Frequency Domain Channel Estimation and Detection for Impulse Radio Systems

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***Abstract*—This paper deals with channel estimation, and detection in ultra wide band (UWB) bi-phase impulse modulated systems. We address the single user and the multiuser scenario assuming a direct sequence spreading code division multiple access (DS-CDMA) scheme. 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. The approach can be extended to time-hopped impulse radio systems.**

***Keywords*—CDMA, Channel estimation, Frequency domain processing, Impulse modulation, Interference Cancellation, Multiuser Detection, Synchronization, Ultra wide band (UWB).**

I. INTRODUCTION[1]

THIS paper deals with synchronization, channel estimation, and detection in impulse radio systems. Several combinations of modulation, and user multiplexing schemes have been proposed for impulse radio communications [3]. The common attractive feature 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. Most of the work has focused so far on schemes that deploy time hopping spreading codes with pulse position modulation. Instead, in this paper we assume bi-phase pulse modulation (BPAM) in conjunction with direct sequence code division multiplexing of users (DS-CDMA) [3], [7]. 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 [1]. 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 [13] 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 [6], [13]. However, this assumption translates into deep performance losses in the non-rare event of clusters of non-separable rays.

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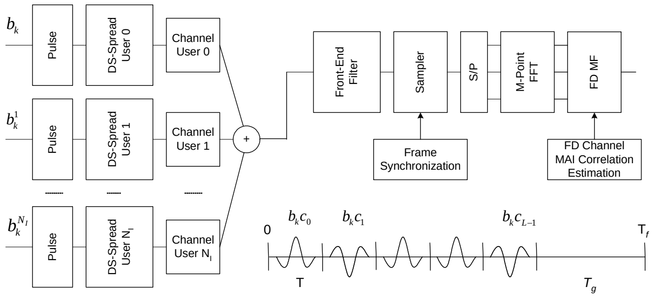

Fig. 1. Impulse modulated system with frequency domain (FD) processing, and frame structure.

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.

Motivated by the above considerations, we propose a frequency domain approach to channel estimation, and detection [2], [12]. The approach is single user based. However, it can include the capability of rejecting the MAI interference. It has been derived from the observation that the optimal matched filter receiver can be equivalently implemented in the frequency domain. 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. Frame timing is crucial. In this paper we also address this problem and we describe a frame timing algorithm.

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.

II. FREQUENCY DOMAIN PROCESSING

In our system model (Fig. 1) we assume bi-phase pulse amplitude (BPAM) modulation [9] such that the signal transmitted by user u can be written as

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

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 user u, and T_f is the bit period (frame duration). We further deploy direct sequence spreading to accommodate for multiplexing of users [3]. 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 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 time dispersion introduced by the channel. If we chose $T \geq D + T_{ch}$ we could avoid also the inter-pulse interference at the expense of a transmission rate penalty. However, we do not restrict ourselves to this case.

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 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 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 case of colored noise can also be considered which yields a different decision metric as described in the next section. 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, 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 $[0\ T_h)$ with $T_h < T_g$, 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., the channel is not necessarily separable. Indeed, other models are possible as comprehensively described in [4].

A. Single User Case

Assuming a single user, with ideal frame synchronization, and under the above assumptions, no inter-symbol and no multiuser interference arises at the receiver side. Thus, the optimal receiver can operate in a symbol by symbol fashion by computing the correlation between the received signal and the real equivalent impulse response $g_{EQ}(t)$, to obtain $z(kT_f) = \int_0^{T_f} y_k(t) g_{EQ}(t) dt$ with $y_k(t) = y(t + kT_f)$ for $0 \leq t < T_f$ [1]. Then, we make a threshold decision, i.e., $\hat{b}_k = sign\{z(kT_f)\}$. This receiver is referred to as matched filter receiver.

If we assume discrete-time processing, the received signal is sampled at the output of the front-end analog filter at sufficiently high rate to obtain $y_k(nT_c)$ with $T_c = T_f / M$. Assuming frame synchronization, the received frame of samples reads

$$y_k(nT_c) = b_k g_{EQ}(nT_c) + \eta_k(nT_c) \qquad n = 0,....,M-1 \quad (5)$$

where $\eta_k(nT_c)$ are zero mean Gaussian random variables. Then, the decision statistic is generated as follows

$$z(kT_f) = \sum_{n=0}^{M-1} T_c y_k(nT_c) g_{EQ}(nT_c) \,. \quad (6)$$

To implement (6) we need to estimate the channel impulse response. Typically, estimation is performed in time domain using a training bit sequence. 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. Maximum likelihood time-domain channel estimation is described in [6], [13] under the assumption of a tapped delay line channel model.

In this paper we take a different approach by proposing channel estimation, and detection in the frequency domain. To proceed, we can interpret (6) as the cross-energy between two discrete time signals that are periodic of $T_f = MT_c$. By Parseval theorem, we can equivalently obtain the decision statistic by operating in the frequency domain as follows

$$z(kT_f) = \frac{1}{MT_c} \sum_{n=0}^{M-1} Y_k(f_n) G_{EQ}^*(f_n) \quad (7)$$

where $Y_k(f_n)$, $G_{EQ}^*(f_n)$ for $f_n = n/MT_c$, $n = 0,...,M-1$, are the M-point discrete Fourier Transform (DFT) outputs of the received frame, and of the matched filter impulse response[2]. The DFT can be efficiently implemented via a fast Fourier Transform (FFT). To obtain (7) we need to estimate $G_{EQ}(f_n)$. The attractive feature in (7) is the fact 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 as we will describe in detail in Section III. 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

$$G(f_n) = G_M(f_n) \sum_{m=0}^{L-1} c_m e^{-j2\pi f_n mT} \,. \quad (8)$$

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. In other words, it does not rely on the assumption of the channel to have a tapped delay line impulse response. Indeed, it has to be pointed out that the frequency domain approach requires frame synchronization. We will propose a practical solution to this problem in Section IV.

B. Multiuser Case

In the presence of N_I other users (interferers), we still pursue a single user detection approach, i.e., the receiver wants to detect the

[2] * denotes the complex conjugate operator. The M-point DFT is defined as $A(f_n) = \sum_{k=0}^{M-1} T_c a(kT_c) e^{-j2\pi kT_c f_n}$ with $f_n = n/MT_c$, $n = 0,...,M-1$.

desired user's bits b_k only. For this purpose, we acquire frame synchronization with the desired user; we estimate its channel equivalent frequency response, and finally, we run the matched filtering operation. In practical scenarios MAI is present due to the deployment of non-orthogonal spreading codes, to the presence of asynchronous users and channel time dispersion [9].

Our goal is to operate in the frequency domain by introducing an appropriate modification of (7) that takes into account the presence of the MAI. We start by collecting M samples at the output of the analog front-end filter in correspondence with the k-th frame of the desired user. We deploy an M-point DFT, obtaining in the presence of MAI

$$Y_k(f_n) = b_k G_{EQ}(f_n) + I_k(f_n) + N_k(f_n) \qquad n = 0,...,M-1 \tag{9}$$

where $N_k(f_n)$ is the DFT of the noise samples $\eta_k(nT_c)$, while $I_k(f_n)$ is the MAI term. 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, respectively reads

$$I_k(f_n)_{async} = \sum_{u=1}^{N_I}\sum_{m=0}^{1} b_{k-m}^u G_{EQ}^u(f_n,m),\ I_k(f_n)_{sync} = \sum_{u=1}^{N_I} b_k^u G_{EQ}^u(f_n,0) \tag{10}$$

where $G_{EQ}^u(f_n,m)$ is a function of the users' time delay, transmitted waveform, and channel. Details are reported in [12]. 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 [12]. Assuming the transmitted bits to be i.i.d. and equally likely, the process has zero mean, and *time-frequency* correlation matrix equal to[3]

$$\mathbf{R}(k,m) = E[\mathbf{Z}_k \mathbf{Z}_m^\dagger] \tag{11}$$

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 $|m-k| > 0$ (see [12]).

Now, let us collect the elements of $Y_k(f_n)$, in the vector $\mathbf{Y}_k = [Y_k(f_0),...,Y_k(f_{M-1})]^T$ while the elements of $G_{EQ}(f_n)$ in the vector $\mathbf{G}_{EQ} = [G_{EQ}(f_0),...,G_{EQ}(f_{M-1})]^T$. Then, if we apply the maximum-likelihood criterion in the frequency domain, we need to search for the sequence of transmitted bits $\{\hat{b}_k\}$, $k = -\infty,...,+\infty$ (belonging to the desired user) that maximizes the logarithm of the probability density function of the received signal $\{\mathbf{Y}_k\}$ conditional on a given hypothetical transmitted bit sequence, i.e., $\log p(\{\mathbf{Y}_k\} | \{\hat{b}_k \mathbf{G}_{EQ}\})$. It follows that we have to search for the bit sequence of the desired user that maximizes the following log-likelihood function [8]

$$\Lambda(\{\hat{b}_k\}) = -\sum_{k=-\infty}^{\infty}\sum_{m=-\infty}^{\infty} [\mathbf{Y}_k - \hat{b}_k \mathbf{G}_{EQ}]^\dagger \mathbf{R}^{-1}(k,m)[\mathbf{Y}_m - \hat{b}_m \mathbf{G}_{EQ}]\,. \tag{12}$$

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

$$\Lambda(\hat{b}_k) \sim \hat{b}_k \operatorname{Re}\left\{\mathbf{G}_{EQ}^\dagger \mathbf{R}^{-1}(k,k)\mathbf{Y}_k\right\}. \tag{13}$$

[3] ^{T} denotes the transpose operator. $\dagger$ denotes the conjugate and transpose operator.

Therefore, according to (13) the frequency domain receiver operates on a frame by frame basis, and it exploits the frequency correlation of the MAI. The computation in (13) 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

$$z_{IC}(kT_f) = \mathbf{G}_{EQ}^\dagger \mathbf{R}^{-1}(k,k)\mathbf{Y}_k\,. \tag{14}$$

Then, we make a decision on the transmitted bit looking at the sign of (14). Note that (14) is real, given that the quantities involved have Hermitian symmetry.

In the absence of MAI, and with white noise, the correlation matrix is diagonal with diagonal elements equal to the noise variance. In such a case the algorithm collapses to the one that we have described in the previous section. 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. Similar, in spirit, approaches have been proposed for co-channel interference cancellation in systems that deploy receive antenna arrays and use spatial interference decorrelation through combining of the received antenna signals [8], [10], [11].

III. 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 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., $\mathbf{R} = \mathbf{R}(k,k)$.

C. RLS Frequency Domain Estimation of Desired User's Channel

With a training sequence of N bits, the M-bins channel frequency response can be obtained via a recursive least squares (RLS) algorithm that operates independently over the sub-channels [9]. To do so we approximate the frequency response of the equivalent channel of the desired user as $\hat{G}_{EQ}(f_n) \approx G(f_n)\hat{H}(f_n)$ $n = 0,...,M-1$, where $G(f_n)$ denotes the M-point DFT of the desired user's waveform (at frequency f_n), and $\hat{H}(f_n)$ denotes the estimate of the channel frequency response that includes the effect of the front-end filter and is estimated via the RLS algorithm [12].

D. Estimation of the Frequency Domain MAI Correlation Matrix

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 $\hat{\mathbf{R}} = \mathbf{R}(k,k)$. 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 $\{b_i\}$, $i = 0,...,N-1$, is the sequence of known training bits of the desired user. Then, we estimate the correlation matrix as

$$\hat{\mathbf{R}} = 1/N \sum_{i=0}^{N-1} \mathbf{e}_i \mathbf{e}_i^\dagger\,. \tag{15}$$

Further, to introduce a tradeoff between the effects of noise, and the effects of the MAI we add diagonal loading as follows [12]: $\hat{\mathbf{R}} = (1-\rho)\hat{\mathbf{R}} + \rho\sigma_N^2\mathbf{I}$, with $\rho \le 1$. For practical purposes the noise variance can be set to an appropriate value according to the range of operating signal-to-noise ratios.

IV. Frame Synchronization

Our frame synchronization scheme operates in the time domain and uses the training bit sequence $\{b_i\}$ of length N. The method is divided in two steps. First, we acquire coarse synchronization with the desired user's training sequence. Second, we acquire fine timing. The scheme is by no means optimal, but it has been chosen as a good tradeoff between performance and complexity [12].

Recall that frame timing is needed to implement the FD channel estimator.

First Step - Coarse Timing. Coarse timing is obtained by locking on the time instant where the channel exhibits the highest energy. We refer to it as the highest energy channel tap. 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

$$\hat{p}_1 = \arg\max_{p \in \mathbb{Z}} \left\{ \left| M_1(p) \right|^2 \right\} \qquad M_1(p) = \frac{1}{N} \sum_{i=0}^{N-1} b_i \, y(pT_c + iMT_c) \, . \quad (16)$$

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 exactly establishing where the frame is located around the highest energy channel tap. We do not make any assumption on the channel, i.e., we do not assume it to have, for instance, a single or double sided exponential power delay profile. The fine synchronization strategy that we propose in this paper 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 $t_2 = \hat{p}_1 T_c + \hat{p}_2 M_w$ for a given $\hat{p}_2 \in \{-M/M_w, ..., M/M_w\}$. The integer $\hat{p}_2$ is determined via the following maximization

$$\hat{p}_2 = \arg\max_{p \in \{-M/M_w, ..., M/M_w\}} \sum_{i=0}^{M/M_w - 2} M_2(pM_w + iM_w) \qquad (17)$$

$$M_2(p) = \frac{1}{2M_w} \sum_{k=-M_w}^{M_w - 1} | M_1(p + \hat{p}_1 + k) |^2 \, . \qquad (18)$$

Note that (18) 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, (17) 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.

V. Considerations on Complexity

The complexity of the proposed estimation and detection approach is a function of the system design parameters in terms of occupied bandwidth, and frame duration (symbol rate). Typically, the frame duration is much larger than the pulse duration. Nevertheless, it has to be noted that the sampling rate requirement can be high depending on the pulse bandwidth. Intrinsically, this is a source of complexity for any UWB system. In our approach the number of samples per frame that we need to process, and consequently the number of DFT points, is directly related to the channel resolvability capability of the receiver. The higher is the desired time resolution for channel estimation, the higher the number of samples (and DFT points) per frame we need to process. If we do not possess any a priori knowledge on the channel we need to deploy the DFT over the whole frame. However, if the channel is sparse, and manifests itself as a number of separable clusters, we can simplify complexity by deploying a *pruned* DFT. That is, we can set to zero the input samples that do not carry useful signal energy. The energy content of the received signal can be determined re-using the synchronization metrics.

We point out that in certain conditions it is possible, and convenient, to deploy channel estimation in the frequency domain while detection in the time domain. As an example consider the case when the channel exhibits a small number of resolvable rays *K*. Then, the time-domain rake receiver needs to combine only *K* fingers [1], [5]. Following our approach, we can first perform channel estimation in the frequency domain. Then, we can compute the rake fingers (delays and amplitudes) via an inverse DFT. Finally, detection can be performed in the time domain by combining the rake fingers at low symbol rate. Indeed, since no a priori knowledge of the channel is available, we need to run channel estimation at high sampling rate. However, during training it is possible to lower the requirements on the analog-to-digital converter sampling rate with the approach that has been proposed in [5] and that is based on the idea of repeating K_R times the training sequence of length *N*, and using a polyphase sampling filter bank of size *K*. In combination or in alternative to this technique we can also resort to conventional interpolation techniques as suggested in [6].

VI. Performance Evaluation

Performance is assed via simulations. A *N* bits training sequence is followed by 1000 information bits that are used to estimate the bit-error-rate for a given channel/estimation realization. Averaging over 1000 channel realizations is then performed.

E. Single User Case

We start looking at the single user case. We assume no spreading, i.e., we deploy a length *L*=1 code, and we assume to deploy a guard time sufficiently long to absorb the channel time dispersion. The channel model in (4) has $N_P = 10$ paths that have uniformly distributed delays within [0, 3D], Rayleigh amplitude, and equally likely sign flip. The power delay profile is exponential, i.e., $E[\alpha_p^2] = e^{-\tau_p / \tau_{rms}}$ [6]. 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.

In Fig. 2, we assume perfect frame synchronization and we show that the RLS frequency domain channel estimator convergence is very fast and accurate. The forgetting factor is set to 0.999. We also plot the curves that are obtained by combining only the frequency bins for which $|G(f_n)| \geq 0.1 \max\{|G(f_k)|\}$ (curves labeled with Above 10% Amplitude Bins). The performance improvement for the 10% curves can be explained by the fact that estimation over frequency bins that have small signal energy is poor, and can negatively affect the BER performance. It should be noted that for the 10% curves, the number of bins over which channel estimation is actually performed is only 17 out of 256 (taking also into account the Hermitian symmetry).

In Fig. 3, the performance of the overall algorithm that combines frame synchronization, and channel estimation is shown. We deploy 100 training bits. The proposed frequency domain estimator is within 0.5 dB from the ideal matched filter curve. 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 [13] (in particular, formulas (31) and (32) of [13]) assuming 100 training bits. This algorithm is based on the assumption of a separable 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*.

F. Multiuser Case

The performance of the proposed algorithm in a multiuser scenario is shown in Fig. 4. We deploy random (pseudo noise) short codes of length $L = 8$. Longer codes shall yield improved performance. We take 25 samples per monocycle of duration *T*=*D*, and *M*=256 samples per frame of duration $T_f = LT + T_{ch}$. The training sequence has length *N*=150 bits. Users' channels are independent with $N_P = 5$, tap delays uniformly distributed in

$[0,\ 2T]$, 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. 4 shows that a sensible performance degradation arises in the presence of multiple users if the algorithm that does not cancel the MAI is deployed (algorithm of Section II.A). Note that we simulate also an overloaded system scenario, i.e., we allocate more than L=8 users. Curves labelled with ideal have been obtained assuming ideal knowledge of the channel and frame timing of the desired user, while the curves labelled with practical have been obtained by estimating both the frame timing, and the channel (with the frequency domain RLS algorithm) with 150 bits known bits. Performance can be significantly improved by deploying the proposed frequency domain MAI cancelling algorithm of Section II.B. In both the ideal, and the practical case the interference correlation matrix has been estimated over the training sequence of length 150 bits according to (15). Diagonal loading with a factor 0.5 has been used. Further, as for Fig. 2-3, we combine, and do interference cancellation, only over the frequency bins for which $|G(f_n)| \geq 0.1\max\{|G(f_k)|\}$, to keep the complexity at low levels. We point out that the training parameters have been kept fixed for all scenarios. Further improvements are expected by deploying longer training sequences, by optimizing the parameters, and by performing interference cancellation over a higher number of freq. bins [12].

VII. Conclusions

We have considered the synchronization, channel estimation, and detection problem in impulse radio systems with DS-CDMA. We have proposed to carry out channel estimation via a RLS algorithm in the frequency domain. Frame synchronization is acquired in the time domain. Detection can be performed in the frequency domain, and it deploys the estimated channel frequency response. In the presence of multiple access interference, the frequency domain detection approach allows including the capability of canceling the MAI exploiting the MAI correlation in the frequency domain. The estimation of the MAI correlation matrix has also been considered. An interesting aspect of the proposed channel estimation approach is its moderate complexity compared to maximum likelihood time domain channel estimation. This is due to the deployment of a fast Fourier transform, and to the fact that channel estimation needs to be performed only over a small fraction of the overall number of frequency bins. In fact only the sufficiently high energy frequency bins contribute to the detection metric, and need to be processed.

Several numerical results have been reported and demonstrate that the proposed approach exhibits fast convergence, and high performance with or without synchronous/asynchronous multiple access interference. Finally, we point out that the proposed estimation, and detection approach can be extended to impulse modulation systems that deploy time-hopping.

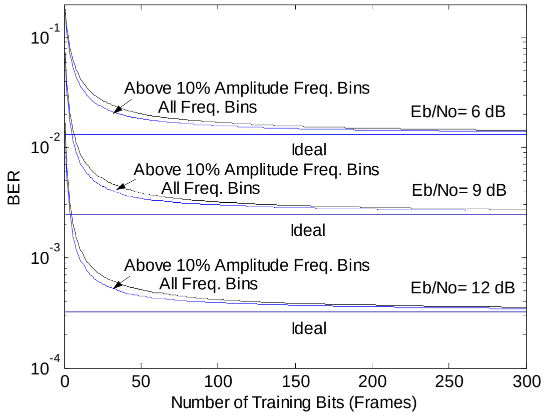

Fig. 2. BER as a function of training bits number in the single user case.

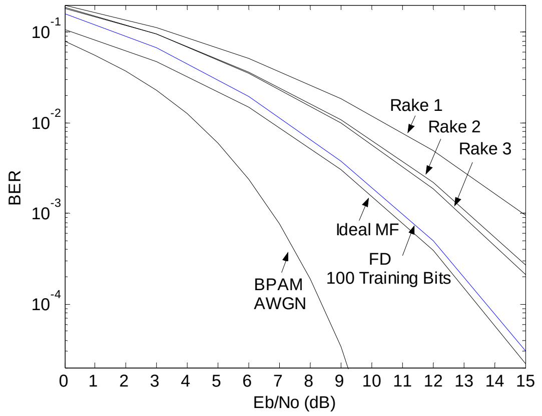

Fig. 3. BER in a single user scenario with ideal matched filtering, with practical frame synchronization and frequency domain (FD) channel estimation, and with a practical rake receiver with up three 3 fingers.

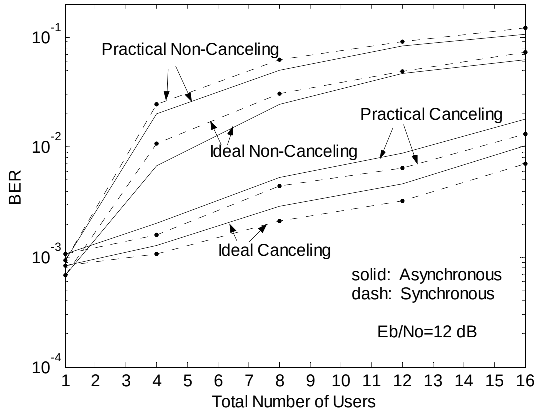

Fig. 4. Multiuser scenario with synchronous and asynchronous equal power users. Random codes of length 8. Only frequency bins with at least 10% amplitude relative to the maximum are used.

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