An Impulse Modulation Based PLC System with Frequency Domain Receiver Processing

Andrea M. Tonello

DIEGM - Università di Udine
Via delle Scienze 208 - 33100 Udine – Italy - e-mail: tonello@uniud.it

***Abstract*—In this paper we propose a wideband impulse modulated based system for indoor power line communications (PLC). Impulse modulation is combined with DS-CDMA to allow for user multiplexing. We describe the key system parameters and we focus on the receiver algorithms. The receiver operates in the frequency domain (FD). We describe various FD algorithms that include the capability of canceling the inter-code interference and the multiple access interference. Simulation results are reported for a proposed wide band statistical channel model.**

***Keywords*—Impulse modulation, frequency domain processing.**

I. INTRODUCTION

In this paper we propose the deployment of wide band impulse modulation for powerline (PL) multiuser communications (Fig.1). In particular we consider indoor applications such as local area networks (LAN), peripheral office connectivity, and industrial automation. Up to date, impulse modulation has been mostly considered for application over wide band wireless channels [1], [2]. 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. 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) [3]. Instead, in this paper we consider the use of wide band (beyond 20 MHz) impulse modulation. The basic idea behind impulse modulation is to convey information by mapping an information symbol stream into a sequence of short duration pulses. No carrier modulation is required. Pulses (referred to as monocycles) are followed by a guard time in order to cope with the channel time dispersion. The monocycle can be appropriately designed to shape the spectrum occupied by the transmission system and in particular avoid the low 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 the use of DS-CDMA [4]-[6]. Information of a given user is conveyed by a certain user's signature waveform (Fig.1). The waveform is a repetition of time delayed and weighted monocycles that spans a transmission frame. Frames are separated by a guard time to cope with the channel time dispersion.

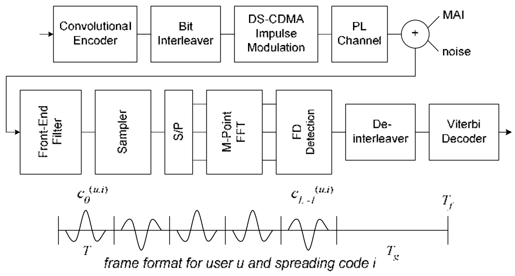

Fig. 1. Impulse modulated DS-CDMA system.

The simplest receiver is the matched filter receiver that correlates the received signal with a template waveform. The template waveform has to be matched to the equivalent impulse response that comprises the user's waveform, and the channel impulse response. Accurate estimation of the channel is necessary, and such an estimation can be complex if performed in the time domain since the signal occupies a wide bandwidth and the channel is highly frequency selective [7]. Simplified, although sub-optimal, time domain channel estimation approaches were presented in [4]. Further, the channel frequency selectivity introduces inter-code interference (ICI) (interference among the codes that are assigned to the same user) and multiple access interference (MAI) when multiple users access the network. This translates into performance losses, such that some form of multiuser detection or interference cancellation is advisable. Motivated by the above considerations, we consider a novel frequency domain (FD) detection approach. Several different (FD) algorithms are described. They all includes the capability of rejecting the ICI/MAI and they take into account the presence of colored background noise. The approach comprises the following stages (Fig.1). First we acquire frame synchronization with the desired user. Second, we run a discrete Fourier transform (DFT) on the received frames. Then, we perform FD detection. Channel coding is also considered and it is based on bit-interleaved convolutional codes.

II. WIDE BAND IMPULSE MODULATED SYSTEM MODEL

We consider a system where 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. Each node deploys wide band impulse

Proceedings of the 9th International Symposium on Power-Line Communications and its Applications (*ISPLC 2005*)
April 6-8, 2005, Vancouver, Canada
ISBN 0-7803-8844-5 © IEEE

modulation combined with DS data spreading as described in what follows. Nodes multiplexing is done in a CDMA fashion as described in Section II.A.

In our PLC system the transmission scheme uses multilevel pulse amplitude (PAM) modulation combined with direct sequence (DS) data spreading. Users multiplexing is obtained allocating different spreading codes among the users. The signal transmitted by user *u* can be written as

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

where $g^{(u,i)}(t)$ is the waveform (signature code) used to convey information for the user *u* and the *i*-th information symbol $b_k^{(u,i)}$ that is transmitted during the *k*-th frame. Each symbol belongs to the PAM alphabet and carries $\log_2 M_S$ bits where M_S is the number of PAM levels. T_f is the symbol period (frame duration). Note that the user *u* transmits C_u information symbols per frame where C_u denotes the set of signature code indices that are allocated to user *u*. The signature code comprises the weighted repetition of $L \geq 1$ narrow pulses (monocycles), i.e.,

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

where $c_m^{(u,i)} = \pm 1$ are the codeword elements (chips), and T is the chip period. We can choose the codewords to be either orthogonal, e.g., Walsh-Hadamard, or random (pseudo-noise). 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, $g_M(t) \sim \exp(-\pi/2((t - D/2)/T_0)^2)$, where $D \approx 5T_0$ is the monocycle duration (Fig.2). An interesting property is that its spectrum does not occupy the low frequencies where we experience higher levels of man-made background noise.

In typical system design we can 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. The frame duration has, therefore, duration $T_f = LT + T_g$.

A. User Multiplexing

In order to multiplex the nodes we can follow either a TDMA approach or a CDMA approach. In this paper we consider the latter approach, and in particular we allocate distinct codes to distinct users. In our design the codes are defined as follows:

$$c_m^{(u,i)} = c_m^{(u)} c_m^{(i)} \quad m = 0, ..., L-1 \tag{3}$$

where $\{c_m^{(u)}\}$ is a binary (± 1) random sequence of length L allocated to user *u*, while $\{c_m^{(i)}\}$ is a 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 $R = L/T_f = 1/(T + T_g/L)$ symb/s. It approaches $\log_2 M_S/T$ bit/s with long codes. Clearly, signals belonging to distinct transmitting nodes are not orthogonal. The random code $\{c_m^{(u)}\}$ is used to randomize the effect of the multiple access interference.

B. Received Signal

In the multiple access channel that we consider, the signals that are transmitted by distinct nodes (users) propagate through distinct channels with impulse response $h^{(u)}(t)$. At the receiver side, we deploy a band-pass front-end filter with impulse response $g_{FE}(t) = g_M(-t)$ that is matched to the transmit monocycle and that suppresses out of band noise and interference. Then, the composite received signal in the presence of N_I other users (interferers), reads

$$y(t) = \sum_k \sum_{i \in C_0} b_k^{(0,i)} g_{EQ}^{(0,i)}(t - kT_f) + i(t) + \eta(t) \tag{4}$$

$$i(t) = \sum_k \sum_{u=1}^{N_I} \sum_{i \in C_u} b_k^{(u,i)} g_{EQ}^{(u,i)}(t - kT_f - \Delta_u)$$

where the equivalent impulse response of user *u* and symbol *i* is denoted as $g_{EQ}^{(u,i)}(t) = g^{(u,i)} * h^{(u)} * g_{FE}(t)$. The index $u = 0$ denotes the desired user. Δ_u denotes the time delay of user *u* with respect to the desired user's frame timing. $\eta(t)$ denotes the additive noise. The equivalent impulse response comprises the convolution of the signature code of index (*u*,*i*) with the channel impulse response of the corresponding user, and the front-end filter. Distinct users experience distinct channels that we assume to introduce identical maximum time dispersion. The channel impulse response is assumed to be time-invariant over several transmitted frames.

III. Channel Model

A. Statistical Channel Impulse Response

We propose to use a statistical channel model to evaluate performance. We start from the band pass channel model in [7] where the frequency response is modelled as

$$H_+(f) = \sum_{p=1}^{N_P} g_p e^{-j\frac{2\pi d_p}{v} f} e^{-(\alpha_0 + \alpha_1 f^K) d_p} \quad 0 \leq B_1 \leq f \leq B_2$$

where N_P is the number of multipath (echoes), $|g_p| \leq 1$ is the transmission/reflection factor, d_p is the length of the path, $v = c/\sqrt{\varepsilon_r}$ with c speed of light and ε_r dielectric constant. The parameters α_0, α_1, K are chosen to adapt the model to a specific network. This, model can realistically represent a true frequency response. A limit of this model is that it is not much useful to assess the average performance of a PLC system and of the related algorithms. In other words, as it is common practice in the wireless context, it is beneficial to deal with a statistical model that allows to capture the ensemble of network topologies. Motivated by this, we propose to add some statistical property to it. In particular, we assume the reflectors (that generate the paths) to be placed over a finite distance interval. We fix the first reflector at distance d_1, and we assume the other reflectors to be located according to a Poisson arrival process with intensity $\Lambda\ [m^{-1}]$. The reflection factors g_p are assumed to be real, independent and uniformly distributed in $[-1,1]$. Finally, we appropriately choose α_0, α_1, K to a fixed value. If we further assume K=1, the real impulse response can be obtained in closed form. This allows to easily generate a real channel realization (corresponding to a realization of the random parameters N_P, g_p, d_p) as follows

$$h^{(u)}(t) = 2\,\mathrm{Re}\left\{ \sum_{p=1}^{N_P} g_p e^{-\alpha_0 d_p} \frac{\alpha_1 d_p + j2\pi(t - d_p/v)}{(\alpha_1 d_p)^2 + 4\pi^2 (t - d_p/v)^2} \times \left(e^{B_1(j2\pi(t-d_p/v)-\alpha_1 d_p)} - e^{B_2(j2\pi(t-d_p/v)-\alpha_1 d_p)}\right) \right\}. \quad (5)$$

We assume distinct users to experience independent channels.

B. Background Noise

We assume the background noise to be in general a colored Gaussian process. Furthermore, in the simulation to test the effect of impulse noise we use the two terms Gaussian model [4] whose probability density function can be defined as $p_\eta(a) \to (1-\varepsilon)N(0,\sigma_1^2) + \varepsilon N(0,\sigma_2^2)$. The first term gives the zero mean Gaussian background noise with variance σ_1^2. The second term represents the impulsive component and it has variance $\sigma_2^2 = 100\sigma_1^2$. The occurrence probability is ε. The impulse noise lasts for a given time interval. The spectrum of this noise model can be shaped to increase its low frequency components. However, if we do not do so we get a worst case scenario especially in our system where the transmission spectrum has significant energy at the high frequencies.

IV. CORRELATION RECEIVER

The matched filter receiver operates in a symbol by symbol fashion by computing the correlation between the received signal frame $y_k(t) = y(t + kT_f)$, $0 \le t < T_f$, and the real equivalent impulse response $g_{EQ}^{(0,i)}(t)$ to obtain the decision metric $z_{DM}^{(0,i)}(kT_f) = \int_0^{T_f} y_k(t) g_{EQ}^{(0,i)}(t)dt$, for the symbol i of user 0. Then, a threshold decision is made to detect the k-th transmitted symbol. For instance with 2-PAM, $\hat{b}_k^{(0,i)} = sign\left\{z_{DM}^{(0,i)}(kT_f)\right\}$. To implement the correlation receiver we need to estimate the channel. Time-domain channel estimation is complicated by the large time dispersion of the PL channel that implies that $g_{EQ}^{(0,i)}(t)$ is an involved function of the channel and the transmitted waveform. Furthermore, the correlation receiver suffers of the presence of inter-code interference and multiple access interference.

V. FREQUENCY DOMAIN RECEIVER

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

$$Y_k(f_n) = \sum_{i \in C_0} b_k^{(0,i)} G_{EQ}^{(0,i)}(f_n) + I_k(f_n) + N_k(f_n) \quad (6)$$

where $Y_k(f_n)$, $G_{EQ}^{(0,i)}(f_n)$, $I_k(f_n)$, $N_k(f_n)$, for $f_n = n/(MT_c)$, n=0,…,M-1, are respectively the DFT outputs of: the received frame samples, the desired user-symbol equivalent impulse response $g_{EQ}^{(0,i)}(nT_c)$, the interference, and the noise samples. No inter-frame interference is present for the desired user assuming perfect frame timing, and a sufficiently long guard time. The MAI term is a function of the users' time delay, transmitted waveform, and channel. To derive the proposed receiver, we model the noise-plus-interference $z(kMT_c + nT_c) = z_k(nT_c) = i_k(nT_c) + \eta_k(nT_c)$ with a discrete time colored Gaussian process (not necessarily stationary) with correlation $r(nT_c, lT_c) = E[z(nT_c)z(lT_c)]$. Then, assuming the transmitted symbols of all users to be i.i.d. and equally likely, the DFT outputs $Z_k(f_n) = I_k(f_n) + N_k(f_n)$ are complex Gaussian with zero mean. The impairment multivariate process $\mathbf{Z}_k$ defined as $\mathbf{Z}_k = [Z_k(f_0),...,Z_k(f_{M-1})]^T$, has *time-frequency* correlation matrix equal to

$$\mathbf{R}(k,m) = E[\mathbf{Z}_k \mathbf{Z}_m^\dagger] = \mathbf{F}\mathbf{K}(k,m)\mathbf{F}^\dagger \quad (7)$$

where $\mathbf{K}(k,m)$ is the $M \times M$ matrix with entries $r(kM+n, mM+l)$ for $n,l = 0,...,M-1$, and $\mathbf{F}$ is the M-point DFT orthonormal matrix.

Now, let us collect the elements of $Y_k(f_n)$, and of $G_{EQ}^{(0,i)}(f_n)$ in the vectors $\mathbf{Y}_k$, and $\mathbf{G}_{EQ}^{(u,i)}$. Then, it can be shown by using Parseval theorem, that the maximum-likelihood receiver can be implemented in the frequency domain by searching the sequence of transmitted bits $\{b_k^{(0,i)}\}$ (belonging to the desired user) that maximizes the log-likelihood function[1]

$$\Lambda(\{b_k^{(0,i)}\}) = -\sum_{k=-\infty}^{\infty} \sum_{m=-\infty}^{\infty} [\mathbf{Y}_k - \sum_{i \in C_0} b_k^{(0,i)} \mathbf{G}_{EQ}^{(0,i)}]^\dagger \times \mathbf{R}^{-1}(k,m)[\mathbf{Y}_m - \sum_{n \in C_0} b_m^{(0,n)} \mathbf{G}_{EQ}^{(0,n)}]. \quad (8)$$

Detection is jointly performed for the desired user's symbols.

[1] Let $\mathbf{R}$ be the matrix whose $M \times M$ block of indices (k,m) is $\mathbf{R}(k,m)$, and let $\mathbf{R}^{-1}$ be its inverse. Then, $\mathbf{R}^{-1}(k,m)$ denotes the $M \times M$ block of indices (k,m) of $\mathbf{R}^{-1}$. If $\mathbf{R}$ is block diagonal, e.g., when we neglect the impairment correlation across frames, $\mathbf{R}^{-1}(k,k) = (\mathbf{R}(k,k))^{-1}$.

Note that in (8) all signals belonging to the other nodes are treated as interference whose frequency domain correlation (once estimated) is included in the matrix $\mathbf{R}(k,m)$ together with the correlation of the background noise.

A. Simplified FD Joint Detector

In order to simplify the algorithm complexity we neglect the temporal correlation of the interference (MAI+noise) vector $\mathbf{Z}_k$, i.e., we assume $\mathbf{R}(k,m)=0$ for $k \neq m$. Indeed, the MAI correlation across frames is zero only for the synchronous users case. Then, by dropping the terms that do not depend on the information symbols $\mathbf{b}_k^{(0)} = [b_k^{(0,i)}, i \in C_0]$ of the desired user, the log-likelihood function simplifies to

$$\Lambda(\mathbf{b}_k^{(0)}) \sim \mathrm{Re}\left\{ \sum_{i \in C_0} b_k^{(0,i)} \mathbf{G}_{EQ}^{(0,i)\dagger} \mathbf{R}^{-1}(k,k) \left[\mathbf{Y}_k - \frac{1}{2} \sum_{n \in C_0} b_k^{(0,n)} \mathbf{G}_{EQ}^{(0,n)} \right] \right\}. \tag{9}$$

We make a decision on the transmitted symbols of frame k as $\hat{\mathbf{b}}_k^{(0)} = \arg\max_{\mathbf{b}_k^{(0)}} \left\{ \Lambda(\mathbf{b}_k^{(0)}) \right\}$. Therefore, according to (9) the frequency domain receiver operates on a frame by frame basis, and it exploits the frequency correlation of the MAI+noise. Further note that detection is performed jointly for all symbols that are simultaneously transmitted in a frame by the desired node. We assume the correlation matrix to be full rank, otherwise pseudo-inverse techniques can be used.

To obtain (9) we need to estimate $\mathbf{G}_{EQ}^{(0,i)}$. The attractive feature with this approach is that the matched filter frequency response at a given frequency depends only on the channel response at that frequency. This greatly simplifies the channel estimation task [6]. By exploiting the Hermitian symmetry of $\mathbf{G}_{EQ}^{(0,i)}$, the estimation can be carried out only over $M/2$ frequency bins. A further simplification is obtained by observing that the Fourier transform of the equivalent channel of the desired user has significant energy only over a small fraction of frequency bins, and only here channel estimation has to be performed. Furthermore, no restrictive assumption about the channel impulse response has been made.

B. Iterative FD Joint Detector

The complexity of the *Simplified FD Joint Detector* is still high because it increases exponentially with the number of symbols that are transmitted by the desired user in a frame. A possible way to simplify complexity, is to search the maximum of the metric in an iterative fashion. That is, we first detect symbol $\hat{b}_k^{(0,0)}$ by setting to zero all other bits in $\Lambda(\mathbf{b}_k^{(0)})$. Then, we detect symbol $\hat{b}_k^{(0,1)}$ by setting $b_k^{(0,0)} = \hat{b}_k^{(0,0)}$ in $\Lambda(\mathbf{b}_k^{(0)})$. We detect new symbols exploiting past decisions. Once all symbols are detected, we can re-run an iterative detection pass. This algorithm is similar to successive interference cancellation but it operates in the frequency domain.

C. FD Full Decorrelator

Another possibility is to perform detection of the symbols that belong to the desired node in a symbol by symbol fashion. That is, when we detect one symbol we treat as interference both the other users signals and all signals that are associated to all other codes that belong to the desired user. Thus, the decision metric for the i-th symbol of user 0 becomes

$$\Lambda(b_k^{(0,i)}) \sim \mathrm{Re}\left\{ b_k^{(0,i)} \mathbf{G}_{EQ}^{(0,i)\dagger} \mathbf{R}_{(0,i)}^{-1}(k,k) \left[\mathbf{Y}_k - \frac{1}{2} b_k^{(0,i)} \mathbf{G}_{EQ}^{(0,i)} \right] \right\} \tag{10}$$

where $\mathbf{R}_{(0,i)}^{-1}(k,k)$ is the inverse of the correlation matrix of the ICI+MAI that is seen by symbol i of frame k.

VI. Frequency Domain Parameter Estimation

The practical implementation of the above algorithms requires to first acquire frame synchronization. A training based procedure is described in [6]. Furthermore, we need to estimate the frequency response of the desired user channel, and the interference correlation matrix. For space limitations we do not describe the estimation algorithms. The interested reader is referred to [6] for some details and possible approaches based on training that are applicable also in this context. Training can be done deploying a pilot channel (a Walsh code). Alternatively, we can periodically send a number of frames that carry only training information. We can use a simple one tap (one per frequency bin) RLS algorithm to obtain the desired user channel estimates. The correlation matrix of the interference can be estimated with a least squares approach.

VII. Channel Coding

In this paper we consider the use of bit interleaved convolutional codes. A block of information bits is coded, interleaved, and then modulated as described in Section II. Interleaving spans a packet of N frames.

VIII. Performance Results

We assume a system whose parameters are reported in the following. The sampling period at the receiver filter output is T_c while the frame duration is $T_f = MT_c$, with M=512 (FFT size). Assuming T_c=16 ns, the sampling frequency is F_c=62.5 MHz, the frame duration is T_f=8.192 µs, and the monocycle duration is $D \approx 126$ ns, and T=128 ns. The Walsh codes have length 32. The guard time is T_g=4.096 µs. 2-PAM is assumed and 1 Walsh code is reserved for training. A bit interleaved convolutional code of rate ½ and memory 4 is used. A coded packet has length 124x31=3844 bits, such that the block interleaver spans 124 frames. It follows that the raw transmission rate equals 3.9 Mbit/s, while the net rate with coding is 1.89 Mbit/s per user. It can be increased with higher level PAM or longer orthogonal spreading codes.

We use the statistical channel model of Section III.A with B_1=0 and B_2=55 MHz. Further, the underlying Poisson process has intensity Λ=1/15 m^{-1}, i.e., 1 reflector every 15 m in average, with the first one set at distance 30 m. The maximum path distance is set at 300 m, and we fix K=1, α_0=1e-5 m^{-1}, α_1=1e-9 s/m. The background noise is assumed either white Gaussian or impulsive. In Fig.2 we plot the monocycle impulse and frequency response with the above parameters. Further, we show a channel realization and the equivalent channel response $g_{EQ}(t)=g_M*h*g_{FE}(t)$. Note that the latter is significantly compressed.

We report BER performance averaged over at least 2000 channel realizations. In Fig.3 we assume a single user that deploys all Walsh codes, and we assume knowledge of the channel responses. Looking at Fig.3.A, AWGN case with the channel model described before, the worst performance is obtained with the correlation receiver. Since the front-end receiver colors the noise, the FD receiver (FD-MF) that takes it into account, improves performance. However, the severely dispersive channel introduces inter-code interference, thus an error floor is visible. If we use the FD full decorrelator receiver we get a significant performance gain. Here, to simplify complexity we actually combine only the sufficiently high energy frequency bins. Near ideal performance is achieved with FD Joint Iterative Detection (FD-JD) with just 2 iterations. With channel coding the performance is greatly improved also in the presence of impulsive noise (Fig.3.B). Here we assume the impulse noise to occur with probability ε=0.01 and to have duration $4T_f$. The results show that a degradation occurs. However, it can be recovered by detecting the frames (symbols) that have received energy higher than a given threshold. Then, the branch metric in the Viterbi algorithm is set to zero in correspondence to such symbols. This is a form of combined soft/erasure decoding.

In Fig.4 we assume the presence of 2 or 4 interferers. They deploy all 32 Walsh codes but have also a random code on top of that as described in Section II.B, i.e., they transmit at full rate. Note that this corresponds to a significantly overloaded system. The overall MAI power equals the desired user power. The channels are independently drawn according the statistical model with AWG background noise. Users are asynchronous with a random starting phase. Fig.4 shows that although there is some performance penalty compared to Fig.3.A. due to the MAI, the iterative cancellation algorithm allows to keep such a penalty very small.

IX. CONCLUSIONS

We have proposed a novel wide band impulse modulated approach for PLC communications. We have described various frequency domain receiver algorithms.

REFERENCES

[1] M. Win, R. Scholtz, "Impulse radio: how it works", *IEEE Comm. Letters*, vol. 2, pp. 10-12, January 1998.

[2] G. Durisi, S. Benedetto, "Performance evaluation and comparison of different modulation schemes for UWB multiple access", *Proc. of IEEE ICC 2003*, Anchourage, USA.

[3] S. Baig, N. Gohar, "A discrete multitone transceiver at the heart of the PHY layer of an in-home PLC local are network", *IEEE Comm. Magazine*, pp. 48-53, April 2003.

[4] A. Tonello, R. Rinaldo, M. Bellin, "Synchronization and channel estimation for wide band impulse modulation over power line channels", *Proc. of ISPLC 2004*, Zaragoza, Spain, pp. 206-211, March 2004.

[5] A. Tonello, R. Rinaldo, L. Scarel, "Detection algorithms for wide band impulse modulation over power line channels", *Proc. of ISPLC 2004*, Zaragoza, Spain, pp. 367-372, March 2004.

[6] A. Tonello, R. Rinaldo, "A time-frequency domain approach to synchronization, channel estimation, and detection for DS-CDMA impulse radio systems", *IEEE Trans. on Wireless Comm.* 2005, in press.

[7] M. Zimmermann, K. Dostert, "A multipath model for the powerline channel", *IEEE Trans. on Comm.*, vol. 50, pp. 553-559, April 2002.

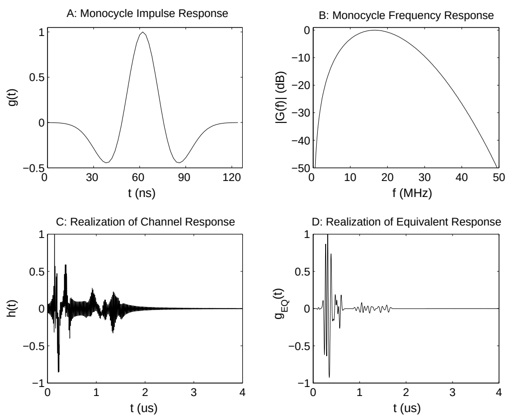

Fig. 2. Monocycle impulse/frequency response, and example of channel realization and equivalent channel realization.

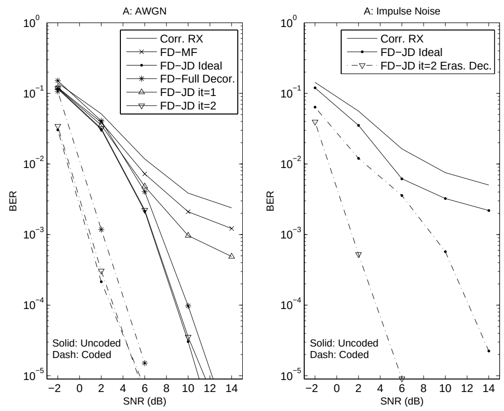

Fig. 3. Average bit-error rate for the single user case at full rate.

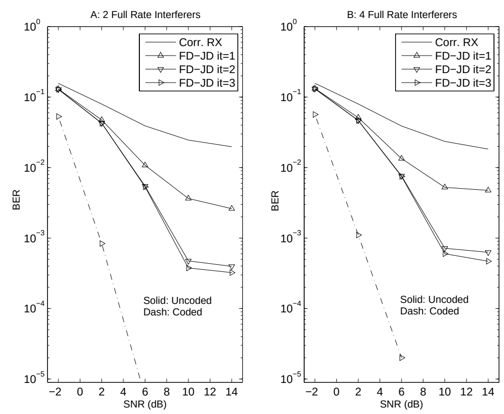

Fig. 4. Average BER for the multiuser case with full rate users (worst case).