Narrowband interference mitigation for a transmitted reference ultra-wideband receiver

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***Abstract*— Narrowband inteference (NBI) is of specific concern in transmitted reference ultrawide band (TR-UWB) communication systems. We consider NBI in high data rate applications where significant interframe interference is present due to a very short frame rate. Oversampling of the correlator output with respect to the frame rate is used to gather more information for the receiver. We formulate an approximate data model that includes the NBI terms, subsequently a receiver algorithm is derived.**

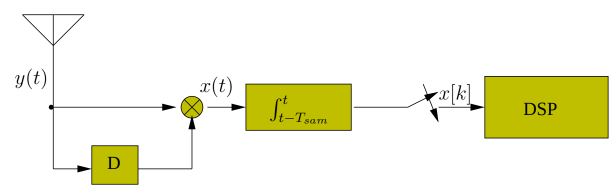

Fig. 1. Autocorrelation receiver

I. Introduction

The transmit-reference (TR) scheme, first introduced in [1], has become a realistic candidate for impulse-based ultra-wideband (UWB) communication systems. In a TR system, without having to estimate all the individual channel taps, the transmitted symbols can be detected by collecting the energy of all the multipath components i.e. the total channel energy. This will greatly reduce the receiver complexity. However, the fact that most papers on TR scheme do not allow interframe interference implicitly limits the data rate of these systems.

In [2], we proposed an alternative TR approach that allows for interframe interference (thus enabling higher data rates). By oversampling (more than one sample per frame), we can collect the energies of the channel segments instead of the total channel energy. A signal processing model and receiver algorithms were derived to detect the transmitted symbols. This paper extends the work by considering how the narrowband interference affects the previously established data model and receiver algorithm.

Although several research papers on TR-UWB have appeared, not many consider the presence of narrow band interference (NBI). The correlation operation in TR-UWB receivers makes it difficult to investigate and thus eliminate the NBI effect. In [3], statistics of the cross terms (due to the correlation operation) "NBI by NBI" and "NBI by data" were studied, where a "code" is used to mitigate the NBI when its frequency is known. In [4], a data model and some receiver algorithms were derived to deal with NBI in low data rate applications with no inter-frame interference. Both mentioned papers make use of a long integration time to average out some of the NBI effects. In this paper, we will analyze the effect of NBI in a high data rate application context, where the integration is much shorter, i.e. with several samples per frame. An approximate signal processing data model, which exploits the high data rate and narrowband nature, is proposed. Subsequently, the performance improvement of the receiver algorithm based on this model is shown.

II. Data model

As introduced in [2], we consider a high rate single user TR-UWB system, with frame rate T_f much less than the channel length T_h. Each frame contains a doublet: two subsequent pulses spaced by D. Each doublet is associated with a symbol value s_i. The assumed channel is specified as uncorrelated dense multipath in a typical UWB indoor environment.

The receiver structure is shown in Fig. 1. The received signal at the antenna output is

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

where $h(t) = h_p(t) * g(t) * a(t)$ is the convolutional product of the physical channel $h_p(t)$, the UWB pulse shape $g(t)$ and the antenna template $a(t)$, and $\gamma(t)$ is the narrowband interference (NBI)

$$\gamma(t) = \sqrt{2}v(t)\cos(2\pi f_I t + \theta)$$

where $v(t)$, f_I and θ are respectively the baseband signal, carrier frequency and random (uniformly distributed) phase of the NBI.

At the multiplier output, the signal $x(t) = y(t)y(t-D)$ will be integrated and dumped at the oversampling rate $P = T_f/T_{sam}$. The resulting discrete signal $x[k]$ will include three cross-terms: the "data by data" term $x^{(1)}[k]$, the "data by NBI" term $x^{(2)}[k]$ and the "NBI by NBI" term $x^{(3)}[k]$.

The first term, "data by data", was considered in our previous research [2]. Putting all samples $x^{(1)}[k]$ into a vector, we have

$$\mathbf{x}^{(1)} = \mathbf{H}\mathbf{s}$$

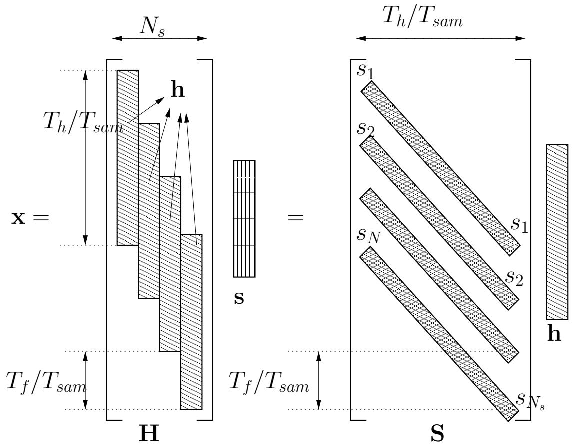

Fig. 2. Two forms of the data model

where $\mathbf{H}$ contains the shifted versions of vector $\mathbf{h}$ with entries $h[k]$, $k = 1, \cdots, (T_h/T_{sam})$, defined (with some abuse of notation) as

$$h[k] = \int_{(k-1)T_{sam}}^{kT_{sam}} h^2(t)dt$$

The structure of the "channel" matrix $\mathbf{H}$ is illustrated in Fig. 2. We thus have replaced the "continuous" channel taps $h(t)$ with the discrete channel parameters $h[k]$, which are the energies of the corresponding channel segments in one T_{sam} interval. The new channel vector $\mathbf{h}$ now has a much smaller number of unknowns, which will benefit the receiver's complexity and performance.

Now we will look at the second and the third term in $x[k]$ that deal with the NBI signal. First, since $v(t)$ is narrowband ($B \ll \frac{1}{T_{sam}}$), we can assume that it is constant during one integration period T_{sam}. Therefore, the "NBI by NBI" term can be expressed as

$$\begin{aligned} x^{(3)}[k] &:= \\ & 2\int_{(k-1)T_{sam}}^{kT_{sam}} v(t)\cos(2\pi f_I t + \theta)v(t-D)\cos(2\pi f_I(t-D)+\theta)dt \\ &= v_k^2 \int_{(k-1)T_{sam}}^{kT_{sam}} [\cos(2\pi f_I(2t-D)+2\theta) + \cos(2\pi f_I D)]dt \\ &= v_k^2 T_{sam}\cos(2\pi f_I D) \\ &\quad + v_k^2 \int_{(k-1)T_{sam}}^{kT_{sam}} \cos(2\pi f_I(2t-D)+2\theta)dt \end{aligned}$$

The second term in the equation above is always less than $v_k^2 \cdot \frac{1}{\pi(2f_I)}$, where $\frac{1}{\pi(2f_I)}$ is the maximum value of the integration of a zero-mean cosine wave of frequency $(2f_I)$. When T_{sam} is in the order of a nanosecond while the NBI carrier f_I is in the GHz range ($T_{sam} \gg 1/(2f_I)$), this can help increase the dominance of the first term. Unfortunately, since the value of $\cos(2\pi f_I D)$ can be arbitrary small, the condition on T_{sam} and f_I is not enough to make any conclusion about the relative magnitudes of the two terms. In the worst case, when $T_{sam}\cos(2\pi f_I D) \gg \frac{1}{\pi(2f_I)}$, the "NBI by NBI" term can be approximated as a constant with a small fluctuation ϵ_k

$$x^{(3)}[k] \approx v_k^2 T_{sam}\cos(2\pi f_I D) + \epsilon_k$$

The "data by NBI" term for one frame can be expressed as

$$\begin{aligned} & x^{(2)}[k] := \textstyle\int_{(k-1)T_{sam}}^{kT_{sam}} [h'(t)\gamma(t-D) + h'(t-D)\gamma(t)]dt \\ & = \sqrt{2}v_k \textstyle\int_{(k-1)T_{sam}}^{kT_{sam}} [h'(t)\cos(2\pi f_I(t-D)+\theta) \\ & \quad + h'(t-D)\cos(2\pi f_I t + \theta)]dt \end{aligned}$$

where $h'(t) = h(t) + s_i h(t-D)$. Note that although we have cross-terms from other frames, they can be ignored due to the highly uncorrelated channel. The question is whether this term is relatively small compared to the "NBI by NBI" term, and how it relates to the signal to interference ratio (SIR).

First, to simplify the expression, we compare two vectors $\mathbf{h}$ and $\mathbf{h}'$, with entries for $k = 1, \cdots, (T_h/T_{sam})$

$$\begin{aligned} h'[k] &= \sqrt{2}v_k \int_{(k-1)T_{sam}}^{kT_{sam}} h(t)\cos(2\pi f_I t + \theta)dt \\ h[k] &= \int_{(k-1)T_{sam}}^{kT_{sam}} h^2(t)dt \end{aligned}$$

Let us remind that, for easier derivation, we use s with unit energy, which implies that the signal energy is embedded in the energy of the composite channel $h(t) = g(t) * h_p(t)$, where $h_p(t)$ is the physical channel. Therefore, the ratio between the norm of $\mathbf{h}$ and $\mathbf{h}'$ is directly related to the SIR. Without loss of generality, we assume $h(t)$ of the user of interest to have normalized energy. We define the SIR at the antenna as the average received signal energy of one UWB pulse spread by a normalized channel over the interference energy within the same time interval.

The ratio between the norms of two vector $\mathbf{h}$ and $\mathbf{h}'$ is

$$\frac{\|\mathbf{h}\|}{\|\mathbf{h}'\|} = \frac{SIR}{2}\Gamma$$

In Fig. 3 and 4, we compare these two vectors for the channel models CM1 and CM3 (the results are averaged over 100 realizations) that include the UWB pulse shape and antenna effect for different sampling intervals. For $SIR = 0$ dB, the factor Γ is

$$\Gamma \approx \begin{cases} 11.7 & \text{for CM1, } T_{sam} = 1\text{ns} \\ 13.4 & \text{for CM1, } T_{sam} = 2\text{ns} \\ 19.2 & \text{for CM1, } T_{sam} = 4\text{ns} \\ 9.2 & \text{for CM3, } T_{sam} = 1\text{ns} \\ 11.1 & \text{for CM3, } T_{sam} = 2\text{ns} \\ 17.0 & \text{for CM3, } T_{sam} = 4\text{ns} \end{cases}$$

It can be concluded that, as we increase the integration interval T_{sam}, the effect of the "NBI by data" term will reduce. Meanwhile, the "NBI by NBI" will become more dominant. In a certain SIR range, this "NBI by data" term can be ignored.

Since our objective is to reduce the frame period $T_f = P * T_{sam}$ for a higher data rate system, we need to lower the sampling rate per frame, which will reduce the system performance slightly. However, the important assumption that

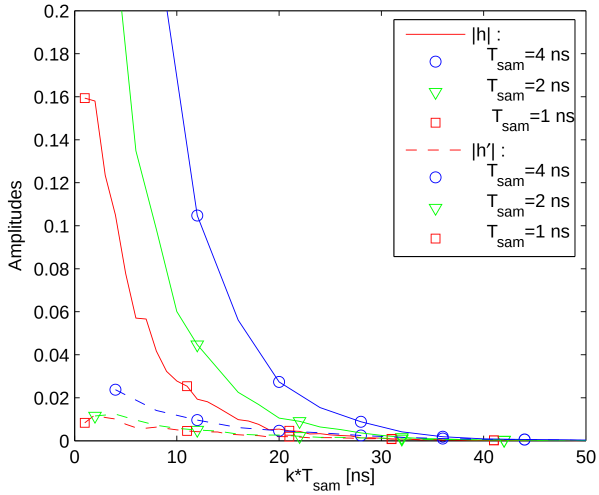

Fig. 3. Comparison between $\mathbf{h}$ and $\mathbf{h}'$ (CM1, SIR=0dB)

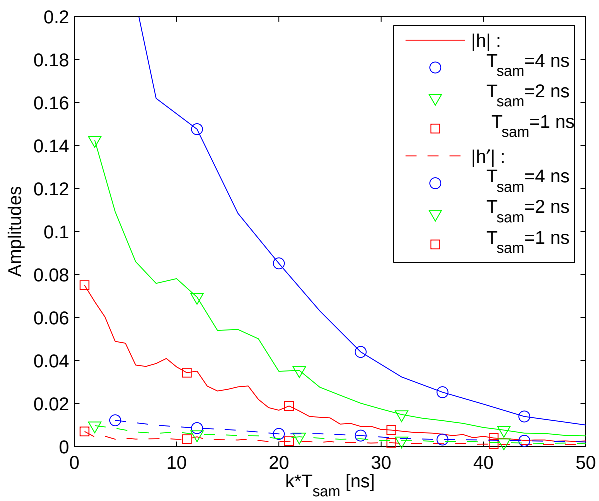

Fig. 4. Comparison between $\mathbf{h}$ and $\mathbf{h}'$ (CM3, SIR=0dB)

the NBI envelope $v(t)$ is constant over the integration period requires T_{sam} to be small enough.

If the dominant term is the "NBI by NBI" term, we can write the data model as

$$\mathbf{x} = \mathbf{H}\mathbf{s} + \mathbf{v} \qquad (2)$$

where $\mathbf{v} = T_{sam}\cos(2\pi f_I D)[v_1^2, v_2^2, \cdots]^T$.

III. Receiver algorithms

In our data model (2), all the parameters on the right hand side are unknowns (more unknowns than received samples), which makes it hard, if not impossible, to find a good estimation solution. Luckily, due to high data rate and the fact that the interference is narrowband, the frame period T_f is much smaller than the reciprocal of the bandwidth of the NBI baseband signal. Therefore, it is valid to assume that all entries of vector $\mathbf{v}$ are approximately constant over one frame period. This assumption greatly reduces the number of unknowns in $\mathbf{v}$ by a factor of P (the number of samples we take per frame $T_f = PT_{sam}$), which makes it possible to solve the problem iteratively.

With this assumption, the NBI vector can be expressed as

$$\begin{aligned} \mathbf{v} &= \mathbf{v}' \otimes \mathbf{1}_P & (3)\\ &= \mathbf{J}\mathbf{v}' & (4) \end{aligned}$$

where $\mathbf{v}' = [v_1', v_2', \cdots]^T$, v_i' is the value of the NBI parameter in the i-th frame, and $\mathbf{J} = \mathbf{I} \otimes \mathbf{1}_P$.

Similar to the algorithms proposed in [2], we can rewrite (2) in two forms, as illustrated in Fig. 2, namely

$$\begin{aligned} \mathbf{x} = \mathbf{H}\mathbf{s} + \mathbf{J}\mathbf{v}' &= [\mathbf{H} \quad \mathbf{J}][\mathbf{s}^T \quad \mathbf{v}'^T]^T & (5)\\ &= [\mathbf{S} \quad \mathbf{J}][\mathbf{h}^T \quad \mathbf{v}'^T]^T. & (6) \end{aligned}$$

An iterative estimation algorithm for $\mathbf{h}$, $\mathbf{s}$ and $\mathbf{v}$ can be straightforwardly implemented. The intial channel channel estimate can be obtained by using a training sequence (assume that the first few data symbols are known).

Note that the data model and the receiver algorithm were derived for the "noise-free" case, which means we ignore 5 cross terms that includes noise among the total number of 9 terms. In the simulation section discussed next, we will consider over which range of parameters this approximation is valid.

IV. Simulation

We simulate the transmission of a TR-UWB scheme at high symbol rate. The frame period is $T_f = 20$ ns. Since there is only one frame per symbol, the symbol rate is 50 Mbps. UWB Gaussian monocycles of width 0.2 ns are transmitted through different IEEE channel models that take into account the nonideal antenna response. The delay between two pulses in one doublet is $D = 0.5$ ns. The channel length can be 100 ns (for CM1) up to 300 ns (for CM4).

We use 1000 Monte Carlo runs to compare the BER vs SIR (signal to interference ratio) plots of receiver algorithms when we take the NBI effect into account and when we ignore it, i.e., use the same algorithm and model but set the NBI term to zero. To emphasize the difference, we implement two receiver algorithms when we have perfect channel estimation, i.e. the channel vector $\mathbf{h}$ is known. Note that we have much less unknown channel parameters now (compared to number of all of the "real" channel taps), which can be easily estimated by some training symbols. Here, the BER and SIR are measured at the output of the antenna (before the correlator), the noise power is taken for the whole pulse sequence.

In Fig. 5 and Fig. 6, we can see that in the low SIR region i.e. strong NBI signal, with our data model, we can significantly improve the BER performance, which can be as much as 5 dB. However, as the NBI signal strength decreases, the improvement also reduces, until a certain threshold SIR $\approx 0 \div 5$dB (depending on the channel model and noise power) we obtain no gain anymore. After that, the old receiver algorithm, which ignores the NBI effect, outperforms the new one. This is foreseeable because when the NBI is small enough, it can be neglected or regarded as part of noise, therefore the old algorithm prevails as the number of parameters it needs to estimate is only half of that of the new algorithm (more

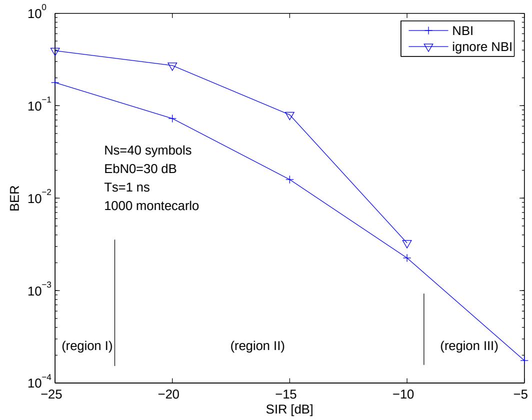

Fig. 5. BER vs. SIR plots for IEEE channel model CM1, SNR=30dB

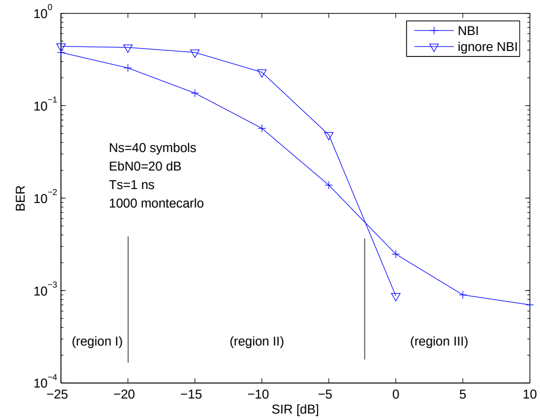

Fig. 7. BER vs. SIR plots for IEEE channel model CM1, SNR=20dB

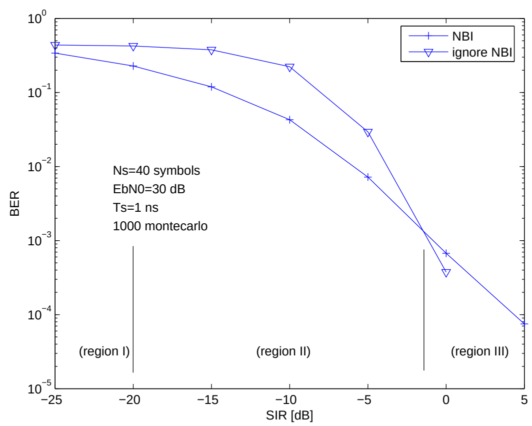

Fig. 6. BER vs. SIR plots for IEEE channel CM2, SNR=30dB

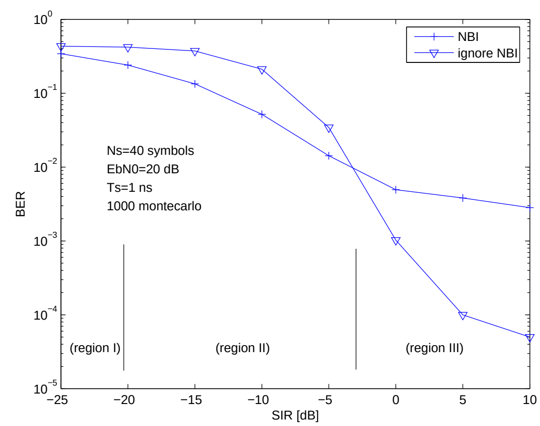

Fig. 8. BER vs. SIR plots for IEEE channel CM2, SNR=20dB

specifically, the matrix it needs to invert is two times more tall than the other).

For a better understanding, we will look more detail into the BER vs. SIR performance with different values of the noise power. Fig. 7 and Fig. 8 show the plots when SNR=20dB. We can see that as the noise increases, its effect will be more visible than the NBI. The floor effect in high SIR region is the performance limit under the current noise power.

We can roughly divide the SIR range into three regions as follows. The high SIR region (III) is when the NBI signal is weak enough to be regarded as noise, which supports the receiver algorithm that ignores the NBI effect. The low SIR region (I) is when the NBI signal is so strong that the cross term "NBI by data" becomes significant, which will destroy the data model in (2). This explains why the performances of the algorithms are limited in this region. The "middle" SIR region (II) satisfies the data model, which gives expected superior performance of the algorithm that deals with the NBI signal.

V. Conclusions

An approximate data model has been derived to deal with NBI problem in a TR-UWB communication system. Simulation results show that at a certain range of signal-to-interference ratio (SIR) we can mitigate the NBI effect. However, the model will not be valid anymore when the NBI signal is too strong, which results in significant increase in the cross-terms: NBI by data and NBI by noise. In this case, we may have to filter the NBI signal out before it enters the autocorrelation receiver.

References

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