A FREQUENCY DOMAIN APPROACH TO CHANNEL ESTIMATION, DETECTION, AND INTERFERENCE CANCELLATION FOR IMPULSE RADIO SYSTEMS

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ABSTRACT

In this paper we propose and investigate a frequency domain approach to channel estimation, and detection for ultra wide band (UWB) direct sequence code division multiple access (DS-CDMA) systems. The channel estimation and detection approach is single user based, and operates in the frequency domain. In the presence of multiple access interference (MAI) the algorithm is appropriately modified to include the capability of canceling the interference through the exploitation of its frequency domain correlation.

1. INTRODUCTION

In this paper we consider the synchronization, channel estimation, and detection problem in impulse radio systems. The attractive feature in impulse radio systems is the carrier-less baseband implementation that involves transmission of short duration pulses. This technology is commonly referred to as ultra wide band (UWB) because the pulses can occupy a very large bandwidth. Our system model assumes bi-phase pulse modulation (BPAM) in conjunction with direct sequence code division multiplexing of users (DS-CDMA) [2], [3]. Binary codewords are assigned to users, and modulate short duration pulses (monocycles). A user's codeword spans a transmission frame. Frames are separated by a guard time to cope with the channel time dispersion. When the guard time is longer than the channel time dispersion, and only a single user accesses the medium, the optimal receiver comprises a matched filter followed by a symbol by symbol threshold detector. The receiver filter has to be matched to the equivalent impulse response that comprises the user's waveform, and the channel impulse response. Since UWB signals can occupy a large bandwidth, the channel is highly frequency selective, and the received signal exhibits a large number of multipath components. Potentially, high frequency diversity gains can be achieved. However, the optimal matched filter receiver has to accurately estimate the channel, and such an estimation can be particularly complex if performed in the time domain. It has been shown in [6] that channel estimation can be partitioned into a two step process if we model it as a tapped delay line. That is, we first determine the channel ray delays, and then we obtain an estimate of the ray amplitudes. Unfortunately, the ray search has a complexity that grows exponentially with their number. Further, false ray detection may occur in the absence of a priori knowledge about the true number of rays. The search can be partially simplified under the assumption of the channel to be separable [4], [6]. However, this assumption translates into deep performance losses in the non-rare event of clusters of non-separable rays.

When the common media is shared by multiple users, multiple access interference (MAI) may arise at the receiver side. In a DS-CDMA system, this is due to the deployment of non orthogonal codes, or to users that are time asynchronous, or to the presence of channel time dispersion. Assuming a single user detection approach the MAI translates into performance losses, such that some form of multiuser detection is advisable [3].

Motivated by the above considerations, we consider a frequency domain approach to channel estimation, and detection [1]. The approach is single user based. However, it can include the capability of rejecting the MAI interference. The approach comprises the following stages. First we acquire frame synchronization with the desired user. Second, we run a discrete Fourier transform (DFT) on the received frames. Third, we perform frequency domain channel estimation for the desired user via a recursive least squares (RLS) algorithm. Finally, detection is accomplished in the frequency domain using the estimated channel frequency response. In the presence of multiple access interference the algorithm is appropriately modified to include the capability of canceling the interference. Interference rejection is accomplished by observing that the MAI manifests itself with a frequency domain correlation that can be estimated and exploited by the detector.

2. SYSTEM MODEL

In our system model (Fig. 1) we assume bi-phase pulse amplitude (BPAM) modulation such that the signal transmitted by user u can be written as

$$s^u(t) = \sum_k b_k^u g^u(t - kT_f) \qquad (1)$$

where $b_k^u = \pm 1$ denotes the information bit transmitted in the k-th frame, $g^u(t)$ is the waveform used to convey information for

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user u, and T_f is the bit period (frame duration). We further deploy direct sequence spreading to accommodate for multiplexing of users [2]. The user's waveform (signature code) comprises the weighted repetition of $L \geq 1$ narrow pulses (monocycles), i.e.,

$$g^u(t) = \sum_{m=0}^{L-1} c_m^u g_M(t - mT) \quad (2)$$

where $c_m^u = \pm 1$ are the codeword elements (chips) of user u, and T is the chip period. We can choose the codewords to be either orthogonal or random (pseudo-noise). We incorporate the differential effect of the transmit-receive antennas into $g_M(t)$, and we assume it to be the second derivative of the Gaussian pulse, $g_M(t) \sim \exp(-\pi/2((t-D/2)/T_0)^2)$. In typical system design we can choose $T \geq D$ where $D \approx 5T_0$ is the monocycle pulse duration. We further insert a guard time T_g between frames to cope with the channel time dispersion, and eliminate the inter-symbol interference (ISI). The frame duration fulfils the relation $T_f > LT + T_{ch}$ with T_{ch} being the channel time dispersion.

As shown in Fig. 1, at the receiver side we first deploy a band-pass front-end filter with impulse response $g_{FE}(t)$ to suppress out of band noise, and interference. Then, the received signal in the presence of N_I other users (interferers), can be written as

$$y(t) = \sum_k b_k g_{EQ}(t - kT_f) + \sum_{u=1}^{N_I} \sum_k b_k^u g_{EQ}^u(t - kT_f - \tau_u) + \eta(t) \quad (3)$$

where u-th user's equivalent impulse response is denoted as $g_{EQ}^u(t) = g^u * h^u * g_{FE}(t)$, while τ_u denotes the time delay of user u with respect to the desired user's frame timing. For easy of notation we drop the index $u = 0$ for the desired user. The additive noise $\eta(t)$ is assumed to be a stationary white Gaussian process within the signal bandwidth, with zero mean, and double sided power spectral density $N_0/2$.

The equivalent impulse response comprises the convolution of the u-th user's transmission waveform (signature code) with its channel impulse response, and the front-end filter. Distinct users experience independent channels that we assume to introduce identical maximum time dispersion. The channel impulse response is assumed to be time-invariant over several transmitted frames. Then, it can change in a random fashion. With the popular discrete multi-path model [4], [6], the channel impulse response of user u can be written as

$$h^u(t) = \sum_{p=1}^{N_P} \alpha_p^u \delta(t - \tau_p^u). \quad (4)$$

As an example, in the numerical results that follow, we assume the tap delays to be independent, and uniformly distributed in an interval smaller than T_g, while the tap gains are assumed to be real, independent, and equal to $\alpha_p^u = \chi_p^u \beta_p^u$ with β_p^u Rayleigh distributed, while χ_p^u takes on the values ± 1 with equal probability. The power delay profile is assumed to be exponential. With this model the rays can appear in clusters of duration less than D, i.e., they are not necessarily resolvable.

3. FREQUENCY DOMAIN PROCESSING

The conventional correlation receiver (matched filter receiver) operates in a symbol by symbol fashion by computing the correlation between the received signal frame $y_k(t) = y(t + kT_f)$, $0 \leq t < T_f$, and the real equivalent impulse response $g_{EQ}(t)$ to obtain $z(kT_f) = \int_0^{T_f} y_k(t) g_{EQ}(t) dt$. Then, a threshold decision is made to detect the k-th transmitted bit, i.e., $\hat{b}_k = sign\{z(kT_f)\}$.

To implement the correlation receiver in the time-domain we need to estimate the channel impulse response. Time-domain channel estimation is complicated by the high number of multipath components exhibited by UWB channels, and by the presence of non resolvable channel rays, i.e., rays with relative time delay smaller than the monocycle duration D. Thus, $g_{EQ}(t)$ can be an involved function of the channel, and the transmitted waveform.

In this paper we propose to perform channel estimation, and detection in the frequency domain. We assume discrete-time processing, such that the received signal is sampled at the output of the front-end analog filter at sufficiently high rate. We acquire frame synchronization with the desired user. Then, we run an M-point discrete Fourier transform (DFT) over the M samples of the k-th frame $y_k(nT_c)$, $T_c = T_f / M$, to obtain

$$Y_k(f_n) = b_k G_{EQ}(f_n) + I_k(f_n) + N_k(f_n) \qquad n = 0,...,M-1 \quad (5)$$

where $Y_k(f_n)$, $G_{EQ}(f_n)$, $I_k(f_n)$, and $N_k(f_n)$ for $f_n = n/MT_c$, are respectively the DFT outputs of : the received frame samples, the desired user equivalent impulse response, the interference, and the noise samples. No ISI is present for the desired user assuming perfect frame timing, and a sufficiently long guard time. The MAI additive term in the presence of asynchronous users, or synchronous users, is a function of the users' time delay, transmitted waveform, and channel. Note that in the asynchronous case two information bits per user may cause interference, while in the synchronous case only one bit generates interference.

Herein, we proceed by modeling $Z_k(f_n) = I_k(f_n) + N_k(f_n)$, $n = 0,...,M-1$, as a multivariate discrete-time Gaussian process. Assuming the transmitted bits to be i.i.d. and equally likely, the process has zero mean, and *time-frequency* correlation matrix equal to

$$\mathbf{R}(k,m) = E[\mathbf{Z}_k \mathbf{Z}_m^\dagger] \quad (6)$$

where the elements of $Z_k(f_n)$, for $n = 0,...,M-1$, have been collected in the vector $\mathbf{Z}_k = [Z_k(f_0),...,Z_k(f_{M-1})]^T$. In the asynchronous case $\mathbf{R}(k,m) = 0$ for $|m-k| > 1$, while in the synchronous case $\mathbf{R}(k,m) = 0$ for $|m-k| > 0$. Now, let us collect the elements of $Y_k(f_n)$, in the vector $\mathbf{Y}_k$ while the elements of $G_{EQ}(f_n)$ in the vector $\mathbf{G}_{EQ}$. It can be shown [5] that the maximum-likelihood detector in the frequency domain searches for the sequence of transmitted bits $\{\hat{b}_k\}$, $k = -\infty,...,+\infty$ (belonging to the desired user) that maximizes the following log-likelihood function

$$\Lambda(\{\hat{b}_k\}) = -\sum_{k=-\infty}^{\infty} \sum_{m=-\infty}^{\infty} [\mathbf{Y}_k - \hat{b}_k \mathbf{G}_{EQ}]^\dagger \mathbf{R}^{-1}(k,m) [\mathbf{Y}_m - \hat{b}_m \mathbf{G}_{EQ}]. \quad (7)$$

In order to simplify the algorithm complexity we neglect the MAI temporal correlation. Indeed, the MAI temporal correlation is zero only for the synchronous case. Then, by dropping the terms that do not depend on the information bit of the desired user, the log-likelihood function simplifies to

$$\Lambda(\hat{b}_k) \sim \hat{b}_k \operatorname{Re}\left\{\mathbf{G}_{EQ}^{\dagger}\mathbf{R}^{-1}(k,k)\mathbf{Y}_k\right\}. \quad (8)$$

Therefore, according to (8) the frequency domain receiver operates on a frame by frame basis, and it exploits the frequency correlation of the MAI. The computation in (8) can be interpreted as the result of matching the frequency response of the *k-th* frame with $\mathbf{G}_{EQ}^{\dagger}\mathbf{R}^{-1}(k,k)$ to obtain $z_{IC}(kT_f) = \mathbf{G}_{EQ}^{\dagger}\mathbf{R}^{-1}(k,k)\mathbf{Y}_k$. Then, we make a decision on the transmitted bit looking at the sign of $z_{IC}(kT_f)$.

In the absence of MAI, and with white noise, the correlation matrix is diagonal with diagonal elements equal to the noise variance. We assume the correlation matrix to be full rank, otherwise pseudo-inverse techniques can be used. The main idea behind the algorithm above is to perform interference cancellation in the frequency domain via decorrelation of the MAI.

To obtain (8) we need to estimate $G_{EQ}(f_n)$. The attractive feature with this approach is that the matched filter frequency response at a given frequency depends only on the channel response at that frequency. This greatly simplifies the channel estimation task. By exploiting the Hermitian symmetry of $G_{EQ}(f_n)$, the estimation can be carried out only over *M*/2 frequency bins. A further simplification is obtained by observing that the desired user's waveform can be written as $G(f_n) = G_M(f_n)\sum_{m=0}^{L-1} c_m e^{-j2\pi f_n mT}$. If we deploy a monocycle that has a frequency concentrated response, as the Gaussian pulse, we can assume that $G_M(f_n) \approx 0$ for, say, $f_n > 2/D$. Therefore, relevant signal energy is present only in a small number of frequency bins, and consequently channel estimation can be performed only over this fraction of bins. If $D=KT_c$, an estimate of the number of such sub-channels is $2M/K$. Another interesting characteristic of the frequency domain channel estimation approach is that no restrictive assumption about the channel impulse response has been made.

3.1. Frequency Domain Parameter Estimation

To estimate the frequency response of the desired user channel, and the interference correlation matrix we assume the deployment of a training sequence of *N* known bits. To keep it simple, we run estimation in a two steps procedure. First, we estimate the desired user's channel. Then, we estimate the interference correlation matrix. We implicitly assume the channel, and the MAI to be stationary over the transmission of several frames, i.e., $\mathbf{R} = \mathbf{R}(k,k)$. In particular, the *M*-bins channel frequency response can be obtained via a recursive least squares (RLS) algorithm that operates independently over the sub-channels [5]. Once we have computed the desired user's frequency domain channel estimate $\hat{\mathbf{G}}_{EQ}$, we compute an estimate of the interference correlation matrix $\hat{\mathbf{R}}$. Let us define the error vector in correspondence with the *i-th* frame as $\mathbf{e}_i = b_i\mathbf{Y}_i - \hat{\mathbf{G}}_{EQ}$ where $\{b_i\}$, $i = 0,\dots,N-1$, is the sequence of known training bits of the desired user. Then, we estimate the correlation matrix as $\hat{\mathbf{R}} = 1/N\sum_{i=0}^{N-1}\mathbf{e}_i\mathbf{e}_i^{\dagger}$. Further, to introduce a tradeoff between the effects of noise, and the effects of the MAI we add diagonal loading as follows: $\hat{\mathbf{R}} = (1-\rho)\hat{\mathbf{R}} + \rho\sigma_N^2\mathbf{I}$, with $\rho \le 1$ and $\mathbf{I}$ being the identity matrix. For practical purposes the noise variance can be set to an appropriate value according to the range of operating signal-to-noise ratios.

3.2 Frame Synchronization

Frame synchronization is acquired in the time domain and uses the training bit sequence. The method is divided in two steps.

First Step - Coarse Timing. Coarse timing is obtained by locking on the time instant where the channel exhibits the highest energy. We assume sampling resolution equal to T_c. Then, the training sequence coarse starting epoch $t_1 = \hat{p}_1 T_c$ is determined as follows

$$\hat{p}_1 = \arg\max_{p\in\mathbb{Z}}\left\{\left|S_1(p)\right|^2\right\} \qquad S_1(p) = \frac{1}{N}\sum_{i=0}^{N-1} b_i\, y(pT_c + iMT_c). \quad (9)$$

The metric derives from the observation that in correspondence with the known training sequence, the frame signals are identical besides the sign flip imposed by the training sequence.

Second Step - Fine Timing*.* Once we have locked into the highest energy channel tap we need to refine the synchronization by establishing where the frame is located around the highest energy channel tap. The fine synchronization strategy that we propose is based on the idea of looking at the received energy content of windows of duration MT_c. The starting epoch of a given window falls in the interval $[(-M+\hat{p}_1)T_c, (M+\hat{p}_1)T_c]$. To keep the complexity at moderate levels, we down-sample that interval by a factor M_w, so that the frame starting epoch is taken to be $t_2 = (\hat{p}_1 + \hat{p}_2 M_w)T_c$ for a given $\hat{p}_2 \in \{-M/M_w,\dots,M/M_w\}$. The integer $\hat{p}_2$ is determined via the following maximization

$$\hat{p}_2 = \arg\max_{p\in\{-M/M_w,\dots,M/M_w\}} \sum_{i=0}^{M/M_w-2} S_2(pM_w + iM_w) \quad (10)$$

$$S_2(p) = \frac{1}{2M_w}\sum_{k=-M_w}^{M_w-1} |S_1(p+\hat{p}_1+k)|^2. \quad (11)$$

Note that (11) yields an estimation of the received energy in a window of duration $2M_w$ that is centred at time instant $pT_c + \hat{p}_1 T_c$. Overall, (10) corresponds to compute the received energy in a frame of duration MT_c, and to smooth by one half the energy content of the two windows of M_w samples at the beginning and the end of the frame itself.

4. PERFORMANCE RESULTS

In Fig. 2, the performance of the overall algorithm that combines frame synchronization, and channel estimation is shown for a single user case. We deploy 100 training bits. The channel model in (4) has $N_P = 10$ paths that have uniformly distributed delays within $[0, 3D]$. The power delay profile is exponential, i.e.,

$E[\alpha_p^2] = e^{-\tau_p/\tau_{rms}}$. The simulation assumes $M = 256$ samples per frame, 63 samples per monocycle of duration $D = 63T_c$, ray delays that are multiple of T_c, i.e., $\tau_p = pT_c$, and delay spread $\tau_{rms} = 0.7663D$. With this model rays can appear in clusters of duration smaller than the pulse duration. The proposed frequency domain estimator is within 0.5 dB from the ideal matched filter curve. It should be noted that we estimate the channel and perform combining only over the frequency bins that have amplitude above 10% of the monocycle peak (17 out of 256). We also report the performance that is obtained with the time-domain rake receiver that combines one, two or three separable rays. The rake receiver is implemented according to the algorithm that is described in the Appendix of [6] (in particular, formulas (31) and (32) of [6]) assuming 100 training bits. This algorithm is based on the assumption of a resolvable channel. Nevertheless, the procedure that searches the ray delays is quite complex. Further, the performance penalty is significant since it is incapable of capturing the channel energy that is associated to clusters of rays of duration smaller than D.

In Fig. 3 we report BER versus number of users performance. We deploy random (pseudo noise) short codes of length L=8. Longer codes shall yield improved performance. We take 25 samples per monocycle and M=256 samples per frame. The training sequence has length N=150 bits. Users' channels are independent with $N_P = 5$ tap delays uniformly distributed in $[0, 2D]$, and power delay profile exponential with $\tau_{rms} = 1.5326D$. Both the synchronous case (dashed curves) and the asynchronous users case (solid curves) are considered. For the asynchronous case, the users' time delays are independent and uniformly distributed within a frame interval.

Fig. 3 shows that a sensible performance degradation arises in the presence of multiple users if no MAI cancellation is deployed both with ideal, and practical channel estimation and frame synchronization (**R** is assumed to be the identity matrix in this case). Performance can be significantly improved by deploying the proposed frequency domain MAI cancelling algorithm. In both the ideal, and the practical case the interference correlation matrix has been estimated over the training sequence of length 150 bits. Diagonal loading with a factor 0.5 has been used. Further, to keep the complexity at low levels, we have done combining and interference cancellation only over the frequency bins that have amplitude higher than 10% of the desired user signature peak.

5. CONCLUSIONS

We have considered a frequency domain processing approach to detection, channel estimation, and MAI cancellation for impulse radio CDMA systems. Frequency domain channel estimation shows fast convergence. The proposed receiver that includes practical estimation of the channel and of the frame timing exhibits near matched filter bound performance and robustness to MAI. The approach can be extended to time-hopping based systems.

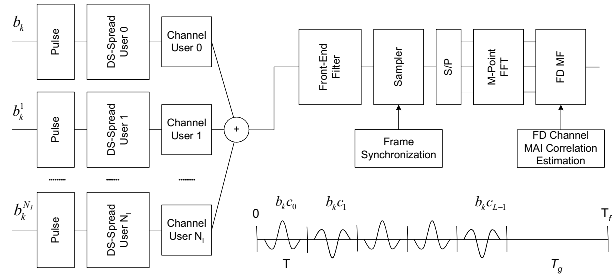

Fig. 1. Impulse modulated system with frequency domain (FD) processing, and frame structure.

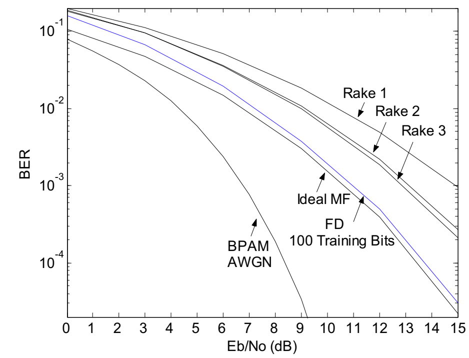

Fig. 2. BER in a single user scenario with ideal matched filtering, with practical frame synchronization and frequency domain (FD) channel estimation, and with a practical rake receiver with up three 3 fingers.

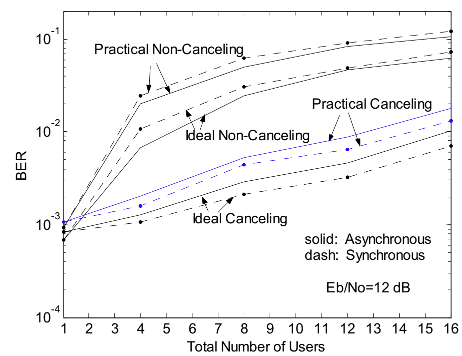

Fig. 3. Multiuser scenario with synchronous and asynchronous equal power users. Random codes of length 8. Training with 150 bits.

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