Resolving inter-frame interference in a transmit-reference ultra-wideband communication system

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***Abstract*— Transmit-reference ultra-wideband (TR-UWB) systems are attractive due to their relatively low complexity at both the transmitter and the receiver. Partly, this is achieved by making restrictive assumptions such as a frame length which should be much larger than the channel length. This limits their use to low data rate applications. In this paper, we lift this restriction and allow inter-frame interference (IFI) to occur. We propose a suitable signal processing data model and corresponding receiver algorithms which take the IFI into account. The performance of the algorithms are verified using simulations.**

I. Introduction

Introduced in [1] but based on much older concepts, transmit-reference (TR) schemes are considered a realistic candidate for impulse-based ultra-wideband (UWB) communication systems. Due to the narrow pulses employed in such systems, many multipath components can be resolved, and it is not efficient to estimate all of them. TR schemes sidestep channel estimation by transmitting a reference signal which undergoes the same distortion by the channel as the data signal. This provides an (unfortunately noisy) template that can be used in a matched filter-like receiver.

The basic TR-UWB scheme transmits pulses in pairs (doublets) [1]: a frame constitutes a reference pulse followed by a data pulse, spaced by a certain delay D. Several related TR schemes have been proposed, the differences are e.g., in the way to choose the delay D (perhaps chosen from a set of possible delays, or perhaps more pulses per frame), the use of repeated frames to represent a chip, etc. Similarly, the basic receiver structure consists of a delay, a correlator (multiplier followed by an integrator) to correlate the received signal by the signal delayed by D, and a sampler. Also a variety of receiver structures have been proposed, e.g., using a bank of multiple correlators [2], [3], [4], averaging multiple reference pulse responses into a template that has less noise [5], or splitting up the integration interval into multiple shorter intervals and combining them using weights such as to limit the noise [6], [7].

Most TR-UWB systems assume that, in a frame, pulses are sufficiently widely spaced such that the channel does not introduce inter-pulse interference: $D > T_h$, where T_h is the channel length. Since T_h can be quite long (50 to 200 ns), data rates cannot be very high in such systems, and thus they are considered a candidate for low-rate low-complexity applications.[1] Furthermore, almost all TR-UWB research papers so far have assumed that there is no inter-frame interference: $T_f > T_h + 2D > 3T_h$, where T_f is the frame duration. This assumption again implies that the resulting system will have a low data rate, say 10 to 50 Mbps at most.

In our previous papers [2], [8], we have discussed a TR-UWB scheme that allows $D \ll T_h$, thus lifting the problems associated with implementing long delays, but we still assumed $T_f > T_h$. The inter-pulse interference was resolved in the receiver by employing a bank of correlators (multiple delays), and by combining the resulting samples. By this, we could improve the system performance and approximately double the potential data rate.

[1]Technologically, it is not trivial at all to implement such long wideband delays in an IC.

In the present paper, we aim at further reducing the frame length, so as to enable higher rate UWB applications. In particular, we consider the extreme case where $D \ll T_h$, $T_f < T_h$ and a short integration interval W — integrate and dump is applied at a rate of P samples per frame. The over-sampling rate P can vary depending on the trade-off between the BER performance and the receiver complexity, e.g., the maximum sample rate of the AD converter.

II. Data model

Consider a single user TR-UWB system, where pulses are transmitted in pairs (doublets) within a frame at a high rate, i.e., the frame period is shorter than the channel length, or $T_f < T_h$. Each doublet is associated with a symbol value s_i while the delay between the reference and the information pulse is a fixed period D. Note that we use only one frame per symbol to be able to achieve a high symbol rate, and we use no chip code since we currently consider only a single user system. For multi-user systems, the data model can be easily extended by introducing a CDMA layer with user codes replacing the present symbols.

The pulses are transmitted over a typical UWB indoor environment, i.e., a channel with uncorrelated dense multipath. The received signal at the antenna output can be written as

$$y(t) = \sum_{i=1}^{\infty} h(t-(i-1)T_f) + s_i h(t-(i-1)T_f - D) \quad (1)$$

where $h(t) = h_p(t) * g(t)$ is the convolutional product of the physical channel $h_p(t)$ and the UWB pulse shape $g(t)$, consisting of the transmit pulse including the antenna distortion and lowpass/bandpass filtering). Although most UWB channels are assumed uncorrelated, the distorted pulse shape $g(t)$ can have a duration of several nanoseconds and thus may introduce some correlations in $h(t)$. However, these correlations are usually small compared to the total channel energy and go down quickly as the correlation lag increases.

The signal at the multiplier output, $x(t) = y(t)y(t-D)$ will be integrated over a period $T_{sam} = T_f/P$, where the integer P is the "oversampling rate", and subsequently sampled (integrate-and-dump). Thus, as in [6], [7], we obtain P samples per frame. The autocorrelation receiver scheme is shown in Fig. 1.

A. Single frame

For simplicity, we first consider the resulting signal after correlation due to one transmitted frame only:

$$\begin{aligned} x_0(t) &= [h(t) + s\cdot h(t-D)][h(t-D) + s\cdot h(t-2D)] \\ &= [h^2(t-D) + h(t)h(t-2D)]s + \\ &\quad +[h(t)h(t-D) + h(t-D)h(t-2D)]\,. \end{aligned} \quad (2)$$

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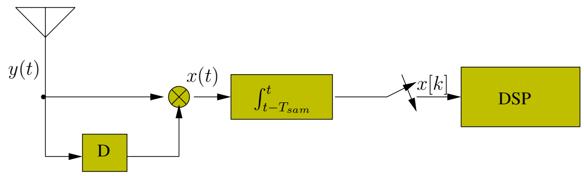

Fig. 1. Autocorrelation receiver

Also define the autocorrelation function

$$R(\tau, k) = \int_{(k-1)T_{sam}}^{kT_{sam}} h(t)h(t-\tau)dt\,.$$

After integrate-and-dump, the received samples are

$$\begin{aligned} x_0[k] &= [R(0, k - \frac{D}{T_{sam}}) + R(2D, k)]s + \\ &\quad +[R(D, k) + R(D, k - \frac{D}{T_{sam}})] \end{aligned} \tag{3}$$

In equation (3), the dominant term in $x_0[k]$, i.e. after integrate-and-dump, is the energy of the channel segments $R(0, k)$, while the autocorrelation terms $R(\tau, k)$ with $\tau \in \{D, 2D\}$ can be assumed zero (for a totally uncorrelated channel) or very small (in case of a highly uncorrelated channel).

Note that in this process we obtain "channel segments" because the integrate-and-dump process virtually divides the spreading channel into P segments (or sub-channels) per frame. Each segment will have its own "channel energy" and "channel autocorrelation function".

B. Multiple frames–Simple data model

If we take into account multiple frames, there will be inter-frame interference (IFI) terms and the cross-correlation terms between different segments of the channel in different frames. However, since the correlation length is rather long (comparable to the frame period T_f and the channel length T_h), these cross-correlation terms can be ignored.

The most simple data model can be straightforwardly derived when we assume a totally uncorrelated channel, which will eliminate all the autocorrelation and cross-correlation terms. The dominant terms left in $x[k]$ are the energy of the channel segments and the overlapping parts (IFI).

Collect NP samples $x[k]$, $k = 0, \cdots, NP-1$ and stack them in a vector $\mathbf{x}$. The sample vector can then be expressed as

$$\mathbf{x} = \mathbf{H}\mathbf{s} + \text{noise} \tag{4}$$

where $\mathbf{s}$ is the unknown symbol vector $\mathbf{s} = [s_1, s_2, \cdots, s_{N_s}]^T$, and $\mathbf{H}$ is the unknown channel matrix containing shifted versions of the channel correlation vector $\mathbf{h} = [h_1, h_2, \cdots, h_L]^T$ where $L = \lceil T_h/T_{sam} \rceil$, and

$$h_i := \int_{(i-1)T_{sam}}^{iT_{sam}} h^2(t)dt\,.$$

This is illustrated in Fig. 2. We can interprete the h_i as samples of the power delay profile (PDP) of the channel, sampled at periods T_{sam}. The presence of IFI is seen from the fact that rows of $\mathbf{H}$ have more than one nonzero entry.

This is a very simple model, but it turns out, as shown later in simulations, that the corresponding receiver algorithm can work quite well especially in the line-of-sight (LOS) case.

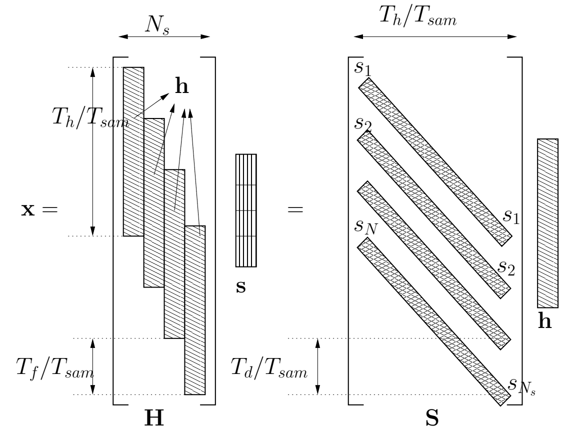

Fig. 2. Two forms of the data model

C. Multiple frames–Improved data model

As discussed earlier, the non-ideal antenna effect can introduce small correlation into the channel but this will go down quickly as the correlation length increases. Therefore, the autocorrelation terms (within one frame) are included into the model, while the cross-correlation terms (between frames) are still left out. This means we will keep the "offset" term in equation (3).

With this, the data model becomes

$$\mathbf{x} = \mathbf{H}\mathbf{s} + \mathbf{B}\mathbf{1} + \text{noise}\,, \tag{5}$$

where $\mathbf{B}$ has the same structure as $\mathbf{H}$, containing shifted versions of the "offset" vector $\mathbf{b} = [b_1, b_2, \cdots, b_{T_h/T_{sam}}]$, where

$$b_i := R(D, i) + R(D, i - \frac{D}{T_{sam}})\,.$$

III. Receiver algorithms

Given $\mathbf{x}$, our aim is to blindly estimate $\mathbf{H}$ and $\mathbf{s}$. The Sylvester structure of $\mathbf{H}$ implies that a range of blind equalization algorithms are applicable, e.g., the subspace methods of Moulines e.a. [9]. These would then have to be extended to apply to the more accurate model (5). Since such techniques may be complex and unreliable if the channel length is not well defined, we will introduce simpler yet powerful algorithms based on the more specific characteristics in this TR-UWB context.

A. Simple blind receiver algorithm

The most simple algorithm follows by ignoring all IFI in our data model, i.e., we approximate $\mathbf{H}$ by a block-diagonal matrix. Dividing the sample vector into segments $\mathbf{x}_k$ of length $P = T_f/T_{sam}$, we obtain from the simple data model (4)

$$\mathbf{x}_k \approx \tilde{\mathbf{h}} s_k$$

where $\tilde{\mathbf{h}} = [h_1, h_2, \cdots, h_P]^T$ is a sub-vector of $\mathbf{h}$. Collecting all vectors $\mathbf{x}_k$ into a matrix $\mathbf{X} = [\mathbf{x}_1, \mathbf{x}_2, \cdots, \mathbf{x}_{N_s}]$ gives

$$\mathbf{X} = \tilde{\mathbf{h}}\mathbf{s}^T$$

¿From this equation, the data symbol vector $\mathbf{s}$ can be easily estimated by taking a rank-1 approximation (SVD) of the sample matrix $\mathbf{X}$.

Perhaps surprisingly, this simple algorithm is seen to work well for most of our measured LOS UWB channels because for LOS scenarios, most channel energy is concentrated in the first few paths.

Therefore, even if we ignore the overlapping parts, the receiver still gather most of the neccessary channel energy to successfully detect the data symbols.

B. Iterative receiver algorithm

For simplicity and clarity, we derive the following algorithm only for the simple data model in (4). The algorithm can be easily extended to the improved model (5).

First, condiser the simple data model that was already derived in the previous section

$$\mathbf{x} = \mathbf{H}\mathbf{s} \tag{6}$$

We can rewrite this equation as

$$\mathbf{x} = \mathbf{S}\mathbf{h} \tag{7}$$

where $\mathbf{S}$ is based on $\mathbf{s}$; the structure of $\mathbf{S}$ is illustrated in Fig. 2. This suggests that the unknown data $\mathbf{s}$ and unknown channel coefficients $\mathbf{h}$ can be estimated by a familiar iterative algorithm, as follows:

- *Inital estimate:* Set the channel vector to the initial estimate $\mathbf{h} = \mathbf{h}_0$.
- *Iteration:*
  (a) With known channel vector $\mathbf{h}$, estimate the symbol vector based on (6), as
  $$\mathbf{s} = \mathbf{H}^{\dagger}\mathbf{x}$$
  where $\dagger$ indicates the Moore-Penrose pseudo-inverse. (b) Make hard decisions on the source symbols $\mathbf{s}$, using their finite alphabet property.
  (c) With known data symbol $\mathbf{s}$, the channel vector can be re-estimated again based on (7), as
  $$\mathbf{h} = \mathbf{S}^{\dagger}\mathbf{x}$$

This algorithm is rather simple yet powerful because the matrices which are inverted ($\mathbf{H}$ and $\mathbf{S}$) are generally tall. In general, i.e., for arbitrary vectors $\mathbf{h}$, the algorithm can take many iterations to converge because of the random initial estimate $\mathbf{h}_0$. However, in our case, we can exploit two important facts about the channel vector $\mathbf{h}$. Firstly, since most of the channel energy is concentrated on the first few paths (for LOS cases), we can use the previously mentioned blind algorithm to obtain a relatively accurate initial estimate for $\mathbf{h}$. Secondly, the vector $\mathbf{h}$ is actually a sampled version of the channel power delay profile (PDP). Therefore, we can use known average PDPs as an initial channel estimate, which will significantly improve the convergence of the iterative algorithms. As a result, we only need a few (one to four) iterations for most cases.

One of the main concerns in every iterative algorithm is about its computational complexity. From Fig. 2, we can see that both $\mathbf{H}$ and $\mathbf{B}$ are very sparse, with known structure. As in [10], there are efficient techniques to invert such matrices with a very low computation complexity.

IV. Simulations

In this section, we simulate our scheme with Monte Carlo runs, each run $N_s = 40$ symbols (or frames) are transmitted at the frame period $T_f = 20$ ns. The UWB pulse is a Gaussian pulse, with width parameter 0.2 ns; the delay between two pulses in a doublet is $D = 0.5$ ns. We consider both measured channels and the IEEE standard channel models for both LOS and NLOS cases. The measured channels are taken in a 40-m wide and 15-m high industrial hall ("API") at TU Delft, and include the transmit and receive antenna response, and a bandpass filter (3.5–8.5 GHz). The theoretical IEEE channel models are CM-1 and CM-4 from IEEE 802.15.3a standard. To obtain the actual channel, we convolve these (twice) with the measured antenna response, see figure 3, which will distort the received pulse shape and give it a length of about 1.5 ns. For LOS cases, the channel length can be as long as 80 ns although most of its energy is concentrated in the first few paths. For NLOS cases, the channel length can be up to 300 ns (CM-4), with no dominant path. The oversampling rate is $P = 20$ samples per frame, which means that the sampling period is $T_{sam} = 1$ ns.

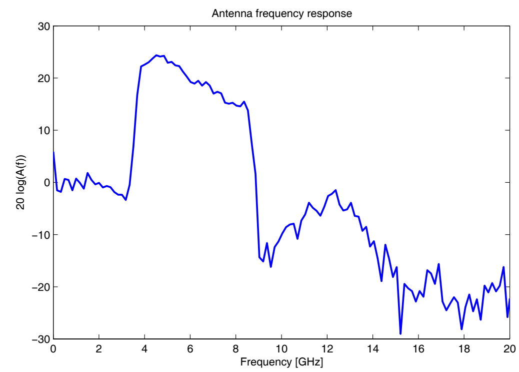

Fig. 3. Measured UWB antenna transfer function

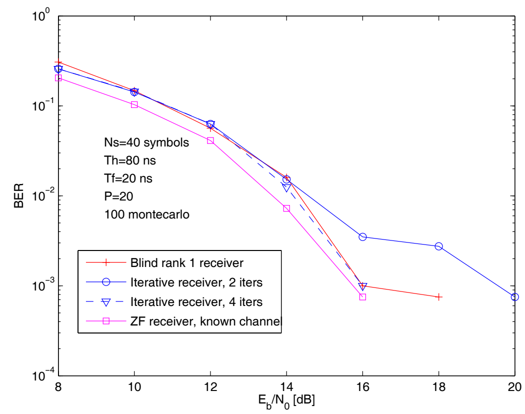

Fig. 4. BER vs. SNR plots for the CM-1 channel model (LOS)

For each channel, we will compare the BER vs. SNR plots of the various algorithms (blind algorithm, iterative algorithms with initial values are chosen as the rough average of the channel power delay profile). The reference curve is that of the zero-forcing (ZF) receiver when channel is known. SNR is defined as the pulse energy over the noise spectral density.

In Fig. 4, we can see that the simple blind receiver algorithm have a very good performance—only about 1 dB less than the "optimal" curve. This is because we use the CM-1 LOS channel model with short channel length, high concentration on the first few paths. However, this algorithm has the floor effect in high SNR region due to the model error (when ignoring the IFI). The iterative algorithm (two iterations) has a good performance but it also has floor effect at high SNR due to "imperfect" initialization. However, significant improvement can be achieved if we use the improved data model and increase the number of iterations, which completely eliminates the floor effect.

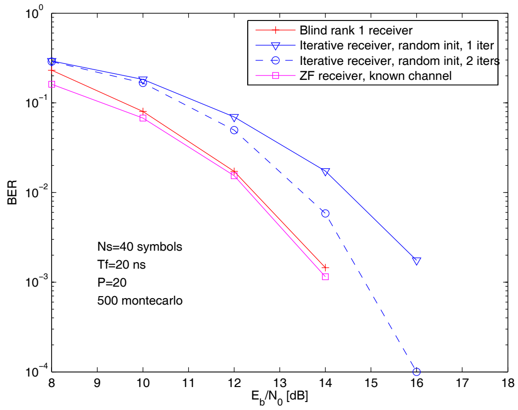

Fig. 5. BER vs. SNR plots for a measured "API" LOS channel

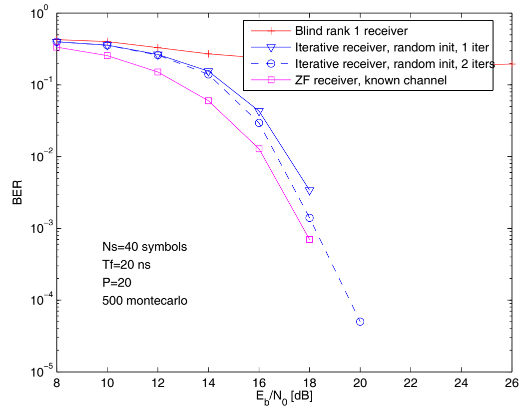

Fig. 7. BER vs. SNR plots for a measured "API" NLOS channel

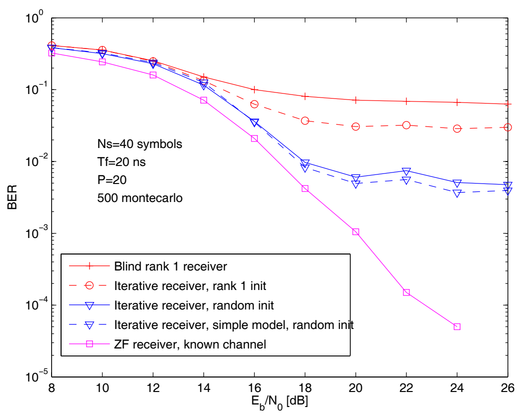

Fig. 6. BER vs. SNR plots for the CM-4 channel model (NLOS)

We obtain similar results for the "API" LOS measured channel in Fig. 5. The only difference is that the curve for the blind receiver is closer to the reference curve and the flooring effect is not as pronounced as before. This is because in our measured LOS channel, the channel energy seems to be more concentrated in the first paths than for the IEEE channel model CM-1.

For the NLOS cases, we can see in both Fig. 6 and 7 that the simple blind receiver does not work anymore due to missing significant channel energy in the overlapping parts. The iterative receiver initialized by random channel estimate (2 iterations) can still work quite well. If we increase the number of iterations and use the improved data model, it can help reduce much of the floor effect. at high SNRs.

V. Conclusions

In this paper, we have introduced a TR-UWB scheme which is more suitable for high data rate applications. Several simple receiver algorithms have also been derived. It is shown, in simulations, that these receivers are robust (even with NLOS channels) and flexible (e.g., the sampling rate can be varied). The use of oversampling also implies a more simple synchronization scheme, which will be considered in future publications. Furthermore, our aim in future extensions of this paper is to find a more complete solution for the initial channel estimate.

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