Considering delay inaccuracies in a transmit-reference UWB communication system

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Abstract— **A Delay-Hopped Transmitted-Reference Ultra-Wideband (UWB) communication system is studied and further developed. We previously proposed an accurate signal processing model and several receiver algorithms. This paper emphasizes on the importance of including the channel parameters in the data model for more general and realistic channels, which will result in more flexible and robust receiver algorithms. Various channel correlation measurement results are shown, and system performance is tested with measured channels under different practical situations. The receiver algorithm is shown to be robust under the common situation where the receiver delays (correlation lags) have a certain offset compared to the transmit delays (spacing between two pulses in a doublet).**

I. Introduction

The classical Transmit-Reference (TR) scheme is considered a promising candidate for practical UWB communication systems. By transmitting the reference signal through the same channel as the information-bearing signal, it allows the data symbols to be detected without directly estimating the channel multipath coefficients (which would be considered unrealistic for UWB channels). The scheme was introduced for UWB by Hoctor and Tomlinson [2], in a form where pulses are transmitted in pairs (doublets), where the first is a reference pulse and serves as a "dirty template" in a matched filter for the second (data) pulse. Some disadvantages of the TR scheme are a 3-dB loss in signal-to-noise ratio (SNR) due to the correlation with a noisy signal, and an obvious loss in transmission capacity due to the repeated transmission of reference pulses. It is possible to improve on these aspects by averaging several reference pulses over time, and by re-using this template for estimating the message in several information-bearing pulses [3]. Alternatively, we can implement a weighting scheme where we integrate only over periods where the reference pulse is present [4].

To avoid inter-pulse interference, most of the research published on TR UWB systems so far assumes either a very short channel length or a large spacing between the reference and the information pulses, which greatly reduces the data rate of the system. In our proposed system, we consider that the spacing between pulses in a doublet (a pair of two pulses) can be much smaller than the channel length. This introduces additional correlations also for non-matched delays at the receiver. Taking this into account, we have derived an accurate signal processing model and corresponding receiver algorithms [5], [6]. The present paper expands upon this work, by demonstrating the improved performance of the receivers over actual measured channels, in the practical case where there is a small offset (200 ps) in delays between the transmitted pulse pairs and the receiver correlation lags. This is in particular relevant when the channel is non-line of sight (NLOS) with correlated taps.

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II. Data model for a DH TR system

We consider a single-user delay-hopped transmit-reference (DH TR) system, where each transmitted symbol consists of N_c chips, each of duration T_c.

In the scheme depicted in figure $1(a)$, for each chip c_j a pair of pulses $g(t)$ is transmitted. The first pulse is the reference, the second pulse has information of the chip value on its polarity. Two pulses are spaced by a time interval of duration D_i, which is selected among a delay set $\{D_1, \dots, D_M\}$ and represents a user-specific delay code. The transmitted pulses for the j-th chip can be expressed as

$$c_j(t) = g(t - jT_c) + c_j\, g(t - jT_c - D_i). \tag{1}$$

This pulse pair (doublet) is propagated through a radio channel $h(t)$, where $h(t) = h_p(t) * g(t)$ is the convolution between the monopulse and physical channel of duration T_h. The received signal will be

$$r_j(t) = h(t - jT_c) + c_j\, h(t - jT_c - D_i). \tag{2}$$

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 $W \geq T_c$, as in figure $1(b)$. The output of the m-th correlator and integrator branch for the received signal (2) can then be written as

$$x_{m,j}(t) = p(t - jT_c)(c_j\, \alpha_{m,i} + \beta_m), \tag{3}$$

where $p(t)$ is a "brick" function (equal to 1 between 0 and W, and zero elsewhere), and $\alpha_{m,i}$ and β_m are the channel coefficients, which can be derived from the channel auto-correlation function $\rho(\Delta) = \int_{-\infty}^{\infty} h(t)h(t-\Delta)dt$ as

$$\begin{aligned} \alpha_{m,i} &= \rho(D_m - D_i) + \rho(D_m + D_i), \\ \beta_m &= 2\rho(D_m). \end{aligned} \tag{4}$$

As in [5], [6], we can derive a matrix-based model for the transmission of N_c consecutive chips for a single symbol s by defining channel matrices $\mathbf{A} = [\alpha_{mi}]_{M\times M}$, $\mathbf{b} = [\beta_m]_{M\times 1}$, the polarity code vector $\mathbf{c} = [c_j]$, and the delay code selector matrix $\mathbf{J} = [J_{ij}]_{M\times N_c}$ of which each column has one nonzero entry corresponding to the transmitted delay index. The vector $\mathbf{x}_m$ which collects all samples at received delay D_m corresponding to one symbol period will have model

$$\mathbf{x}_m = \mathbf{P}\text{diag}(\mathbf{c})\mathbf{J}^T\mathbf{a}_m s + \mathbf{P}\mathbf{1}_{N_c}\beta_m$$

where $\mathbf{P}$ is a Sylvester matrix stacking all the shifted versions of the samples of the "pulse shape function" $p(t)$, and $\mathbf{a}_m$ is the

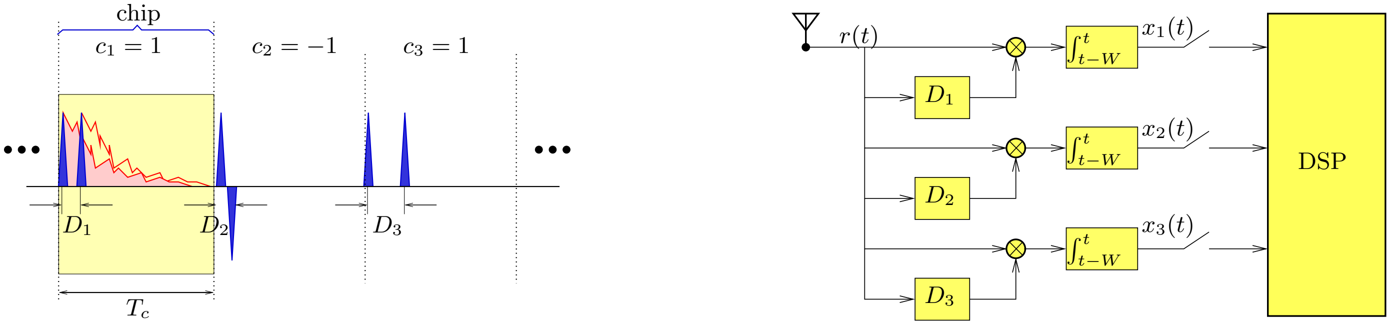

Fig. 1. (*a*) Structure of the transmitted data burst, (*b*) Structure of the auto-correlation receiver.

m-th column of $\mathbf{A}^T$. Collecting all vectors $\mathbf{x}_m$ into a matrix $\mathbf{X} = [\mathbf{x}_1, \ldots, \mathbf{x}_M]$, we have for a single symbol

$$\mathbf{X} = \mathbf{P}\text{diag}(\mathbf{c})\mathbf{J}^T\mathbf{A}^T s + \mathbf{P}\mathbf{1}_{N_c}\mathbf{b}^T .$$

When transmitting multiple symbols $\mathbf{s} = [s_0, \cdots, s_{N_s-1}]^T$, and assuming there is no overlap between consecutive symbols, the model for the k-th symbol is

$$\begin{aligned}\mathbf{X}_k &= \mathbf{P}\text{diag}(\mathbf{c})\mathbf{J}^T\mathbf{A}^T s_k + \mathbf{P}\mathbf{1}_{N_c}\mathbf{b}^T \\ &= \mathbf{P}[\text{diag}(\mathbf{c})\mathbf{J}^T, \mathbf{1}_{N_c}][\mathbf{A}s_k, \mathbf{b}]^T . \end{aligned} \quad (5)$$

In this model, $\mathbf{X}_k$ contains the measured samples, $\mathbf{c}$ is known (user code), $\mathbf{J}$ is known (delay code), and $\mathbf{P}$ is known while $\mathbf{A}$ and $\mathbf{b}$ are unknown (channel correlation coefficients), and s_k is the data symbol to be detected.

A more advanced data model that includes the effect of noise terms can be found in [1]. We skip the details due to lack of space.

III. Channel statistics

In the data model derived in section II, the channel coefficients α_{mi} and β_m are unknown and normally need to be estimated along with the user's data. Simple receivers assume that $\alpha_{mi} = \alpha\delta_{m,i}$, i.e., there is only a response at matched delays, and $\beta_m = 0$: no voltage offset at the output of the correlator. The question is whether this is a valid assumption. It is, therefore, important to understand the statistics of these coefficients.

A. Statistics of channel coefficients for a theoretical channel model

First, we consider a theoretical multipath channel model, where the physical channel impulse response is modelled as a sum of discrete delta pulses,

$$h_p(t) = \sum_{i=0}^{\infty} a_i \delta(t - \tau_i), \quad (6)$$

where a_i are ray amplitudes, and τ_i are their corresponding arrival times. A typical channel model for UWB is assumed be time-invariant and to have uncorrelated ray amplitudes a_i, where ray amplitudes will be negligibly small for large τ_i.

A detailed analysis can be found in [7], [1]. Omitting the equations, figure 2 shows the resulting expected values and standard deviations of $\rho(\Delta)$ for a Gaussian monocycle (duration 0.2 ns), and a multipath channel with exponentially decaying power delay profile with parameters $P_0 = 1$ (normalized channel power), $K = 0$ (non-line-of-sight channel), exponential decay factor $1/\gamma = 15$ ns, and path arrival rate $\lambda = 5$ ns^{-1}. According to this model, $\rho(\Delta)$ is significant only for $\Delta = 0$, which results in $\alpha_{mi} \approx \alpha\delta_{m,i}$ and gives credibility to the model assumptions considered by Hoctor and Tomlinson [2].

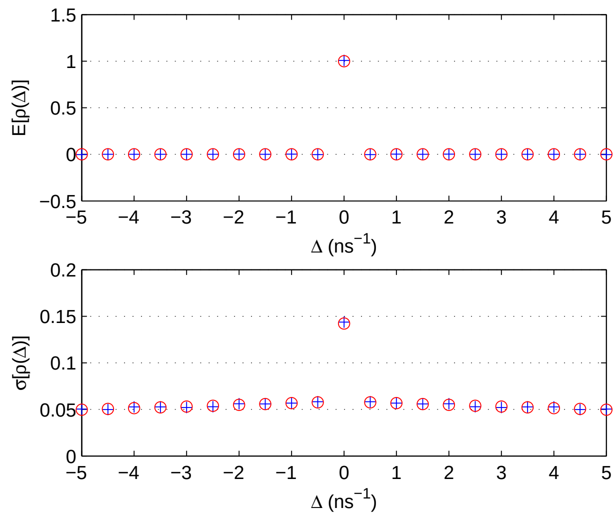

Fig. 2. Statistics of $\rho(\Delta)$ according to the uncorrelated exponentially decaying multipath model (+' denotes a simulated value, 'o' the analytical result).

B. Measured channel correlation coefficients

Within the AIRLINK project at TU Delft, recently the first channel impulse response measurements have been conducted.* An example of impulse response, frequency spectrum and autocorrelation function is shown in figure 3. The measurement data includes the convolution by the pulse shape and the distortion by the biconical antennas. The sampling period is 10 ps, achieved using stroboscopic sampling. However, the effective bandwidth is about 10 GHz, as above this frequency the signal is masked by the noise. The transmitted pulse is about 50 ps, but it is immediately distorted by the antenna to a nonsymmetric monocycle with a duration of about 1.5 ns.

The data includes 7 indoor experiments: four line-of-sight (LOS) at distances of 1.5 to 4 m, two non-line-of-sight (NLOS) from an office to a neighboring office (thin concrete wall), and one NLOS from office to corridor. Table I shows specific values of the auto-correlation function $\rho(\tau)$ for each of the experiments, at a spacing of 0.5 ns.

It is seen from the table that $\rho(0)$ is dominant and typically 3 to 5 times larger than the other values of $\rho(\tau)$. However, the

*We are grateful to Z. Irahhauten, G. Janssen and A. Yarovoy for implementing these experiments. The shown data has been postprocessed: lowpass filtered (to 12.5 GHz), time-shifted, and the DC component was subtracted. Details on the measurements will be published separately at ICU'05. Our main interest here is in the autocorrelation function.

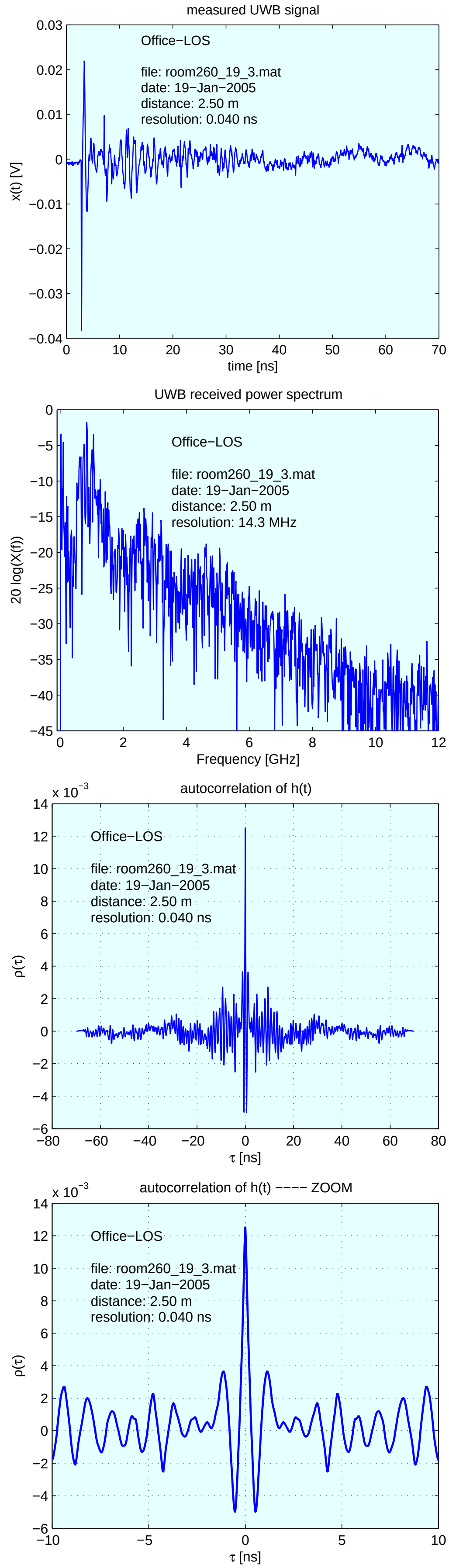

Fig. 3. Measured UWB channel—Office, LOS

correlation peak at 0 is very narrow (about 200 ps). Typical affordable delay lines have tolerances which are higher than this. The values of $\rho(\tau+0.2\text{ ns})$ for each of the experiments are shown in the second column in table I. In this case, the correlation peak is missed, and all values of ρ have about the same magnitude.

This suggests that in practice, most values of α_{mi}, β_m are significant, one cannot assume that $\mathbf{A}$ is diagonally dominant and that $\mathbf{b}$ is zero. The difference with the theoretical model is explained from the fact that it was derived for a pulse duration shorter than the distance between two pulses, which is violated in the measured data due to the spreading by the antenna.

IV. Receiver algorithms and simulation

From the channel statistics described in the previous section, it can be seen that the channel coefficients' statistics can vary for different channel models under different situations. This, in turn, significantly changes the structures and the statistics of channel matrices $\mathbf{A}$ and $\mathbf{b}$. We already developed some receiver algorithms in [6] but did not show their performance. In the following we will briefly introduce these algorithms again and discuss their performance by simulation.

A. Receiver algorithms

The simple matched filter algorithm is based on the assumption that $\rho(0) =: \alpha$ is dominant, so that $\mathbf{A} = \alpha\mathbf{I}$ and $\mathbf{b} = \mathbf{0}$. Thus, the simplified data model is

$$\mathbf{X} = \mathbf{P}\text{diag}(\mathbf{c})\mathbf{J}^T\alpha s$$

and the corresponding receiver is

$$\widehat{\alpha s} \;=\; \text{tr}[\mathbf{J}\text{diag}(\mathbf{c})\mathbf{P}^T\mathbf{X}]$$

where 'tr' is the trace operator. Since α is always positive, it does not change the detected symbol (at least for a BPSK constellation).

The blind multiple symbol receiver [6] can blindly estimate the channel matrices along with the data symbols. Collect data due to N_s symbols:

$$[\mathbf{X}_0,\ldots,\mathbf{X}_{N_s-1}] = \mathbf{P}[\text{diag}(\mathbf{c})\mathbf{J}^T,\mathbf{1}]\begin{bmatrix} \mathbf{A}^T s_0 & \cdots & \mathbf{A}^T s_{N_s-1} \\ \mathbf{b}^T & \cdots & \mathbf{b}^T \end{bmatrix}$$

Note that $\mathbf{Q} := \mathbf{P}[\text{diag}(\mathbf{c})\mathbf{J}^T,\mathbf{1}]$ is completely known. Assuming it is tall, multiplying both sides with the left inverse of $\mathbf{Q}$ gives

$$[\mathbf{Y}_0,\ldots,\mathbf{Y}_{N_s-1}] := \mathbf{Q}^\dagger[\mathbf{X}_0,\ldots,\mathbf{X}_{N_s-1}] = \begin{bmatrix} \mathbf{A}^T s_0 \cdots \mathbf{A}^T s_{N_s-1} \\ \mathbf{b}^T \;\cdots\; \mathbf{b}^T \end{bmatrix}$$

The channel vector $\mathbf{b}$ can be estimated by averaging the last rows of the $\mathbf{Y}_k$, and $\mathbf{A}$ and $\mathbf{s}$ can be estimated up to a scaling from a certain rank-1 decomposition of the remaining data. This algorithm has no specific assumption on the channel matrices $\mathbf{A}$ and $\mathbf{b}$, it is, thus, expected to work for more practical cases where $\mathbf{A}$ and $\mathbf{b}$ can be arbitrary. This receiver algorithm can also be applied when there is only a single received delay branch at a much lower complexity. However, it is obvious that the performance will significantly degrade.

Finally, an iterative algorithm is available to enhance the performance of both previous algorithms [6]. Using one of these algorithms as the initial estimate of the data symbols, we can implement a two-step iterative Least Squares scheme to alternatingly estimate the channel matrices and the data symbols, keeping the other parameters fixed. As we will see, the performance can be greatly improved in this scheme, which is due to the inversion of a much taller matrix (size of $NN_s \times (M+1)$).

TABLE I

MEASURED CHANNEL CORRELATIONS $\rho(\tau + \text{OFFSET})$, NORMALIZED TO $\rho(0)$, FOR 7 CHANNEL REALIZATIONS

	no offset						offset 0.2 ns					
τ [ns]	0	0.5	1.0	1.5	2.0	2.5	0	0.5	1.0	1.5	2.0	2.5
LOS 1	1.000	-0.430	0.236	0.095	-0.100	0.071	0.171	-0.186	0.165	0.093	-0.088	0.096
LOS 2	1.000	-0.346	0.208	0.183	-0.066	0.076	0.198	-0.141	0.255	0.043	0.008	0.122
LOS 3	1.000	-0.380	0.259	0.097	0.036	0.042	0.207	-0.197	0.261	0.019	0.000	0.059
LOS 4	1.000	-0.478	0.422	-0.066	0.056	0.031	0.182	-0.261	0.281	-0.179	0.096	0.058
NLOS 5	1.000	-0.516	0.273	0.053	0.006	0.090	0.167	-0.316	0.368	-0.291	0.219	-0.047
NLOS 6	1.000	-0.376	0.063	0.238	-0.032	-0.100	0.197	-0.266	0.197	0.133	-0.139	0.038
NLOS 7	1.000	-0.100	0.268	0.115	0.086	-0.071	0.383	-0.001	0.204	0.127	0.004	-0.038

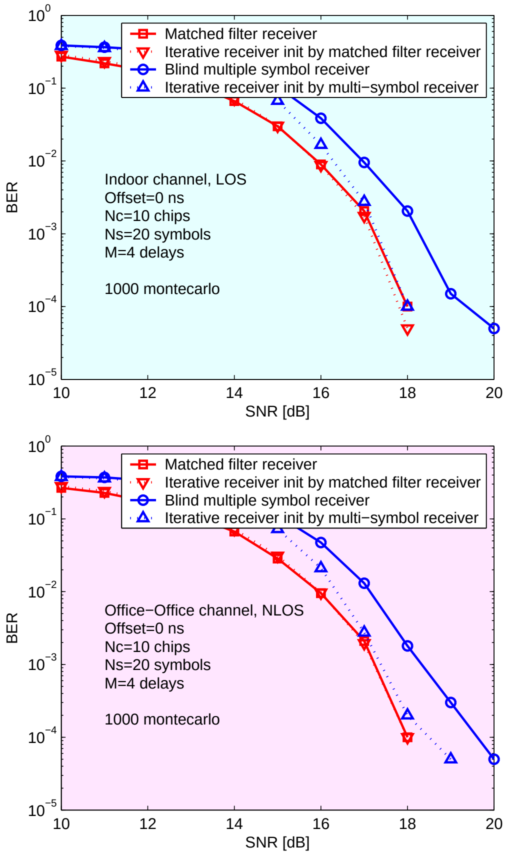

Fig. 4. BER vs. SNR for different receiver algorithms: (a) LOS, (b) NLOS

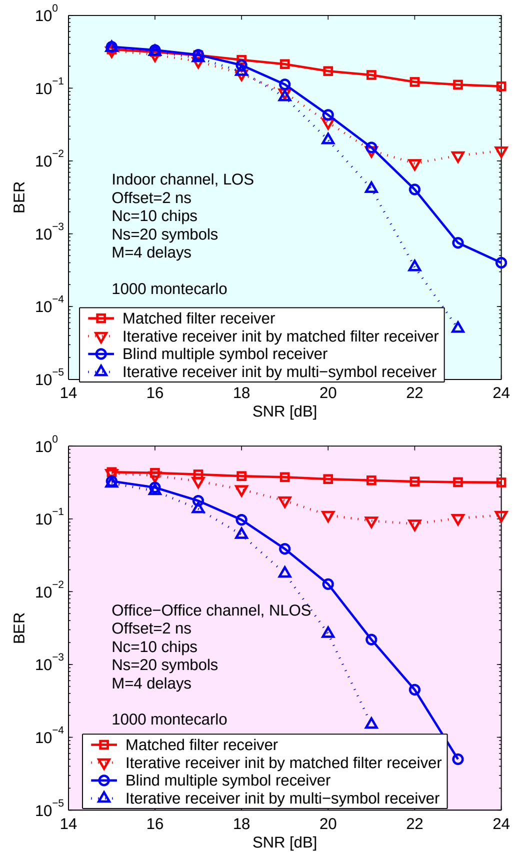

Fig. 5. Receiver performance with receiver delay mismatch (0.2ns): (a) LOS, (b) NLOS

B. Simulation using measured channels

The transmission of $N_s = 20$ symbols is simulated over our measured UWB channels, which were truncated at 50 ns. The system uses $N_c = 10$ chips per symbol, each of duration $T_c = 50$ ns. The two pulses in each chip are separated by $D_m \in \{1, 2, 3, 4\}$ ns. The transmitted pulse is the second derivative of a Gaussian pulse, duration $\tau_m = 0.5$ ns. The integration interval at receive is $W = T_c$. We use 1000 Monte Carlo runs to obtain the BER vs. SNR plots for the various receiver algorithms while the channel is kept fixed. The SNR is defined as the average transmitted pulse energy over the white Gaussian noise density power.

Figure 4 shows the BER versus the SNR for various algorithms in the LOS (office) and NLOS (office-corridor) case. The performance of the simple matched filter and the blind multiple-symbol receiver is about the same, and both can be improved by the iteration (which uses the full data model). The explanation is that in the iterative receiver the estimation of $\mathbf{A}, \mathbf{b}$ avoids the inversion by $\mathbf{Q}$ (a matrix which does not grow with the number of samples) and instead inverts a matrix which does grow with the number of samples and therefore gives less noise enhancement. The same holds for the detection of the symbols.

Figure 5 shows the advantage of the blind multi-symbol receiver over the matched filter receiver for LOS (indoor) and NLOS (office to office) channels when there is a small offset in each received delay bank due to timing inaccuracy, here 0.2 ns. The offset affects the diagonal-dominant structure of the channel matrix $\mathbf{A}$. In this case, the simple matched filter algorithm breaks down while the blind multi-symbol receiver, which takes into account all the elements of matrices $\mathbf{A}$ and $\mathbf{b}$, still maintains a fairly good performance. This effect is even more significant when there is correlation in the multi-path channel, especially in NLOS case. Additional simulations show that when the delay spacings are relatively small, we obtain the same situation where the simple matched filter receiver becomes much worse than the blind multi-symbol receiver.

V. Conclusions

In this paper, an accurate signal processing model was proposed for a transmit-reference UWB system, which considers all relevant channel correlation coefficients. By using different algorithms, we can estimate these coefficients along with data symbols. This guarantees the flexibility and robustness of the system in various situations, e.g., when the channel is NLOS or highly correlated, when there is a small offset in the receiver delay banks due to component inaccuracies, or when the delay spacings are small compared to the channel length. The iterative receiver provides a better performance than the matched-filter and the blind receiver, at a modest increase in complexity.

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