

Synchronization for Multiuser Wide Band Impulse Modulation Systems in Power Line Channels with Unstationary Noise

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Abstract—In this paper we consider a wide band multiuser impulse modulation based system combined with DS-CDMA. We focus on the synchronization problem and we derive an optimal maximum likelihood metric for multiuser synchronization assuming an unstationary Gaussian noise model. Then, with the appropriate simplifications we obtain a simple implementation that is capable of achieving near optimal performance. The results show that the scheme allows for robust transmission over highly dispersive PL channels in the presence of impulse noise and multiple access interference.

Keywords—Wide band power line communication systems, Impulse modulation, Multiuser detection, Synchronization, Ultra wide band (UWB) systems.

I. INTRODUCTION

In this paper we consider a power line (PL) communication system that uses carrier less wide band impulse modulation with bandwidth beyond 20 MHz [1]-[4]. Indoor applications such as local area networks, peripheral office connectivity, and home/industrial control are considered. The basic idea behind impulse modulation is to convey information by mapping an information symbol stream into a sequence of short duration pulses (referred to as monocycles) [5]. The monocycle can be shaped to avoid the low frequencies where we experience higher levels of background noise. User multiplexing is obtained via direct sequence code division multiple access (DS-CDMA) using signature waveforms that are a repetition of time delayed and weighted monocycles [6]. We point out that since our system deploys a fractional bandwidth (ratio between signaling bandwidth and center carrier) larger than 20%) it can be classified as an ultra wide band (UWB) system according to the FCC.

It has been shown that this system provides robust performance in highly time dispersive PL channels, in the presence of impulse noise and multiple access interference. In particular in [4] we have described a frequency domain receiver and we have assessed the problem of estimating the channel frequency response.

In this new contribution we deal with the synchronization problem. The multiple access channel is assumed asynchronous which means that communication is from one node to another, and nodes are not temporally synchronous. The detector needs to acquire frame synchronization with the users that it intends to demodulate. Single user

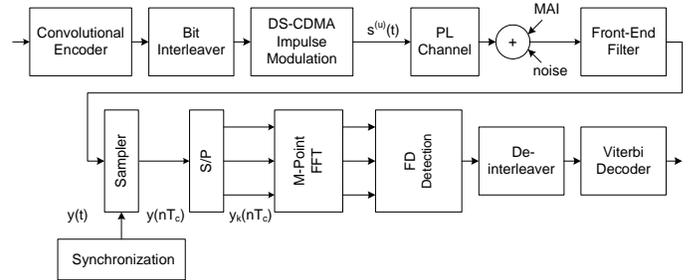


Fig. 1. Wide band impulse modulated system

synchronization, and in particular estimation of the multipath channel profile for typical UWB radio channels has been studied in [11]-[12].

Herein, we consider optimal maximum likelihood synchronization with training in the presence of multiple user transmission and unstationary noise. In PL channels, both stationary and unstationary noise components are present, e.g., colored background noise and impulsive noise [8]-[9]. To derive the synchronization metrics, we model the noise as an unstationary colored Gaussian process. Simplified metrics are also obtained under certain assumptions.

II. SYSTEM MODEL

We consider the system model described in [4] (Fig.1). For clarity, we summarize the main characteristics. A number of nodes (users) wish to communicate sharing the same PL network. Communication is from one node to another node, such that if other nodes simultaneously access the medium they are seen as potential interferers. Users' multiplexing is obtained in a CDMA fashion allocating the spreading codes among the users. The signal transmitted by user u can be written as

$$s^{(u)}(t) = \sum_k \sum_{i \in C_u} b_k^{(u,i)} g^{(u,i)}(t - kT_f) \quad (1)$$

where $g^{(u,i)}(t)$ is the waveform (signature code) used to convey the i -th information symbol $b_k^{(u,i)}$ of user u that is transmitted during the k -th frame. Each symbol belongs to the pulse amplitude modulation (PAM) alphabet, and it carries $\log_2 M_S$ information bits where M_S is the number of PAM levels, e.g., with 2-PAM $b_k^{(u,i)}$ has alphabet ± 1 . T_f is the

symbol period (frame duration). C_u denotes the set of signature code indices that are allocated to user u . Thus, user u can adapt its rate by transmitting C_u information symbols per frame. The signature code comprises the weighted repetition of $L \geq 1$ narrow pulses (monocycles):

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

where $c_m^{(u,i)} = \pm 1$ are the codeword elements (chips), and T is the chip period. The monocycle $g_M(t)$ can be appropriately designed to shape the spectrum occupied by the transmission system. In this paper we consider the second derivative of the Gaussian pulse with duration $D \approx 126$ ns, and -30 dB bandwidth of 40 MHz [4]. An interesting property is that its spectrum does not occupy the low frequencies where we experience higher levels of man-made background noise.

We choose the chip period $T \geq D$, and we further insert a guard time T_g between frames to cope with the channel time dispersion. Thus, the frame duration is $T_f = LT + T_g$.

Distinct codes are allocated to distinct users. In our design the codes are defined as follows:

$$c_m^{(u,i)} = c_{1,m}^{(u)} c_{2,m}^{(i)} \quad m = 0, \dots, L-1 \quad (3)$$

where $\{c_{1,m}^{(u)}\}$ is a binary (± 1) random sequence of length L allocated to user u , while $\{c_{2,m}^{(i)}\}$ is the i -th binary (± 1) Walsh-Hadamard sequence of length L . It should be noted that with this choice each node can use all L Walsh codes, which yields a peak data rate per user equal to $R = L/T_f$ symbol/s. It approaches $\log_2 M_s / T$ bit/s with long codes. Clearly, while the signals of a given user are orthogonal, the ones that belong to distinct transmitting nodes are not. The random code $\{c_{1,m}^{(u)}\}$ is used to randomize the effect of the MAI.

The signals of the N_u users propagate through distinct channels with impulse response $g_{CH}^{(u)}(t)$. The received real signal, at a given node, reads

$$y(t) = x(t) + w(t) = \sum_{u=1}^{N_u} x^{(u)}(t - \Delta^{(u)}) + w(t), \quad (4)$$

$\Delta^{(u)}$ is the time offset of user u , while $x^{(u)}(t)$ is the equivalent impulse response of user u that is obtained from the convolution of the transmitted waveform $s^{(u)}(t)$ and the channel impulse response $g_{CH}^{(u)}(t)$,

$$x^{(u)}(t) = \int_{-\infty}^{+\infty} d\tau g_{CH}^{(u)}(\tau) s^{(u)}(t - \tau). \quad (5)$$

$w(t)$ denotes the additive background noise at the receiver input.

It should be noted that the channel time dispersion introduces both inter-code (ICI) and multiple access interference (MAI) at the receiver.

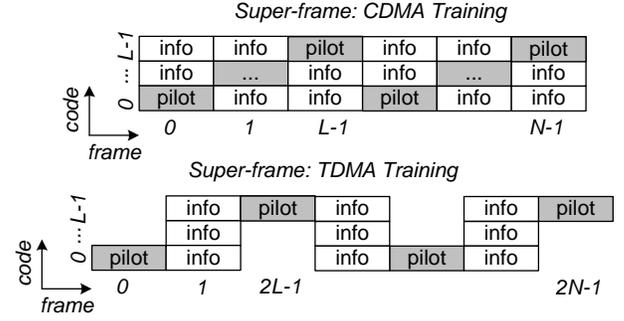


Fig. 2. Training methods.

A. Training

The synchronization algorithm described in this paper is based on sending a training pattern. The training signal is sent via the assignment of a pilot code (one of the codes defined by (3)) to each user. If we assume packet transmission the training word can precede the information packet (TDMA training), or it can be embedded in the packet itself (CDMA training). This is illustrated in Fig.2. The received signal can be written as

$$y(t) = \sum_{u=1}^{N_u} x_{TR}^{(u)}(t - \Delta^{(u)}) + \sum_{u=1}^{N_u} x_{INFO}^{(u)}(t - \Delta^{(u)}) + w(t) \quad (6)$$

where we highlight the fact that the received signal comprises both $x_{TR}^{(u)}(t)$ that is associated to the training signal, and $x_{INFO}^{(u)}(t)$ that is associated to the information signal. If training is embedded in each information packet, $x_{INFO}^{(u)}(t)$ is seen as interference by the synchronization algorithm.

B. Channel Model

The synchronization algorithm that we derive does not rely on a specific channel model. To evaluate performance we use a statistical channel derived from the multipath model in [7] as described in [4]. Distinct users experience distinct channels with identical maximum time dispersion.

C. Noise Model

To derive the receiver algorithms we treat the noise as an unstationary colored Gaussian real process with zero mean. In PL channels both stationary, and unstationary noise components can be present. They include colored stationary background noise, cyclostationary noise synchronous to the mains, and unstationary impulsive noise [8]. Often, the two terms Gaussian model is used [9]. It accounts for a stationary component and an impulsive component that occurs with a certain probability. Herein, the two terms Gaussian model is generalized to the continuous time domain as follows

$$w(t) = w_T(t) + \alpha(t)w_{IM}(t) \quad (7)$$

where $w_T(t)$ is the stationary contribution, and $\alpha(t)w_{IM}(t)$ is the impulse noise component. The multiplicative process $\alpha(t)$ accounts for the presence or absence of impulse noise. That is, at time instant t the random variable $\alpha(t)$ is a Bernoulli random variable with parameter p and alphabet $\{0,1\}$. We refer to it as Bernoulli process. $w_T(t)$ and $w_{IM}(t)$ are treated as independent zero mean Gaussian, with correlation

$$\kappa_T(t; t') = \kappa_T(t - t') = E[w_T(t)w_T(t')],$$

$$\kappa_{IM}(t; t') = \kappa_{IM}(t - t') = E[w_{IM}(t)w_{IM}(t')]. \quad (8)$$

Conditional on the Bernoulli process, the impairment is a Gaussian process with correlation

$$\kappa_w(t; t') = E[w(t)w(t') | \alpha(t), t \in \mathbb{R}] = \kappa_T(t; t') + \alpha(t)\alpha(t')\kappa_{IM}(t; t'). \quad (9)$$

The model describes impulse spikes of certain duration, power decay profile, and colored spectral components. Note that (7) is in general unstationary although the single processes $w_T(t)$ and $w_{IM}(t)$ are herein modeled as stationary according to (8).

D. Receiver

The baseline receiver for the impulse modulated system is the correlation receiver. It correlates the received signal with the desired user equivalent signature waveform [1]. Improved detection performance can be obtained with a frequency domain approach as described in [4]. All these receivers require the knowledge of the users time offsets $\Delta^{(u)}$. In the following we derive maximum likelihood synchronization algorithms that allow to estimate both the offsets $\Delta^{(u)}$, and the users channel impulse responses $g_{CH}^{(u)}(t)$, assuming that a training signal is transmitted by each user.

III. ML JOINT SYNCHRONIZATION

In this section we describe a general framework from which we derive the maximum likelihood (ML) joint synchronization metric. We start from the received signal model in (6), and we define the vector of time offsets

$$\Delta = [\Delta^{(1)}, \Delta^{(2)}, \dots, \Delta^{(N_U)}]^T \quad (10)$$

whose N_U elements are the time offsets $\Delta^{(u)}$, and the vector of channel impulse responses

$$\mathbf{c}(t) = [g_{CH}^{(1)}(t), g_{CH}^{(2)}(t), \dots, g_{CH}^{(N_U)}(t)]^T. \quad (11)$$

Further, we model the overall impairment (interference plus noise) $i(t) = \sum_{u=1}^{N_U} x_{INFO}^{(u)}(t) + w(t)$ as a zero mean unstationary Gaussian process with correlation function

$$\kappa(t; t') = E[i(t)i(t')] = \kappa_{INFO}(t; t') + \kappa_w(t; t') \quad (12)$$

where $\kappa_{INFO}(t; t')$ is the correlation function of the information signals, i.e., interference seen by the known training signals, which is not null with CDMA training, while $\kappa_w(t; t')$ is the correlation of the background noise. This is Gaussian if the background noise follows the two terms mixture model conditionally on the knowledge of the Bernoulli process realizations. The interference is only approximately Gaussian.

Then, the ML estimates of Δ , and $\mathbf{c}(t)$ are the values $\hat{\Delta}$, $\hat{\mathbf{c}}(t)$ that minimize the log-likelihood function [13]-[14]

$$\Lambda(y(t) | \Delta, \mathbf{c}(t)) = \int_{-\infty}^{+\infty} dt \int_{-\infty}^{+\infty} dt' \left(y(t) - \sum_{u=1}^{N_U} x_{TR}^{(u)}(t - \Delta^{(u)}) \right) \times \kappa^{-1}(t; t') \left(y(t') - \sum_{u=1}^{N_U} x_{TR}^{(u)}(t' - \Delta^{(u)}) \right) \quad (13)$$

where $\kappa^{-1}(t; t')$ is the inverse function (that we assume to exist) of (12) with respect to the product operator [14], i.e.,

$$\int_{-\infty}^{+\infty} dt' \kappa(t; t') \kappa^{-1}(t'; t'') = \delta(t; t'') \quad (14)$$

where $\delta(t; t'')$ is the unit function in the vector space of bi-dimensional function to which $\kappa(t; t')$, and $\kappa^{-1}(t'; t'')$ belong.

It can be shown, after some manipulation, that the minimization of $\Lambda(y(t) | \Delta, \mathbf{c}(t))$ is equivalent to the minimization of the following likelihood function

$$\hat{\Lambda}(\Delta, \mathbf{c}(t)) = \int_{-\infty}^{+\infty} d\tau \int_{-\infty}^{+\infty} d\tau' \mathbf{c}^T(\tau) \mathbf{R}(\Delta, \tau, \tau') \mathbf{c}(\tau') - \int_{-\infty}^{+\infty} d\tau \mathbf{c}^T(\tau) \chi(\Delta, \tau) - \int_{-\infty}^{+\infty} d\tau \chi^T(\Delta, \tau) \mathbf{c}(\tau). \quad (15)$$

To obtain (15) we have defined the correlation matrix of the user training waveforms $s_{TR}^{(u)}(t)$

$$\mathbf{R}(\Delta, \tau, \tau') = \begin{bmatrix} R(\Delta^{(1)}, \Delta^{(1)}, \tau, \tau') & \dots & R(\Delta^{(1)}, \Delta^{(N_U)}, \tau, \tau') \\ \vdots & \ddots & \vdots \\ R(\Delta^{(N_U)}, \Delta^{(1)}, \tau, \tau') & \dots & R(\Delta^{(N_U)}, \Delta^{(N_U)}, \tau, \tau') \end{bmatrix} \quad (16)$$

where each element of the matrix is given by

$$R^{(u, u')}(\Delta^{(u)}, \Delta^{(u')}, \tau, \tau') = \int_{-\infty}^{+\infty} dt \int_{-\infty}^{+\infty} dt' s_{TR}^{(u)}(t - \Delta^{(u)} - \tau) \kappa^{-1}(t; t') s_{TR}^{(u')}(t' - \Delta^{(u')} - \tau'). \quad (17)$$

Further, we have defined the vector of mutual correlation between the received signal and the transmitted waveform

$$\chi(\Delta, \tau) = [\chi^{(1)}(\Delta^{(1)}, \tau), \chi^{(2)}(\Delta^{(2)}, \tau), \dots, \chi^{(N_U)}(\Delta^{(N_U)}, \tau)]^T \quad (18)$$

where each element is given by

$$\chi^{(u)}(\Delta^{(u)}, \tau) = \int_{-\infty}^{+\infty} dt \int_{-\infty}^{+\infty} dt' s_{TR}^{(u)}(t - \Delta^{(u)} - \tau) \kappa^{-1}(t; t') y(t'). \quad (19)$$

To proceed in the derivation of the algorithm, we define

$$\lambda(\Delta, \tau) = \int_{-\infty}^{+\infty} d\tau' \mathbf{R}^{-1}(\Delta, \tau, \tau') \chi(\Delta, \tau'). \quad (20)$$

$\chi(\Delta, \tau)$ can be rewritten as

$$\chi(\Delta, \tau) = \int_{-\infty}^{+\infty} d\tau' \mathbf{R}(\Delta, \tau, \tau') \lambda(\Delta, \tau') \quad (21)$$

where $\int_{-\infty}^{+\infty} d\tau' \mathbf{R}^{-1}(\Delta, \tau, \tau') \mathbf{R}(\Delta, \tau', \tau'') = \mathbf{I} \delta(\tau, \tau'')$, and \mathbf{I} is the identity matrix of size $N_U \times N_U$.

Now, if we substitute (21) in (15), we add and subtract the term $\int_{-\infty}^{+\infty} d\tau \int_{-\infty}^{+\infty} d\tau' \lambda^T(\Delta, \tau) \mathbf{R}(\Delta, \tau, \tau') \lambda(\Delta, \tau')$, and we group equal terms, we obtain a new form for the likelihood function

$$\hat{\Lambda}(\Delta, \mathbf{c}(t)) = \underbrace{\int_{-\infty}^{+\infty} d\tau \int_{-\infty}^{+\infty} d\tau' [\mathbf{c}(\tau) - \lambda(\Delta, \tau)]^T \mathbf{R}(\Delta, \tau, \tau') [\mathbf{c}(\tau') - \lambda(\Delta, \tau')]}_A - \underbrace{\int_{-\infty}^{+\infty} d\tau \int_{-\infty}^{+\infty} d\tau' \lambda^T(\Delta, \tau) \mathbf{R}(\Delta, \tau, \tau') \lambda(\Delta, \tau')}_{B}. \quad (22)$$

A and B in (22) are non-negative quadratic forms such that they are always larger than or equal to zero (the proof is omitted for space limitations). Therefore, $\hat{\Lambda}(\Delta, \mathbf{c}(t))$ is minimized when we choose Δ that maximizes $B(\Delta)$, and then we choose $\mathbf{c}(t)$ that sets $A(\Delta, \mathbf{c}(t))$ to zero. The condition $A(\Delta, \mathbf{c}(t)) = 0$ implies that

$$\mathbf{c}(t) = \lambda(\Delta, t) = \int_{-\infty}^{+\infty} d\tau \mathbf{R}^{-1}(\Delta, t, \tau) \chi(\Delta, \tau). \quad (23)$$

Further, from (20)-(21) we have that

$$B(\Delta) = \int_{-\infty}^{+\infty} d\tau \int_{-\infty}^{+\infty} d\tau' \chi^T(\Delta, \tau) \mathbf{R}^{-1}(\Delta, \tau, \tau') \chi(\Delta, \tau'). \quad (24)$$

Thus, the ML estimates of the user time offsets and channel impulse responses is done in two separate steps. First, we search for the vector of time offsets that maximizes (24), i.e.,

$$\hat{\Delta} = \arg \max_{\Delta} \left\{ \int_{-\infty}^{+\infty} d\tau \int_{-\infty}^{+\infty} d\tau' \chi^T(\Delta, \tau) \mathbf{R}^{-1}(\Delta, \tau, \tau') \chi(\Delta, \tau') \right\}. \quad (25)$$

Then, we determine the vector of channel impulse responses as

$$\hat{\mathbf{c}}(t) = \int_{-\infty}^{+\infty} d\tau \mathbf{R}^{-1}(\hat{\Delta}, t, \tau) \chi(\hat{\Delta}, \tau). \quad (26)$$

Indeed, the ML joint estimation problem in (25) is optimal but complex. Thus, to make it practical we proceed making certain assumptions that allow us to derive a simplified algorithm.

A. Assumption of Orthogonality

We can simplify the synchronization algorithm if we assume

$$\mathbf{R}^{-1}(\Delta, \tau, \tau') = \mathbf{I} \delta(\tau' - \tau) \quad (27)$$

when τ is in $[-T_1, T_1]$, with $T_1 \leq N_{b,TR} T_f$, where $N_{b,TR}$ is equal to the number of training bits, and zero otherwise. This hypothesis holds true if the transmitted waveforms are orthogonal and the noise is white stationary. We obtain a sub-optimal metric that can be implemented by a decoupled search of the time offsets. This is because the joint search for the N_U time offsets $\Delta^{(u)}$ that solves (25) is equivalent to the solution of N_U separate searches, one for each user, as follows

$$\hat{\Delta}^{(u)} = \arg \max_{\Delta^{(u)}} \left\{ \int_{-\infty}^{+\infty} d\tau \left[\chi^{(u)}(\Delta^{(u)}, \tau) \right]^2 \right\}. \quad (28)$$

Under the hypothesis above, if we substitute the expression for the training waveforms according to (1) we obtain the following timing metric

$$\hat{\Delta}^{(u)} = \arg \max_{\Delta^{(u)}} \left\{ \int_{-T_1}^{T_1} d\tau \left[\sum_{k=0}^{N_{b,TR}-1} b_{k,TR}^{(u)} \sum_{m=0}^{L-1} c_{m,TR}^{(u,k)} z(\tau + \Delta^{(u)} + kT_f + mT) \right]^2 \right\} \quad (29)$$

where we have defined $z(t) = y * g_{M-}(t)$. $z(t)$ is the received signal filtered by the impulse matched to the transmit monocycle $g_M(t)$ (front-end analog filter). $b_{k,TR}^{(u)}$, and $c_{m,TR}^{(u,k)}$ are the training bits and spreading chips assigned to user u . If we

set $T_1 = (T_f - LT)/2$ this metric is similar to the one proposed in [10]. If we know the channel duration we can set $T_1 = T_{channel}/2$.

B. Remarks on the Practical Implementation

The practical implementation of the algorithm in (25)-(26) requires the knowledge of the time-variant correlation function that includes the interference and the unstationary background noise. Clearly, the correlation of the unstationary noise component cannot be estimated in practice. However, experimental measurements have shown that the unstationary component is due to impulse noise that can be either cyclostationary or in general repetitive [8]. That is, it manifests itself as a noise spike whose statistics (variance and correlation) over its duration remain identical. Thus, they can be estimated by first estimating when the impulsive noise occurs, and then via time averaging over the occurrence bursts. A significant simplification is obtained by estimating whether a received frame is hit by impulse noise (by looking at the received energy) [4] and then neglecting such frames in the synchronization metric (29).

IV. SIMULATION RESULTS

We assume a frame duration $T_f = 4.096 \mu\text{s}$ and a monocycle of duration $D \approx 126 \text{ ns}$, and -30 dB bandwidth of 40 MHz [4]. The guard time is $T_g = 2.048 \mu\text{s}$. The front-end filter output signal is sampled with period $T_c = 16 \text{ ns}$. Thus, we collect $M = 256$ samples per frame. User synchronization is done with the algorithm described in Section III, while detection, channel estimation, and noise correlation estimation, is done in the frequency domain using an FFT of size 256 [4]. The spreading codes have length $L = 16$ with a chip period $T = 128 \text{ ns}$. The codes are obtained by the chip by chip product of the 16 Walsh codes and a random code for each user to be multiplexed. One code is reserved for training in the CDMA fashion as shown in Fig.2. The code is cyclically shifted so that at each new frame a different Walsh code is assigned to the pilot channel. We consider binary data symbols. A bit interleaved convolutional code of rate $1/2$ and memory 4 is also used. Full rate users are considered, i.e., all 16 Walsh codes are assigned to them. The super-frame spans $N = 540$ frames over which block interleaving is applied. The uncoded data rate is equal to 3.66 Mbit/s/user , while the net rate with coding is half of that. It can be increased with higher level PAM or longer codes.

To test performance we have used the statistical channel model in [4] whose impulse response has duration equal to $4 \mu\text{s}$. Synchronization is done according to the simplified metric of this paper. After having acquired frame timing, detection algorithm is done with the iterative frequency domain joint detector described in [4]. 3 iterations are used. The frequency response of the channel is estimated via an RLS algorithm.

In Fig.3 a single user transmits at full rate (all 16 Walsh codes are used). No channel coding is considered herein. The input background noise is AWG, however the front-end filter colors it. The results show that with ideal synchronization and channel estimation the BER curve is within 0.5 dB from the single code performance (the degradation is due to the ICI). With practical time domain synchronization the performance is

good, and diverges from the ideal only for low bit error rates.

In Fig.4 the set up is identical to that of Fig.3. However, channel coding is applied, and the background noise is impulsive according to the two terms Gaussian model with occurrence probability equal to $\varepsilon = 0.01$. The impulse noise has variance 100 times the first term. The noise spikes last for 4 frames but are asynchronous with them. The impulse noise degrades performance severely. However, it can be compensated by detecting it, and neglecting the frames that are hit by it both in the synchronization metric and in the corresponding trellis sections of the Viterbi decoder. In this case, as the curves labeled with impulse noise compensation show, the performance comes close to the single code bounds.

In Fig.5 three equal power interferers at full rate with channel coding are present. They have a random time phase. The performance with practical synchronization is about 1 dB worse than that with ideal synchronization. We emphasize that here 4 users are transmitting at full rate thus the amount of ICI and MAI is relevant.

V. CONCLUSIONS

We have derived a ML synchronization algorithm for multiuser impulse modulated CDMA systems in the presence of unstationary noise. The practical application of the algorithm in the presence of impulse noise has been discussed, and a simplified metric has been also reported. Simulation results show that the performance of the receiver with practical synchronization and channel estimation is close to the ideal performance, such that the scheme allows for robust transmission over highly dispersive PL channels in the presence of impulse noise and multiple access interference.

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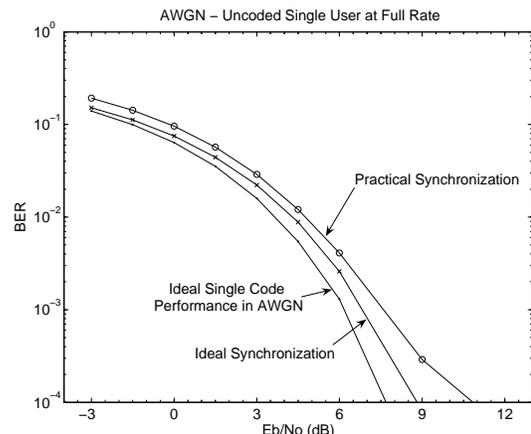


Fig. 3. One uncoded full rate user with ideal and practical synchronization.

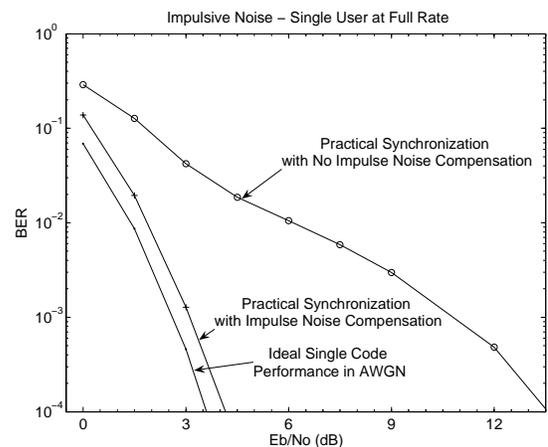


Fig. 4. One full rate coded user with impulsive noise.

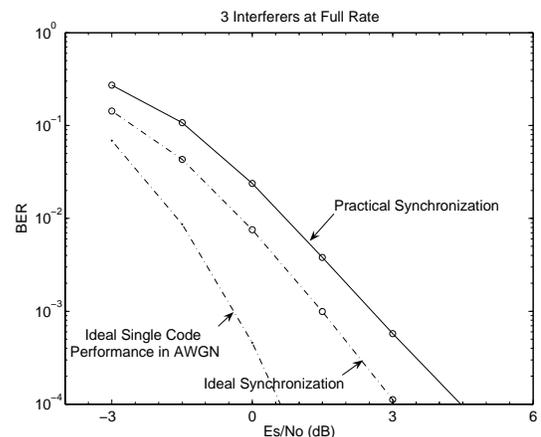


Fig. 5. One full rate coded user with three full rate interferers.

