mathbf{A}, \mathbf{b}$ are known, estimate $\mathbf{s}$. For the first step, we rewrite the data model (13) as

$$\begin{bmatrix} \mathbf{X}_1 \\ \vdots \\ \mathbf{X}_{N_s} \end{bmatrix} = \begin{bmatrix} \mathbf{P}[s_1 \mathrm{diag}(\mathbf{c})\mathbf{J}^T \;\; \mathbf{1}] \\ \vdots \\ \mathbf{P}[s_{N_s} \mathrm{diag}(\mathbf{c})\mathbf{J}^T \;\; \mathbf{1}] \end{bmatrix} \begin{bmatrix} \mathbf{A}^T \\ \mathbf{b}^T \end{bmatrix} \tag{16}$$

from which $\mathbf{A}$, $\mathbf{b}$ can be estimated using least squares. For the second step, we partition $\mathbf{Q}$ in (13) as $\mathbf{Q} = [\mathbf{Q}' \quad \mathbf{q}]$ and obtain

$$\mathrm{vec}(\mathbf{X}_n) = \mathrm{vec}(\mathbf{Q}'\mathbf{A}^T)\mathbf{s}_n \; + \; \mathbf{b}\otimes\mathbf{q} \tag{17}$$

Therefore, a LS solution is

$$\hat{s}_n = [\mathrm{vec}(\mathbf{Q}'\mathbf{A}^T)]^{\dagger}(\mathrm{vec}(\mathbf{X}_n) - \mathbf{b}\otimes\mathbf{q})$$

which is straightforward to evaluate.

4. DATA MODEL WITH NOISE TERMS

4.1. Single doublet

We will now analyze the effect of noise on the sampled signal after integration. Due to the correlation, it has two components: a cross-term with the data, and a cross-term with itself. We first look at the single doublet case, for which we only give the main results; the derivations are omitted due to lack of space.

For one doublet in additive white Gaussian noise with variance $\sigma^2 = \frac{N_0}{2}$, the resulting data model is found to be

$$x_m(t) \approx b(t)\{(\beta_m + \gamma'_m) \; + \; c\,(\alpha_{mi} + \gamma''_{mi})\} \; + \; n_2(t) \tag{18}$$

where $b(t)$ is the brick function, c is the chip value, and α_{mi}, β_m are as in (2), (3). These parameters only depend on the channel correlation function $\rho(\cdot)$ and are constant over multiple doublets.

The first noise term in (18) is due to the correlation of the received pulses with the noise. It gives rise to perturbations on α_{mi}, β_m,* and is given by

$$n_1(t) \approx b(t)(\gamma'_m + c\,\gamma''_{mi}) \tag{19}$$

where $b(t)$ is the brick function, and γ'_m and γ''_{mi} are random variables, selected once for each doublet and different for the next doublet. We can derive that

$$\begin{aligned} \mathrm{E}(\gamma'_m) \quad &= \mathrm{E}(\gamma''_{mi}) \quad = 0 \\ \mathrm{E}(|\gamma'_m|^2) \;\; &= \mathrm{E}(|\gamma''_{mi}|^2) = 2\sigma^2[\rho(0) + \rho(2D_m)] \\ \mathrm{E}(\gamma'_m\gamma''_{mi}) &= \sigma^2[2\rho(D_i) + \rho(2D_m - D_i) + \rho(2D_m + D_i)] \end{aligned} \tag{20}$$

Thus, the variance of these variables depends both on the noise power σ^2 and the channel correlation function $\rho(\cdot)$, but not on the integration length W. Note that the variance depends on $\rho(0)$, which is always positive and the corresponds to the maximum of the correlation function. In contrast, α_{mi} depends on $\rho(D_i - D_m)$, which is maximum only for $D_m = D_i$.

The second noise term is caused by the correlation of the noise with itself, i.e.,

$$n_2(t) := \int_{t-W}^{t} n(\tau)n(\tau - D_m)\,d\tau\,. \tag{21}$$

When we assume white Gaussian noise with noise density N_0, and a prefilter with a processing bandwidth B, the expected value and the variance of this term are

$$\mathrm{E}(n_2(t)) = r_{nn}(D_m)\cdot W \approx 0 \tag{22}$$

$$\mathrm{var}(n_2(t)) = \frac{N_0^2 BW}{2} \tag{23}$$

where W is the integration length, $r_{nn}(\cdot)$ is the autocorrelation function of noise.

4.2. Matrix formulation

Following the same derivation as in section 2.2, the sampled integrated data $x_{mk} = x_m(k\frac{T_c}{P})$ due to a single symbol (N_c chips $\{c_j\}$, one doublet per chip) has the model

$$x_{mk} = \sum_{i=1}^{M}\sum_{j=1}^{N_c} p_{kj}[(\alpha_{mi}+\gamma''_{mij})J_{ij}c_j + (\beta_m+\gamma'_{mj})J_{ij}] + (n_2)_{mk}$$

Note that the noise terms γ'_{mij} and γ''_{mj} change from chip to chip. As before, we collect N samples x_{mk} into a vector $\mathbf{x}_m$.

To simplify the derivation, we first consider only the "gain" part of the channel (omitting the "offset" and the noise n_2). A derivation similar to (9) shows that

$$\begin{aligned} \mathbf{x}_{m,\mathrm{gain}} &= \mathbf{P}\mathrm{diag}(\mathbf{c})\mathbf{J}^T\mathbf{a}_m + \mathbf{P}\mathrm{diag}(\mathbf{c})\mathbf{J}'^T\boldsymbol{\gamma}''_m \\ &=: \mathbf{P}\mathrm{diag}(\mathbf{c})(\tilde{\mathbf{a}}_m + \tilde{\boldsymbol{\gamma}}''_m) \end{aligned}$$

where $\mathbf{a}_m^T$ is the m-th row of $\mathbf{A}$, $\boldsymbol{\gamma}''_{mj}$ is a stacking of $\{\gamma''_{mij}, i = 1,\cdots,M\}$, $\boldsymbol{\gamma}''_m$ is the stacking of all vectors $\{\boldsymbol{\gamma}''_{mj}, j = 1,\cdots,N_c\}$ and $\mathbf{J}' = \mathbf{J}\circ\mathbf{I}_{N_c}$. Remind that $\mathbf{J}$ is only the "code delay" selection matrix (each column has exactly one nonzero entry), therefore the "signal" term $\tilde{\mathbf{a}}_m := \mathbf{J}^T\mathbf{a}_m$ can be considered as a vector of which the entries are mapped from $\mathbf{a}_m$. Similarly, the "noise" term $\tilde{\boldsymbol{\gamma}}''_m = \mathbf{J}'^T\boldsymbol{\gamma}''_m$ can also be considered as a vector of which

*We could thus consider these parameters as random variables with nonzero mean and which are selected once for each transmitted doublet.

the j-th element is mapped from γ''_{mj}. The noise term has a known covariance, with uncorrelated entries:

$$\mathrm{E}\{\tilde{\gamma}''_m \tilde{\gamma}''^T_m\} = \sigma''^2_m \mathbf{I}\,,$$

where σ''^2 can be computed using (20).

For the offset part we follow the same procedures, and note that β_m and γ'_{mj} do not depend on index i, and $\sum_{i=1}^{M} J_{ij} = 1$ for all j. Therefore we obtain

$$\mathbf{x}_{m,\text{offset}} = \mathbf{P}\mathbf{1}_{N_c}\beta_m + \mathbf{P}\boldsymbol{\gamma}'_m \tag{24}$$

where $\boldsymbol{\gamma}'_m$ is a stacking of $\{\gamma'_{mj}, j = 1, \cdots, N_c\}$. Note that $\mathbf{J}$ is removed from (24) because the offset term does not depend on the transmitted delays. Finally, collecting the gain and offset parts, stacking all samples $\mathbf{x}_m$ into a matrix, and taking the symbol value s_n into account, we obtain the complete data model as follows:

$$\mathbf{X}_n = \mathbf{P}\mathrm{diag}(\mathbf{c})(\tilde{\mathbf{A}} + \tilde{\mathbf{\Gamma}}''_n)s_n + \mathbf{P}(\tilde{\mathbf{B}} + \tilde{\mathbf{\Gamma}}'_n) + \mathbf{N}_2 \tag{25}$$

where

$$\begin{aligned} \tilde{\mathbf{A}} &= \mathbf{J}^T\mathbf{A}^T\,, & \tilde{\mathbf{B}} &= \mathbf{1}_{N_c}\mathbf{b}^T \\ \tilde{\mathbf{\Gamma}}'' &= [\tilde{\gamma}''_1 \;\cdots\; \tilde{\gamma}''_M]\,, & \tilde{\mathbf{\Gamma}}' &= [\gamma'_1 \;\cdots\; \gamma'_M] \end{aligned}$$

In this model, $\mathbf{P}$, $\tilde{\mathbf{A}}$, $\tilde{\mathbf{B}}$ depend on the channel correlation, $\mathbf{c}$ is known, and the noise terms have the following properties:

$$\mathrm{E}\{\tilde{\mathbf{\Gamma}}'\} = \mathrm{E}\{\tilde{\mathbf{\Gamma}}''\} = 0$$

$$\mathrm{E}\{\tilde{\mathbf{\Gamma}}''\tilde{\mathbf{\Gamma}}''^T\} = \sum_{m=1}^{M} \mathrm{E}\{\tilde{\gamma}''_m\tilde{\gamma}''^T_m\} = \mathbf{I}\sum_{m=1}^{M}\sigma''^2_m \tag{26}$$

and

$$\mathrm{E}\{\tilde{\mathbf{\Gamma}}'\tilde{\mathbf{\Gamma}}'^T\} = \sum_{m=1}^{M} \mathrm{E}\{\tilde{\gamma}'_m\tilde{\gamma}'^T_m\} = \mathbf{I}\sum_{m=1}^{M}\sigma'^2_m$$

4.3. A receiver algorithm with noise terms

We briefly indicate a training-based receiver algorithm based on the data model for the single doublet case is derived. Assuming a single data symbol s_k is known, the data model in (25) can be expressed as

$$\mathbf{X}_k = \mathbf{P}\mathrm{diag}(\mathbf{c})\tilde{\mathbf{A}}s_k + \mathbf{P}\tilde{\mathbf{B}} + \tilde{\mathbf{N}} \tag{27}$$

In this equation, we have an aggregate noise term $\tilde{\mathbf{N}} := (\mathbf{P}\mathrm{diag}(\mathbf{c})s_k\tilde{\mathbf{\Gamma}}'' + \mathbf{P}\tilde{\mathbf{\Gamma}}'_k + \mathbf{N}_2)$ with known covariance $\tilde{\mathbf{R}}$ (scaled by an unknown constant), which can be derived straightforwardly from (26). To take the noise into account, a receiver algorithm can first whiten it, i.e., pre-multiply both sides of (27) with $\tilde{\mathbf{R}}^{-\frac{1}{2}}$. Subsequently, the two-step iteration described in section 3.3 can be applied.

5. SIMULATION

We simulate the transmission of $N_s = 20$ symbols over an exponentially decaying channel with $M = 4$ delay positions, $N_c = 10$ chips per symbol, single doublet per chip, and $P = 4$ times oversampling at the integrators' outputs. The transmitted pulses duration is 0.5 ns. Two pulses in a doublet are separated by 1, 2, 3 or 4 ns, the doublets are separated by $T_d = 50$ ns.

Due to lack of space, we only show the performance of the receiver algorithms that ignore the noise terms (section 3): the matched filter receiver, the blind multi-symbol receiver and the iterative receivers initialized by one of the previous algorithms. BER vs. SNR plots of these algorithms are shown in Fig. 2. It can be seen that the iterative algorithms give significant performance improvement over both proposed non-iterative algorithms. The gain becomes arbitrary large when SNR increases.

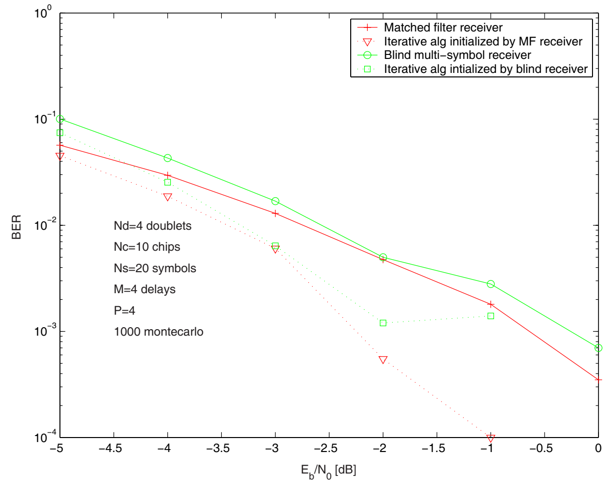

Figure 2. BER vs. SNR for different receiver algorithms

References

[1] N. van Stralen, A. Dentinger, K. Welles, R. Gauss, R. Hoctor, and H. Tomlinson, "Dealy hopped transmitted reference experimental results," in *IEEE Conference on Ultra Wideband Systems and Technologies*, pp. 93–98, 2002.

[2] R. Hoctor and H. Tomlinson, "Delay-hopped transmitted-reference RF communications," in *IEEE Conference on Ultra Wideband Systems and Technologies*, pp. 265–270, 2002.

[3] A. Trindade, Q. Dang, and A. van der Veen, "Signal processing model for a transmit-reference UWB wireless communication system," in *IEEE Conf. on Ultra Wideband Systems and Technologies (UWBST2003)*, (Reston, Virginia), Oct. 2003.

[4] K. Witrisal, M. Pausini, and A. Trindade, "Multiuser interference and inter-frame interference in UWB transmitted reference systems," in *Int. Workshop on Ultra Wideband Systems (IWUWBS 2004/UWBST2004)*, (Kyoto, Japan), May 2004.

[5] M. Pausini and G. Janssen, "Analysis and comparison of autocorrelation receivers for IR-UWB signals based on differential detection," in *IEEE ICASSP*, (Montreal, Canada), May 2004.

[6] M. Pausini, G. Janssen, and K. Witrisal, "Analysis of ISI for an IR-UWB differential autocorrelation receiver," in *IEEE VTC-Fall*, Oct. 2004.

[7] S. B. S. Lee and W. Serdijn, "A quadrature downconversion autocorrelation receiver architecture for UWB," in *Int. Workshop on Ultra Wideband Systems (IWUWBS2004/UWBST2004)*, (Kyoto, Japan), May 2004.

[8] R. Djapic, A. van der Veen, and G. Leus, "Blind synchronization in asynchronous multiuser UWB networks based on the transmit-reference scheme," in *Asilomar Conference on Signals, Systems, and Computers*, Nov. 2004.