Synchronization and Channel Estimation for Wide Band Impulse Modulation over Power Line Channels

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Abstract **— In this paper we consider the deployment of impulse modulation for powerline multiuser communications combined with either time hopping, or direct sequence CDMA. We focus on the synchronization and channel estimation problem in the presence of multipath propagation and impulse noise.**

Keywords – impulse modulation, multiple access techniques, power line communications, synchronization algorithms.

I. Introduction[1]

In this paper we consider the synchronization and channel estimation problem in a powerline communication system that deploys wide band impulse modulation. Up to date, impulse modulation has been proposed for application over wide band wireless channels only [8] where it shows interesting properties in terms of simple baseband implementation, and robustness against frequency selective fading, and interference. Most of the current wide band power line transmission proposals consider the deployment of orthogonal frequency division multiplexing (OFDM) [1]. Instead, in this paper we consider a bi-phase impulse modulation scheme that comprises users multiplexing via time-hopping (TH) or direct sequence code division multiple access (DS-CDMA) [2]. The occupied spectrum can be parameterized according to the choice of the pulse shape.

The basic idea behind impulse modulation, is to convey information by mapping an information bit stream into a sequence of short duration pulses. Pulses (referred to as monocycles) are followed by a guard time in order to cope with the channel time dispersion. If the guard time is sufficiently long, no inter-symbol interference arises at the receiver side such that detection simplifies to a matched filter receiver that basically correlates the received signal with a template waveform [6]. The monocycle can be appropriately designed to shape the spectrum occupied by the transmission system. As an example, in Fig. 1 we plot

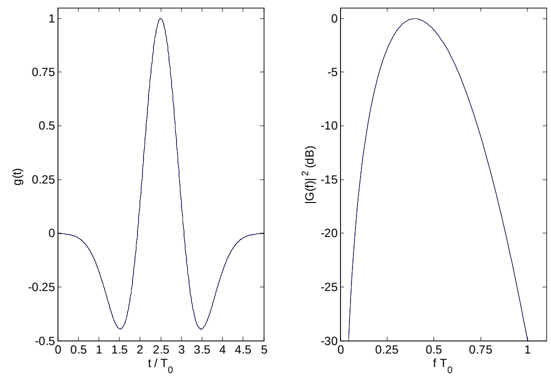

Fig. 1. W-shaped monocycle impulse and frequency response.

the impulse and frequency response of the second derivative of a Gaussian pulse,

$$g(t) = \left(1 - \pi\left((t - D/2)/T_0\right)^2\right) e^{-\frac{\pi}{2}\left((t-D/2)/T_0\right)^2} \quad (1)$$

where $D \approx 5T_0$ is the monocycle duration. An interesting property is that its spectrum does not occupy the low frequencies where we typically experience higher levels of man-made background noise.

In this paper we focus on the multiple access forward link. That is, we consider transmission from a central transmitter to several receivers located in different points across the power grid (Fig. 2). The reverse link, i.e., connection from multiple transmitters to multiple receivers, is treated in [6]. The multiplexing of information belonging to distinct users (receivers) is done with two possible methods. A first method is based on time-hopping (TH), the second method is based on direct sequence code division multiple access (DS-CDMA) [2]. In both schemes binary information of a given user is conveyed by a signature waveform. In turn the signature waveform is a repetition of monocycles over a number of time-slots in the TH scheme, while it is a repetition of time delayed and weighted monocycles in the DS-CDMA scheme.

It has to be emphasized that power line channels are corrupted by several background disturbances [9]-[10] that include colored background noise, and impulse noise. Optimal detection, in the proposed impulse modulation based system, requires accurate synchronization, and

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channel estimation. In this paper we address such a problem. We assume to use a training bit sequence, and we devise the maximum likelihood estimation algorithm for synchronization and channel estimation. Low complexity solutions are also proposed.

In order to assess performance it is important to use appropriate channel, and noise models. Aiming at capturing the ensemble of network topologies we describe a simple statistical channel model. It is appropriate for describing both the forward, and the reverse link. In the forward link, a given receiver sees the superposition of the signals coming from the central transmitter. Even if signal multiplexing is done in a orthogonal fashion via orthogonal TH or DS-CDMA codes, multiple access interference is experienced by the receiver. This is due to the channel frequency selectivity generated by the reflections in the power grid [9]. In terms of noise model, we also consider the presence of impulse noise, and we model it with the common two-term Gaussian distribution [4].

We report performance results from simulations for several channel models, and noise scenarios. We study the sensitivity of the algorithms to the presence of impulse noise, and multiple access interference.

II. Impulse Modulation Transmission Model

In this section we describe the transmission model. It is based on impulse modulation combined with either time hopping (TH) or direct sequence code division multiple access (DS-CDMA). We note that herein we consider the forward link, so that the transmitter sends information to multiple receivers. We denote with $s_u(t)$ the information signal to be transmitted to user u. Thus, the composite signal sent over the power grid by the central transmitter is

$$s(t) = \sum_{u=0}^{N_U-1} s_u(t) \ . \qquad (2)$$

A. Time Hopped Impulse Modulation Scheme

In time hopped impulse modulation, transmission of a given information bit takes place over a number of frames (Fig. 3). A frame has duration T_f, and comprises N_S time slots of duration T followed by a guard time of duration T_g. We assume bi-phase pulse amplitude (BPAM) modulation such that the signal to be transmitted to user u can be written as

$$s_u(t) = \sum_k b_{u,k} \sum_{l=0}^{L-1} g(t - c_{u,l}^{TH} T - lT_f - kLT_f) \qquad (3)$$

where $b_{u,k} = \pm 1$ denotes the *k*-th information bit of user u, $g(t)$ is the pulse used to convey information (referred to as monocycle), and T_f is the bit period (frame duration). Further, $c_{u,l}^{TH}$ for $l = 0,....,L-1$ is the time hopping pattern

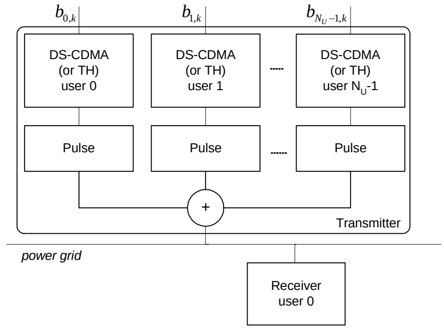

Fig. 2. Forward link (central transmitter to multiple receivers link) of impulse modulated PLC multiuser system.

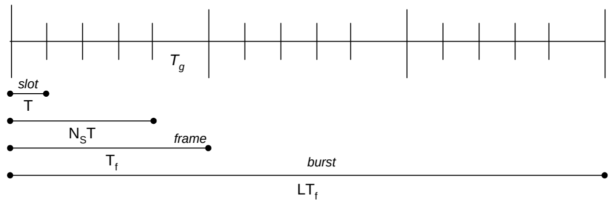

Fig. 3. Slot, frame, and burst format.

of user u (hopping codeword) that we assume to have period L. Each element $c_{u,l}^{TH}$ takes values in $\{0,....,N_S-1\}$. The monocycle is a short duration pulse that is designed to fulfill with the spectrum requirements. As an example, we assume the second derivative of the Gaussian pulse whose impulse response is given in (1). In typical system design we can choose $T \geq D$ where D is the monocycle pulse duration. Therefore, transmission comprises the time hopped repetition of each information bit over L frames. Equivalently, in TH pulse modulation, the binary information at rate $1/(LT_f)$ is conveyed by the user's signature waveform

$$v_u^{TH}(t) = \sum_{l=0}^{L-1} g(t - c_{u,l}^{TH} T - lT_f) \ . \qquad (4)$$

The hopping codewords can be appropriately designed to minimize the interference at the receiver side. For the forward link, we choose them to be orthogonal.

B. DS-CDMA Impulse Modulation Scheme

The DS-CDMA impulse modulated scheme that we consider herein, is designed to be consistent with the TH scheme such that they both have the same frame duration, and data rate. In our DS-CDMA scheme, a bit transmission spans a burst of L frames. Two spreading codes are assigned to a given user. A inner codeword $c_{u,m}^I$, $m = 0,....,N_S-1$, of length N_S is used to spread a bit within a frame. An outer codeword $c_{u,l}^O$, $l = 0,....,L-1$, of length L is used to spread the bit over L frames. Thus the signal to be transmitted to user u can be written as

$$s_u(t)=\sum_k b_{u,k}\sum_{l=0}^{L-1}c_{u,l}^{O}\sum_{m=0}^{N_S-1}c_{u,m}^{I}g(t-mT-lT_f-kLT_f)\,. \qquad (5)$$

Therefore, the binary information at rate $1/(LT_f)$ is conveyed by the user's signature waveform

$$v_u^{DS}(t)=\sum_{l=0}^{L-1}c_{u,l}^{O}\sum_{m=0}^{N_S-1}c_{u,m}^{I}g(t-mT-lT_f)\,. \qquad (6)$$

We choose the inner and outer codewords to be orthogonal. The outer codewords can be chosen to be identical to the inner codewords.

It may be argued that we could use a single codeword of length LN_S chips, and define a frame of duration $T_f=LN_ST+T_g$. Our choice, besides the consistency to the TH scheme, is dictated by the idea of allowing multirate transmission with a high degree of flexibility. In fact a high rate user can be assigned with the inner codeword only.

C. Example of System Parameters

With LN_S overall time slots (number of orthogonal codewords), in both the TH and the DS-CDMA system, the aggregate data rate equals $1/(T+T_g/N_S)$. Therefore, it approaches $1/D$ as the number of time slots increases. Preliminary measurements of in-home power lines of length less than 100 m, have shown that the channel maximum dispersion is below 0.5 μs. As an example, if we choose $T=D$, a monocycle with $T_0=D/5=30\ ns$ (-30 dB bandwidth of approximately 30 MHz), and a guard time of duration $T_g=0.5\ \mu s$, the aggregate data rate is 4.7 Mb/s with $N_S=8$, and can be increased by using longer codewords.

III. Forward Link Powerline Channel and Impulse Noise Model

In this section we describe the channel, and noise models that we use to evaluate the performance of the system. In particular we describe a statistical channel model that can be considered representative of the ensemble of network topologies.

A. Channel Impulse Response

Wide band power line channels are characterized by high frequency selectivity [9]. The channel frequency and impulse response are clearly a function of the system topology, and bandwidth. The frequency selectivity translates into time dispersion. We herein consider the forward link (central transmitter to multiple receivers connection). The reverse link (transmitters to receivers connection) is treated in [6]. We denote as $g^{CH}(t)$ the channel impulse response that is seen by the desired receiver. Aiming at capturing the ensemble of network topologies we assume the impulse response to exhibit a given number N_P of multipath components (echoes) so that it can be written as

$$g^{CH}(t)=\sum_{p=1}^{N_P}\alpha_p\delta(t-\tau_p) \qquad (7)$$

α_p denotes the sequence of real echo amplitudes, while τ_p denotes the associated delays. The channel echoes are due to the reflections that take place in the power grid. We assume the delays to be independent, and uniformly distributed in $[0,T_{CH}]$ where T_{CH} is the maximum channel time dispersion. In our model, the echo amplitudes are obtained by sampling an exponential profile, i.e.,

$$|\alpha_p|=Ae^{-\tau_p/\tau_D} \qquad (8)$$

where τ_D is a decaying factor. To incorporate the sign flip that is due to reflections, we weight each echo amplitudes by a random variable χ_p that takes on the values ±1 with equal probability, so that $\alpha_p=\chi_p|\alpha_p|$. The random variables χ_p are assumed to be independent.

The parameters of this model can be chosen to fit with measured channels.

B. Impulse Noise

We denote with $\eta(t)$ the background noise. The background noise is modeled with a two term Gaussian model as described in [4] to incorporate the presence of impulse noise. This is a good model for many natural impulsive noise sources, e.g., man-made impulsive noise. The probability density function can be obtained as follows

$$p_\eta(a)\to(1-\varepsilon)N(0,\sigma_1^2)+\varepsilon N(0,\sigma_2^2)\,. \qquad (9)$$

That is, the first term gives the zero mean Gaussian background noise term with variance σ_1^2. The second term represents the impulse component, and has variance $\sigma_2^2=100\sigma_1^2$. The probability of the impulse to occur is ε. In the simulations we assume $\varepsilon=0.05$. Further, when the impulse noise occurs we assume it to last for a given number of frames.

IV. Received Signal

The signal received by a given user, say, user 0, reads

$$y(t)=x(t)+\eta(t) \qquad (10)$$

where $x(t)$ is the equivalent signal that is obtained by the convolution of the composite transmitted signal with the channel impulse response, i.e.,

$$x(t)=s*g^{CH}(t-\Delta t)=\sum_{u=0}^{N_U-1}\sum_k b_{u,k}v_u^{EQ}(t-kLT_f-\Delta t) \qquad (11)$$

where $v_u^{EQ}(t)$ is the convolution of the user's signature waveform with the channel impulse response, while Δt denotes the time delay. We can write the equivalent signature waveform, respectively for the TH and DS-CDMA system as

$$v_u^{EQ,TH}(t)=\sum_{l=0}^{L-1} g^{EQ}(t-c_{u,l}^{TH}T-lT_f) \quad (12)$$

$$v_u^{EQ,DS}(t)=\sum_{l=0}^{L-1} c_{u,l}^{O}\sum_{m=0}^{N_S-1} c_{u,m}^{I} g^{EQ}(t-mT-lT_f) \quad (13)$$

with $g^{EQ}(t)=g*g^{CH}(t)$ being the convolution of the monocycle with the channel. If we assume to acquire perfect timing (burst synchronization), and we observe the received signal in correspondence to the k-th transmitted bit, we obtain

$$y_k(t)=y(t+kLT_f+\Delta t)=b_{0,k}v_0^{EQ}(t)+\sum_{u=1}^{N_U-1} b_{u,k}v_u^{EQ}(t)+\eta_k(t) \quad (14)$$

under the assumption of using a guard time longer than the maximum channel time dispersion, i.e., $T_g \ge T_{CH}$. Therefore, the k-th received burst is not affected by intersymbol interference (ISI). However, there may exist multiple access interference (MAI) as shown in the next section. The MAI term is a function of the multiplexing scheme that we are using, and the propagation conditions. In both the TH and the DS-CDMA scheme, MAI arises in the forward link because of the loss of the codeword orthogonality introduced by the channel time dispersion.

V. Single User Receiver

If the background noise is a stationary white Gaussian process, the optimal receiver is essentially a correlation receiver [6]. It passes the received signal through a front-end filter that is matched to the equivalent signature waveform of the desired user (user 0)

$$g_0^{MF}(t)=v_0^{EQ}(-t). \quad (15)$$

Then, samples are taken at rate $1/(LT_f)$ to obtain

$$Z_0(k)=\int_{-\infty}^{\infty} y(t)g_0^{MF}(kLT_f+\Delta t-t)dt. \quad (16)$$

Bit decisions are made by looking at the sign of (16),

$$\hat{b}_{0,k}=sign\{Z_0(k)\}. \quad (17)$$

We can rewrite (16) as follows

$$\begin{aligned} Z_0(k)&=b_{0,k}\int_0^{LT_f} |v_0^{EQ}(t)|^2\,dt \\ &+\sum_{u=1}^{N_U-1} b_{u,k}\int_0^{LT_f} v_u^{EQ}(t)v_0^{EQ}(t)dt+\int_0^{LT_f}\eta_k(t)v_0^{EQ}(t)dt \end{aligned} \quad (18)$$

Therefore, the matched filter output sample sees MAI whenever the cross-correlation of the equivalent signature waveforms is not zero. Even if we choose orthogonal signature waveforms, at the receiver side the cross correlations in (18) are not zero because of multipath propagation.

VI. Synchronization and Channel Estimation

To synthesize the matched filter in (15) we need to estimate the channel impulse response. We herein consider a data aided approach. That is, we assume that for each user a known training bit sequence is transmitted before sending the information bits. Alternatively a pilot channel associated with a given code can be used. Further, we assume a single user detection, and channel estimation approach, i.e., we neglect the presence of the other users signals. Focusing on the user of index 0, maximum likelihood channel estimation is based on finding the waveform $\hat{v}_0^{EQ}(t)$ that minimizes the likelihood function

$$\Delta(\hat{v}_0^{EQ})=\int_{-\infty}^{\infty} |\,y(t)-\sum_{k=1}^{N_T}\hat{b}_{0,k}\hat{v}_0^{EQ}(t-kLT_f-\widehat{\Delta t})|^2\,dt. \quad (19)$$

$\hat{v}_0^{EQ}(t)$ is an estimate of the equivalent signature waveform, and $\hat{b}_{0,k}$, are the N_T known training bits of user 0. To proceed we assume to synthesize such a waveform as follows

$$\hat{v}_0^{EQ,TH}(t)=\sum_{l=0}^{L-1}\sum_{p=1}^{N_P}\hat{\alpha}_p g(t-c_{0,l}^{TH}T-lT_f-\hat{\tau}_p) \quad (20)$$

$$\hat{v}_0^{EQ,DS}(t)=\sum_{l=0}^{L-1} c_{0,l}^{O}\sum_{m=0}^{N_S-1} c_{0,m}^{I}\sum_{p=1}^{N_P}\hat{\alpha}_p g(t-mT-lT_f-\hat{\tau}_p). \quad (21)$$

It follows that the waveform synthesis translates into the estimation of the channel amplitudes $\hat{\alpha}_p$, and the delays $\hat{\tau}_p$ that minimize (19). Indeed, the synthesis is optimal when the channel comprises a finite number of echoes as assumed in the model (7).

It can be shown that the estimation algorithm can be partitioned in a two-step procedure where first we estimate the delays, and then the amplitudes [3], [7], under the following assumption

$$\int_{-\infty}^{\infty} v_0(t-\tau_p)v_0(t-\tau_m)dt=0 \ \text{ for } \ \tau_p\ne\tau_m \quad (22)$$

where $v_0(t)$ is the signature waveform of user 0 that is defined in (4) for the TH scheme, and in (6) for the DS-CDMA scheme. In this case, we start by computing

$$\chi(\tau)=\sum_{k=1}^{N_T}\hat{b}_{0,k}\int_{-\infty}^{\infty} y(t)v_0(t-kLT_f-\tau_p-\Delta t)dt. \quad (23)$$

Then, we search the N_P delays $\boldsymbol{\tau}=[\tau_1,...,\tau_{N_P}]$ that jointly maximize the metric $\Lambda(\boldsymbol{\tau})$ as shown below:

$$\Lambda(\boldsymbol{\tau})=\sum_{p=1}^{N_P}\{\chi^2(\tau_p)\} \qquad \hat{\boldsymbol{\tau}}=\arg\max_{\boldsymbol{\tau}\in\Gamma}\{\Lambda(\boldsymbol{\tau})\} \quad (24)$$

where Γ is the set of vectors of N_P delays for which (22) holds true. Once we have determined the delays, the channel amplitudes are obtained as follows

$$\hat{\alpha}_p=\chi(\hat{\tau}_p)/E_v \quad (25)$$

where E_v is the energy of $v_0(t)$. Note that the burst timing Δt is included in the delays search. Finally, we note that (22) holds strictly true only in the TH scheme when the channel is separable, i.e., the minimum delay difference among the echoes is larger than the monocycle duration D. The joint maximization in (24) is particularly complex. To simplify the complexity we can just find the relative maxima of $\chi^2(\tau)$. In particular, in the numerical results that follow, we estimate the delays by finding the principle $\hat{N}_P$ maxima of $\chi^2(\tau)$. Further, we limit the search over echoes that have relative delay larger than *D*, or than *D*/2, i.e.,

$$\hat{\tau}_p = \underset{\substack{\tau,\, |\tau_p-\tau_{p-1}|>D \\ p=1,\ldots,\hat{N}_P}}{\arg\max}\{\chi^2(\tau)\} \quad (26) \qquad \hat{\tau}_p = \underset{\substack{\tau,\, |\tau_p-\tau_{p-1}|>D/2 \\ p=1,\ldots,\hat{N}_P}}{\arg\max}\{\chi^2(\tau)\}. \quad (27)$$

We refer to (26) as EST-D, while to (27) as EST/D-2.

A. Rake Receiver

The receiver can be implemented with a front-end stage where the received signal is filtered with a filter matched to the monocycle $g(t)$. Then, the output $Z_0(t)$ is sampled at sufficiently high rate to perform synchronization and channel estimation according to (23). After channel estimation, the signal $Z_0(t)$ is sampled at the appropriate time instants, and the samples are weighted by the tap amplitudes (and the spreading chips in the DS-CDMA case) and finally combined to generate $Z_0(k)$. This receiver implementation is referred to as *rake receiver*.

VII. Erasure Combining in Impulse Noise

The effect of the impulse noise is to raise the noise level for a certain time period (impulse duration). If we assume that the impulse spans only a fraction of a burst of duration LT_f, it is possible to improve the performance by combining at the detection stage only the frames (within a burst) that are not hit by the impulse noise. The resulting detector implements a form of erasure decoding, and it can be used with both TH and DS-CDMA.

Our algorithm works as follows. First, we run synchronization, and channel estimation. Then, we estimate the received average signal energy over each single frame within a burst. Finally, we disregard the bursts that have an estimated energy higher than a threshold. Only the remaining received frames are used to compute the correlation in (16). The threshold is chosen to be equal to twice the long term average received energy.

VIII. Numerical Results

Herein, we report several numerical results to characterize the performance of the impulse modulated system.

A. Single User System

In the following numerical results we assume a single user scenario with neither TH nor DS-CDMA. That is, we assume transmission of unspread w-shaped pulses followed by a guard time. The frame has duration $T_f = 5D$, where $D = 5T_0$ is the monocycle duration. The channel is according to the model (7), with $T_{CH} = 4D$, and $\tau_D = 0.6D$. Two models are considered. They differ on the delay distribution. Channel 1 (CH-1) assumes $N_P = 3$ separable taps with delays uniformly distributed in $[0, T_{CH}]$. Channel 2 (CH-2) assumes $N_P = 10$ non-separable taps with delays uniformly distributed in $[0, T_{CH}]$.

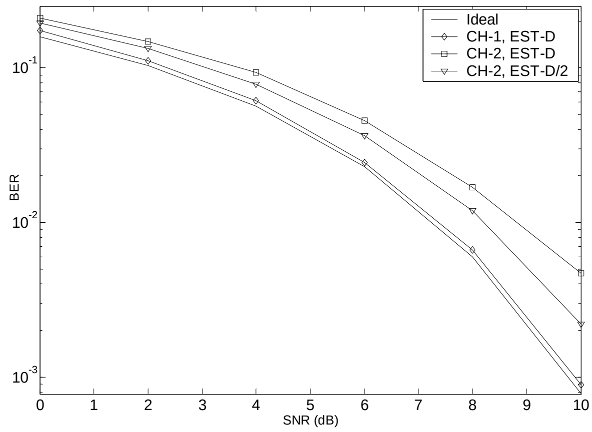

Fig. 4. BER in a single user system with separable and non separable channels, and with practical synchronization and channel estimation.

In Fig. 4, we report bit-error-rate performance assuming white Gaussian noise. The SNR is defined as the energy-per-bit over the noise variance at the matched filter output when all channel realizations are normalized such that $\int |v_0^{EQ}(t)|^2\, dt = E_b$. Curves labelled with *ideal* have been obtained with perfect channel knowledge. For the other curves we have performed practical synchronization and channel estimation by searching the relative maxima of $\chi^2(\tau)$ with the constraint of being separated by at least *D* (curves labelled with EST-D) or by at least *D*/2 (curves labelled with EST-D/2). The number of searched channel taps is limited to $\hat{N}_P = 3$ to lower the complexity of the receiver. The training sequence has length 100 bits. The figure shows that the EST-D algorithm achieves performance close to the ideal for the separable channel case. When the channel is not separable EST-D exhibits a performance penalty that can be partially recovered with the EST-D/2 algorithm.

B. Multiuser System

In the following numerical results we assume a multiuser system scenario using either TH, or DS-CDMA. The frame has duration $T_f = 12D$. The channel is according to the model (7), with $N_P = 10$, $T_{CH} = 4D$, and $\tau_D = 0.6D$. Ray delays are uniformly distributed in $[0, T_{CH}]$. Further, we allocate $N_S = 8$ time slots per frame each with duration $T = D$. Users have equal power. The hopping patterns and spreading codes are orthogonal. In all Fig. 6, 7, 8 the channel is estimated assuming a training sequence of length

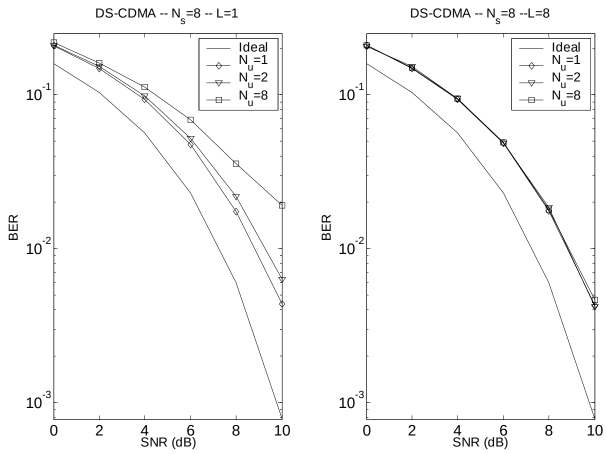

Fig. 5. BER in a multiuser system with DS-CDMA in white Gaussian noise, non separable channel and practical estimation (EST-D/2).

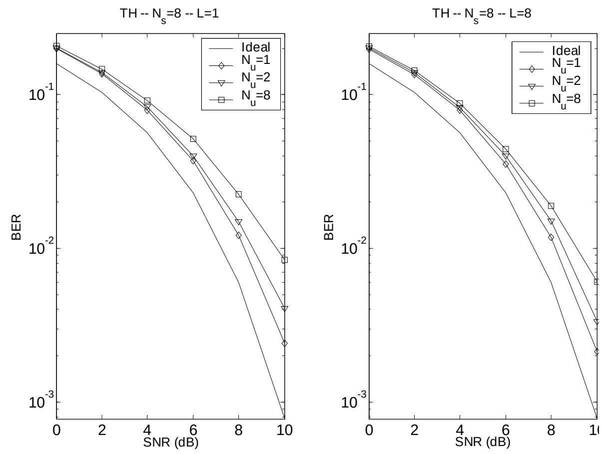

Fig. 6. BER in a multiuser system with TH in white Gaussian noise, with non separable channel, and practical estimation (EST-D/2).

100 and search of delays according to the EST-D/2 algorithm. The number of searched channel taps is limited to 3. Curves labelled with *ideal* assume a single user with perfect channel knowledge in white Gaussian noise.

In Fig. 5 we report performance for DS-CDMA, while in Fig. 6 for TH. The left plots assume $L=1$. That is, in DS-CDMA we use only the inner codeword of length 8. In the TH scheme we allocate up to 8 users in distinct time slots within a frame. The right plots assume codewords of length $L=8$. That is, in the DS-CDMA scheme we use both a inner codeword of length 8, and an outer codeword of length 8. In the TH scheme up to 8 users are allocated and use a hopping codeword of length 8.

The figures show that by increasing the codeword length it is possible to improve performance since less MAI is experienced. Further, the DS-CDMA scheme is more robust when a high number of users is present.

In Fig. 7 we assume also the presence of impulse noise. The duration of an impulse is assumed to be equal to 2 frames. However, the impulse is not frame synchronized. The results show that a sensible performance improvement is obtained with the proposed erasure combining algorithm.

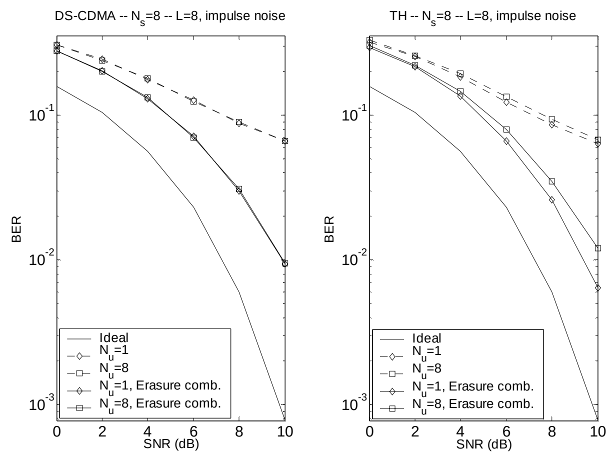

Fig. 7. BER in a multiuser system with DS-CDMA and TH in impulse noise, with non separable channel, practical estimation (EST-D/2), and erasure combining.

IX. Conclusions

We have described two transmission and multiplexing schemes for power line communications that are based on impulse modulation. Users multiplexing is done with time hopping or DS-CDMA. We have devised synchronization and channel estimation algorithms. We have also considered the presence of impulse noise, and we have described a simple detection algorithm based on an erasure decoding approach. Other detection and estimation algorithm approaches can be found in [5]-[6].

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