Detection Algorithms for Wide Band Impulse Modulation Based Systems 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. The performance is evaluated assuming simple detection algorithms that can take into account the noise and interference statistics.**

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

I. Introduction[1]

In this paper we consider the deployment of wide band impulse modulation for powerline multiuser communications. Both time hopping, and direct sequence code division multiple access schemes are described (Fig. 1). Simple detection algorithms are proposed and their performance is numerically evaluated. In particular, we focus on the reverse link, that is the connection between multiple asynchronous transmitters, and multiple receivers.

Up to date, impulse modulation has been proposed for application over wide band wireless channels only [8]. It shows interesting properties in terms of simple baseband implementation, and robustness against frequency selective fading and interference. It has been recognized that some similarity between wireless channels and power line channels exists. In particular the channel frequency selectivity translates into intersymbol interference. Based on this, several transmission approaches that are used in wireless may be of practical interest also for application over power line channels. Most of the current proposals consider the deployment of orthogonal frequency division multiplexing (OFDM) [1]. Instead, in this paper we consider the deployment of wide band impulse modulation. We specialize the detection algorithms taking into account the channel frequency selectivity and the presence of colored Gaussian noise [9]-[10].

The basic idea behind impulse modulation, is to convey information by mapping an information bit stream into a

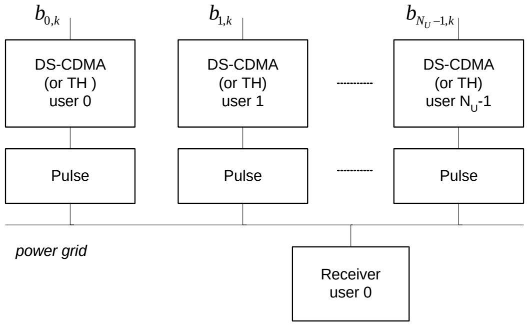

Fig. 1. Reverse link of impulse modulated PLC multiuser system.

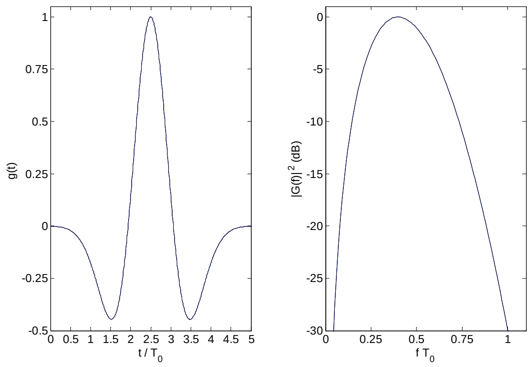

Fig. 2. W-shaped monocycle impulse and frequency response.

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. The monocycle can be appropriately designed to shape the spectrum occupied by the transmission system. As an example, in Fig. 2 we plot the impulse and frequency response of the second derivative of a Gaussian pulse,

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

where $D \approx 5T_0$ is the monocycle duration. An interesting property is that its spectrum does not occupy the low

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frequencies where we typically experience higher levels of man-made background noise.

In order to allow for users' multiplexing, and add robustness against interference, in this paper we consider to possible solutions. One is based on time-hopping (TH) and the other on direct sequence code division multiple access (DS-CDMA) [3]. We carry out a unified treatment of the two schemes. This is made possible by noting that in both schemes binary information of a given user is conveyed by a certain user's 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.

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 the reverse link (users-to-receiver) where a given receiver sees the superposition of the signals transmitted by several asynchronous users accessing the same powerline grid in distinct points. Further, we also consider the presence of colored background noise.

Several results from simulation are reported to compare the various schemes, and proposed receivers.

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).

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 modulation (BPAM) such that the signal transmitted by user u can be written as

$$s_u(t) = \sum_k b_{u,k} \sum_{l=0}^{L-1} g(t - c_{u,l}^{TH} T - lT_f - kLT_f) \qquad (2)$$

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

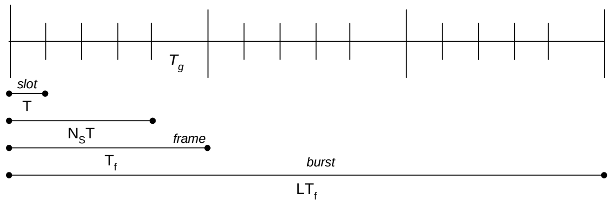

Fig. 3. Slot, frame, and block format.

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

$$v_u^{TH}(t) = \sum_{l=0}^{L-1} g(t - c_{u,l}^{TH} T - lT_f) \,. \qquad (3)$$

The hopping codewords can be appropriately designed to minimize the interference at the receiver side. In this paper, we assume them to be randomly generated.

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 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 inside 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 transmitted by user u can be written as

$$s_u(t) = \sum_k b_{u,k} \sum_{l=0}^{L-1} c_{u,l}^{O} \sum_{m=0}^{N_S-1} c_{u,m}^{I} g(t - mT - lT_f - kLT_f) \,. \qquad (4)$$

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

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

It may be argued that we could use a single codeword of length LN_S chips, and define a frame of duration $T_f = LN_S T + T_g$. 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 we can choose LN_S orthogonal codewords. For this case, in both the TH and the DS-CDMA system the aggregate data rate equals $1/(T+T_g/N_S)$. Therefore, it approaches $1/T$ 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. PLC Channel and Noise Model

In this section we describe the channel, and noise models that have been used for evaluating the performance of the system. In particular we describe a statistical channel model that can be considered representative of the channels 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. In the reverse link (users to receiver connection) the channels of distinct users are in general different. We denote as $g_u^{CH}(t)$ the channel impulse response of user u. Aiming at capturing the ensemble of network topologies we assume the impulse responses to exhibit a given number N_P of multipath components (echoes) such that it can be written as

$$g_u^{CH}(t) = \sum_{p=1}^{N_P} \alpha_{u,p}\delta(t-\tau_{u,p}) \tag{6}$$

$\alpha_{u,p}$ denotes the sequence of real echo amplitudes, while $\tau_{u,p}$ denotes the associated delays. 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.,

$$|\alpha_{u,p}| = Ae^{-\tau_{u,p}/\tau_D} \tag{7}$$

where the decaying factor τ_D can be in general different for distinct users. However, in the numerical results we assume it to be the same for all users. Further, to incorporate the sign flip that is due to reflections, we weight each echo amplitudes by a random variable $\chi_{u,p}$ that takes on the values ± 1 with equal probability, so that

$$\alpha_{u,p} = \chi_{u,p}|\alpha_{u,p}|. \tag{8}$$

The random variables $\chi_{u,p}$ are assumed to be independent.

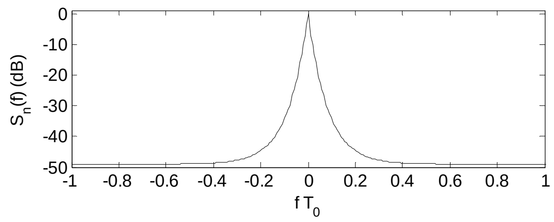

Fig. 4. Background noise PSD.

In the simulations the average channel energy has been normalized to one.

B. Colored Background Noise

The background noise $\eta(t)$ is assumed to be a stationary real Gaussian process with zero mean and correlation $R(\tau) = E[\eta(t)\eta(t+\tau)]$. If the noise is white, then $R(\tau) = N_0\delta(\tau)$. For the correlated noise, various spectral density models can be used [2], [10]. In this paper, we assume the noise power spectral density to be double sided exponential in the log-scale, i.e.,

$$S_\eta(f)_{dBm} = N_0 + N_1 e^{-|f|/F_0} \tag{9}$$

for given constants N_0, N_1, and F_0. It is consistent with the practical observation that the background noise is higher at low frequencies. In Fig. 4, we plot an example of its power spectral density.

C. Impulse Noise

Another noise source in powerline communications is the impulse noise [10]. In this paper, we do not consider it. However, the effect on performance of impulse noise, and possible ways to counteract it in an impulse modulated system can be found in [5].

IV. Received Signal

The received composite signal can be written as

$$y(t) = \sum_{u=0}^{N_U-1} x_u(t) + \eta(t) \tag{10}$$

where $x_u(t)$ is the equivalent signal that is obtained by the convolution of the *u*-th user transmitted signal with its channel impulse response, i.e.,

$$x_u(t) = s_u * g_u^{CH}(t-\Delta t_u) = \sum_k b_{u,k} v_u^{EQ}(t-kLT_f-\Delta t_u) \tag{11}$$

where $v_u^{EQ}(t)$ is the convolution of the user's signature waveform with the corresponding channel impulse response, while Δt_u denotes the time delay of user *u* with respect to the receiver timing reference. The equivalent user signature waveform, can be respectively written for the TH, and DS-CDMA system as

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

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

with $g_u^{EQ}(t) = g * g_u^{CH}(t)$ being the convolution of the monocycle with the channel of a given user. We note that the support of the equivalent signature waveform has duration LT_f assuming $T_g \geq T_{CH}$.

If we assume to acquire perfect timing with the desired user of index $u = 0$, and we observe the received signal in correspondence to the *k*-th transmitted bit, we obtain

$$y_k(t) = y(t + kLT_f + \Delta t_0) = b_{0,k} v_0^{EQ}(t) + MAI_k(t) + \eta_k(t) \quad (14)$$

under the assumption of using a guard time longer than the maximum channel time dispersion, i.e., $T_g \geq T_{CH}$. Therefore, the *k*-th received burst is not affected by intersymbol interference (ISI) for the desired user. However, there may exist multiple access interference (MAI). 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 whenever the users are asynchronous, or the users' codewords are not orthogonal, or the channel is time dispersive.

If we include the MAI in the noise additive term, we can write that the *k*-th received burst reads $y_k(t) = b_{0,k} v_0^{EQ}(t) + z_k(t)$ where with refer to $z_k(t)$ as the impairment signal.

V. Single User Receiver in Colored Gaussian Noise and Interference

In this paper we consider a single user based detection approach. However, the detection approach can include the capability of rejecting the MAI interference by the exploitation of its correlation. To this purpose, we assume to model the overall additive impairment term $z(t)$ that includes the MAI, and the background noise, with a zero mean colored Gaussian noise process with correlation

$$K(t;\tau) = R(\tau - t) + R_{MAI}(t;\tau) \,. \quad (15)$$

Due to the presence of the MAI, the impairment is not stationary, but ciclo-stationary with period LT_f. Recall, in fact, that we assume the users' signature waveforms to be periodic of LT_f. We herein consider three receivers structures. The first two receivers are basically correlation receivers. They pass the received signal through a front-end filter with an appropriate impulse response $g^{FE}(t)$. Then samples are taken at rate $1/(LT_f)$ to obtain

$$Z_0(k) = \int_{-\infty}^{\infty} y(t) g^{FE}(kLT_f + \Delta t_0 - t) dt \,. \quad (16)$$

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

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

The third receiver is a simple interference rejection detector.

A. Matched Filter Receiver

In a first receiver scheme that we refer to as *matched filter receiver* the front-end filter is a pulse matched to the equivalent signature waveform,

$$g^{FE}(t) = g_0^{MF}(t) = v_0^{EQ}(-t) \,. \quad (18)$$

This is the optimal receiver in stationary white Gaussian background noise. Recall that due to the deployment of the guard time the desired user does not experience ISI. Indeed, to synthesize the matched filter in (18) we need to estimate the channel impulse response. To achieve this goal we can first filter the received signal with a filter matched to the monocycle, sample the output at sufficiently high rate, and then perform channel estimation. Details can be found in [5]. If we assume the channel model in (6), channel estimation requires estimation of the echo delays and amplitudes for the desired user. Practical estimation is treated in [5]. If we assume to have such a knowledge, the equivalent signature waveform can be written as

$$v_u^{EQ,TH}(t) = \sum_{l=0}^{L-1} \sum_{p=1}^{N_P} \alpha_{u,p} g(t - c_{u,l}^{TH} T - lT_f - \tau_{u,p}) \quad (19)$$

$$v_u^{EQ,DS}(t) = \sum_{l=0}^{L-1} c_{u,l}^{O} \sum_{m=0}^{N_S-1} c_{u,m}^{I} \sum_{p=1}^{N_P} \alpha_{u,p} g(t - mT - lT_f - \tau_{u,p}) \,. \quad (20)$$

Thus, 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 the appropriate time instants, and the samples are combined after being weighted by the tap amplitudes (and the spreading chips in the DS-CDMA case). This receiver implementation is referred to as *rake receiver* [4]. To simplify its complexity it is possible to combine only a finite number of channel echoes that correspond to the highest energy taps.

In general, this receiver can be affected by the MAI. In fact we can rewrite (16) as follows

$$\begin{aligned} Z_0(k) &= b_{0,k} \int_0^{LT_f} | v_0^{EQ}(t) |^2 \, dt \\ &+ \sum_{u=1}^{N_U-1} \sum_{n=0}^{1} b_{u,k-n} \mathrm{X}(u,n) + \int_0^{LT_f} \eta_k(t) v_0^{EQ}(t) dt \end{aligned} \quad (21)$$

with $\mathrm{X}(u,n) = \int_0^{LT_f} v_u^{EQ}(t + nLT_f + \Delta t_0 - \Delta t_u) v_0^{EQ}(t) dt$.

Therefore, the matched filter output sample sees MAI whenever the cross-correlation $\mathrm{X}(u,n)$ of the equivalent signature waveforms is not zero. Even if we choose orthogonal signature waveforms, at the receiver side the cross correlations in (21) are not zero because of multipath propagation and users' time delays.

B. Noise-Matched Filter Receiver

If we assume the presence of colored stationary Gaussian noise, the previous receiver is not optimal. In order to maximize the receive filter output signal-to-noise ratio we should use a filter matched to the noise correlation, and channel [7]. That, is we can filter the received signal with

$$g^{FE}(t) = g_0^{NMF}(t) = \int_{-\infty}^{\infty} R^{-1}(\tau) v_0^{EQ}(\tau - t) d\tau \qquad (22)$$

where $R^{-1}(\tau)$ is the convolution inverse of the noise correlation function, i.e. $R * R^{-1}(\tau) = \delta(\tau)$. The inverse exists if the noise PSD is not zero over the signal bandwidth. This receiver is referred to as *noise-matched filter receiver*. Similarly to the previous case, the receiver can be implemented with a mixture of analog, and digital components.

C. MAI Rejection Detector

In the presence of multiple-access interference it is possible to include some interference rejection capability by using an interference cancellation detector that still operates on a burst by burst base. The detector works as follows. We first pass the received signal to a front-end stage that rejects out-of-band noise and interference. Then, we sample the output to obtain in correspondence of the *k*-th frame of the desired user

$$y_k(nT_c) = b_{0,k} v_0^{EQ}(nT_c) + z_k(nT_c) \qquad (23)$$

where $z_k(nT_c) = MAI_k(nT_c) + \eta_k(nT_c)$ is the impairment sample. Now, let M be the number of samples we collect per burst, so that we can define the following vectors

$$\mathbf{y}_k = [y_k(0),..., y_k((M-1)T_c)]^T \qquad (24)$$

$$\mathbf{v}_{0,k}^{EQ} = [v_{0,k}^{EQ}(0),..., v_{0,k}^{EQ}((M-1)T_c)]^T \qquad (25)$$

$$\mathbf{z}_k = [z_k(0),..., z_k((M-1)T_c)]^T . \qquad (26)$$

The impairment vector, noise plus MAI, has zero mean and correlation matrix $\mathbf{K} = E[\mathbf{z}\mathbf{z}^{\dagger}]$. If we assume the impairment to be Gaussian, the maximum likelihood receiver [4] decides in favor of the bit $\hat{b}_{0,k}$ of the desired user that minimizes the likelihood function

$$\Delta(\hat{b}_{0,k}) = (\mathbf{y}_k - \hat{b}_{0,k}\mathbf{v}_{0,k}^{EQ})^{\dagger} \mathbf{K}^{-1} (\mathbf{y}_k - \hat{b}_{0,k}\mathbf{v}_{0,k}^{EQ}) \qquad (27)$$

It follows that the bit decision is accomplished according to

$$\hat{b}_{0,k} = sign\left\{ \mathbf{v}_0^{EQ^{\dagger}} \mathbf{K}^{-1} \mathbf{y}_k \right\} \qquad (28)$$

The implementation requires to acquire synchronization with the desired user, to estimate its equivalent signature waveform, and finally, to estimate the impairment inverse correlation matrix. The equivalent signature waveform is known when the channel impulse response is estimated.

VI. Autocorrelation Receiver

A much simpler receiver that does not need any parameter estimation, is based on a autocorrelation approach. We describe it for the TH case only. Further, we assume differential encoding of the transmitted bits, i.e., the information bits $b_{u,k}$ are mapped into $d_{u,k} = b_k d_{u,k-1}$. The correlation receiver first filters the received signal with $g(-t)$ to reject out of band noise and interference, to obtain the signal $r(t)$. Then, assuming frame synchronization with the desired user, the signals corresponding to two adjacent frames are correlated to generate an estimate of the matched filter output as follows:

$$\hat{Z}_0(kT_f) = \sum_{l=0}^{L-1} \int_{lT_f + c_{0,l}T}^{lT_f + c_{0,l}T + T + T_1} r_k(t) r_{k-1}(t) dt \qquad (29)$$

where $0 \le T_1 \le D + T_{CH}$ can be optimally chosen to maximize performance. Finally, bit decisions are made by looking at the sign of (29). The drawback with such a simple approach is essentially a noise enhancement (see next section).

VII. Numerical Results

In this section we report several numerical results to characterize the performance of the impulse modulated systems, and receivers that we have described. The channel impulse response is assumed to be known.

A. Single User System

In the following numerical results we consider 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 $T_f = 4D$, where $D = 5T_0$ is the monocycle duration. The channel is according to the model (6), with $N_P = 10$, $T_{CH} = 3D$, and $\tau_D = 0.432D$.

We evaluate the performance in both white and colored stationary Gaussian noise. In particular we plot the bit-error-rate performance for the matched filter receiver, and the noise-matched filter receiver. For the correlated noise case, we use the model in (9) with $N_1 T_0 = 2.668\text{e-}6$ dBm, $F_0 T_0 = 0.0828$ (see also Fig. 4). Further, we define the average energy per bit (E_b/N_0) at the output of the front-end filter that is matched to the monocycle. The average channel energy is set to one.

Fig. 5 shows that some performance advantage is achievable when we take into account the noise correlation (assumed known). The autocorrelation receiver is characterized by a severe performance loss.

B. Multiuser System

In the following numerical results we assume a multiuser system scenario using both TH and DS-CDMA. The frame has duration $T_f = 10D$. The channel is according to the model (6), with $N_P = 5$, $T_{CH} = 2D$, and $\tau_D = 0.284D$. We allocate $N_S = 8$ time slots per frame each with duration $T = D$. The hopping patterns and spreading codes are randomly chosen.

In Fig. 6 we assume the noise to be white, but the users are asynchronous, and have a uniformly distributed delay in [0, LT_f] with respect to the desired user. We use the simple matched filter receiver (rake receiver). As the curves show there is a performance degradation as the number of users increases that is more pronounced for the DS with no outer spreading compared to the TH scheme with L=1. However, when we use the outer spreading, DS-CDMA performs better than TH with L=8.

Finally, in Fig. 7 we assume a DS-CDMA system without outer spreading (only inner codeword of length 8). We compare the performance of the matched filter receiver with the performance of the MAI canceling algorithm of Section V.c. As the curves show there is a sensible performance improvement with the latter scheme. We point out that in this figure the MAI interference correlation matrix has been estimated using a 500 bits known training sequence. More details can be found in [6].

VIII. Conclusions

We have described two transmission and multiplexing schemes for power line communications that are based on impulse modulation. They comprise either time hopping or DS-CDMA. We have described simple receivers that have the capability of exploiting both the background noise and the multiple access interference correlation. The channel estimation problem and the effect of impulse noise are addressed in [5]. Other detection and estimation algorithm approaches can be found in [6].

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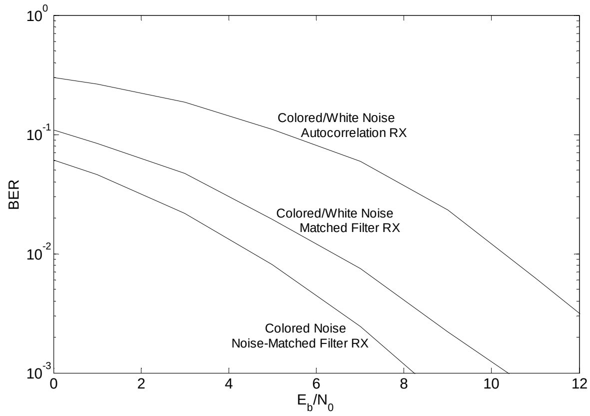

Fig. 5. BER in a single user system in the presence of white and colored noise and for different receivers.

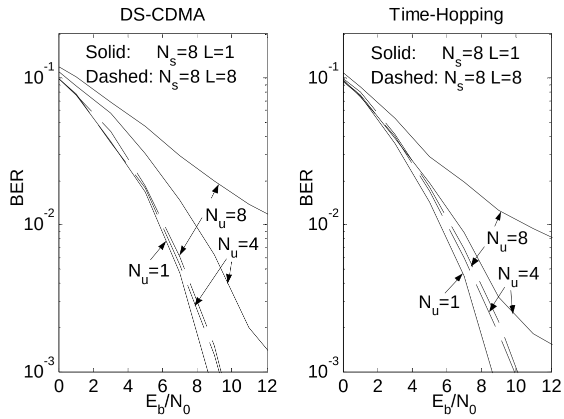

Fig. 6. BER in an asynchronous multiuser system with DS-CDMA or TH, and equal power users, assuming a single user receiver.

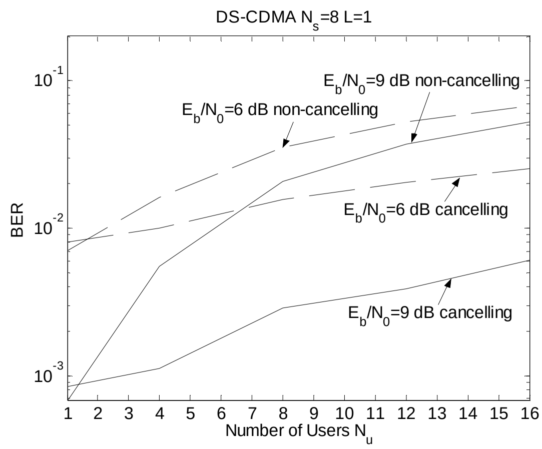

Fig. 7. BER in an asynchronous multiuser system with DS-CDMA and equal power users, with and without MAI cancellation.