Iterative Multiuser Detection for DS-CDMA/MC-CDMA Powerline Communications

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Abstract: In this paper, signal processing techniques to combat the adverse communications environment on power lines are addressed, so as to enable reliable high speed data communications over low-voltage (LV) power distribution networks for Internet access and in-home/office networking. Direct sequence code-division multiple-access (DS-CDMA), Multicarrier CDMA (MC-CDMA), multiuser detection (MUD), and turbo decoding, having demonstrated their limit-approaching capacity in digital subscriber line (DSL) and wireless communication systems, are readily applied to powerline communications, successfully mitigating the influence of time-varying channel attenuation, multipath frequency-selective fading, multiple-access interference, and background noise. Some strategies to deal with the most unfavorable noise, the impulse noise, are also discussed.

I. Introduction

The increasing ubiquity of the Internet is creating a rapidly growing demand for larger bandwidth to the home. Currently the narrowband twisted-pair access network from an optical network unit (ONU) or a central office (CO) to a customer's premises, the so-called "last dirty mile", is the bottleneck for Internet traffic. The increasing demand for a home/office network further necessitates a flexible broadband network access. Electric power lines, which can be found in essentially all buildings and residences, naturally exhibit potential as a convenient and cheap communication alternative. Also, in rural areas where services from telephone companies or cable companies do not reach, and where radio coverage is poor or very expensive through one-way satellite access, communication through power lines may be the only feasible solution. As to home networking, the powerline is inherently the most attractive medium due to its universal existence in homes, the ubiquity of outlets, and the simplicity of the power plug. In comparison, the phone line/cable suffers from too few connection points, and wireless suffers from congestion and interference in the unlicensed bands.

Even though powerline communications is an attractive substitute for broadband Internet access for the last mile and in-home/office networking, many difficulties and challenges exist, as power lines were not originally intended or designed for data communications [1]. To facilitate reliable high-speed communications over low-voltage (LV) power distribution networks, advanced signal processing techniques should be pursued to combat the adverse environment. The dominant sources of impairment for powerline communications are time-varying channel attenuation, multipath frequency-selective fading, multiple-access interference, and impulse noise. These phenomena naturally remind us of the similar impairments and corresponding mitigating techniques used in digital subscriber line (DSL) and wireless communications.

Given the similarities among DSL, wireless and powerline channels, techniques developed for the first two types of channels are natural candidates for application in powerline communications. In this study we consider such application. We adopt DS-CDMA and MC-CDMA as modulation and multiple-access methods, based on which powerline communication model is under consideration (see Section II). On forming a serially concatenated system at the transmitter through introducing an interleaver between coding and modulation modules, multiuser detection and turbo decoding are used at the receiver for data detection and decoding, further details of which are given in Section III. Some numerical results are given in Section IV to demonstrate the performance of the proposed signal processing techniques, in comparison with traditional ones. Finally, Section V concludes the paper.

II. Powerline Communication Model

A mathematical multipath propagation model for the transfer function of powerline channels has been proposed in [11].

$$H(f) = \sum_{i=1}^{N_p} g_i \cdot e^{-(a_0 + a_1 f^{\kappa}) d_i} \cdot e^{-j 2\pi f (d_i / v_P)} \,. \qquad (1)$$

This model is based on physical signal propagation effects in mains networks including numerous branches and impedance mismatching. Besides multipath propagation accompanied by frequency selective fading, signal attenuation of typical power cables increasing with length and frequency is considered. The principal advantage of this model is the comparatively small set of parameters needed. These are the weighing factor g_i, the length d_i of path i with total number of paths N_p, and general parameters of a_0, a_1 and κ for signal attenuation with respect to length and frequency. The propagation speed, v_p, is a constant depending on the cable's insulation material. It has been verified that this model allows a very accurate reproduction of powerline channel behavior and will be used here as basis for channel emulation.

Consider a direct-sequence CDMA communication system with K users employing normalized spreading waveforms $s_1,\ldots,s_K$ given by

$$s_k(t)=\frac{1}{\sqrt{N}}\sum_{j=0}^{N-1}c_k(j)\psi(t-jT_c)\ ,\ 0\le t\le T\ ,\ 1\le k\le K\ , \tag{2}$$

where N is the processing gain, $\{c_k(j)\}_{j=0}^{N-1}$ is a signature sequence of ± 1's assigned to the kth user, and $\psi(\cdot)$ is a normalized chip waveform of duration $T_c=T/N$ with T the symbol duration. User k (for $1\le k\le K$) transmits a frame of M independent equiprobable BPSK symbols $b_k(i)\in\{+1,-1\}$, $0\le i\le M-1$; and the symbol sequences from different users are assumed to be mutually independent. Note that $\{b_k(i)\}$ may be encoded streams derived from underlying information symbols. The transmitted baseband signal due to the kth user is thus given by

$$x_k(t)=A_k\sum_{i=0}^{M-1}b_k(i)s_k(t-iT)\ ,\ 1\le k\le K\ , \tag{3}$$

where A_k is the amplitude associated with user k's transmission. The kth user's signal $x_k(t)$ propagates through a multipath channel with impulse response $h_k(t)$, whose transfer function $H_k(f)$ is in the form of (1). At the receiver, the received signal due to the kth user is then given by

$$y_k(t)=x_k(t)*h_k(t)=A_k\sum_{i=0}^{M-1}b_k(i)\sum_{j=0}^{N-1}c_k(j)\tilde{h}_k(t-iT-jT_c)\ , \tag{4}$$

where $\tilde{h}_k(t)=\psi(t)*h_k(t)$, and * denotes convolution. The signal at the receiver is the superposition of the K users' signals plus the ambient noise, given by

$$r(t)=\sum_{k=1}^{K}y_k(t)+n(t)\ . \tag{5}$$

Usually it is convenient to deal with a discrete-time sufficient statistic, which is derived by passing the received signal $r(t)$ through a chip-matched filter and then sampling at the chip rate. The resulting signal sample at the nth chip interval of the lth symbol interval is given by

$$r_n(l)=\int_{lT+nT_c}^{lT+(n+1)T_c}r(t)\psi(t-lT-nT_c)dt=\sum_{k=1}^{K}y_{kn}(l)+n(l), \tag{6}$$

where

$$\begin{aligned}y_{k,n}(l)&=A_k\sum_{i=0}^{M-1}b_k(i)\sum_{j=0}^{N-1}c_k(j)\int_{lT+nT}^{lT+(n+1)T_c}\tilde{h}_k(t-iT-jT_c)\psi(t-lT-nT_c)dt\\&=A_k\sum_{i=0}^{M-1}b_k(i)\sum_{j=0}^{N-1}c_k(j)g_k(N(l-i)+(n-j))\end{aligned} \tag{7}$$

with the discrete-time channel response for the kth user given by

$$g_k(m)=\int_{mT_c}^{(m+1)T_c}\tilde{h}_k(t)\psi(t-mT_c)dt\ . \tag{8}$$

On denoting $f_k(m)=A_kc_k(m)*g_k(m)$, we have

$$y_{k,n}(l)=\sum_{i=0}^{M-1}b_k(i)f_k(n+N(l-i))\ . \tag{9}$$

With $\mathbf{f}_k(l-i)=[f_k(N(l-i)),\cdots,f_k((N-1)+N(l-i))]^T$ (2)
and $\mathbf{y}_k(l)=[y_{k,0}(l),\cdots,y_{k,(N-1)}(l)]^T$, from (9) we have

$$\mathbf{y}_k(l)=\sum_i b_k(i)\mathbf{f}_k(l-i)=\sum_{j=0}^{L_k-1}\mathbf{f}_k(j)b_k(l-j)\ , \tag{10}$$

where L_k is the effective length of the composite channel response for the kth user relative to the symbol duration. On denoting $L=\max_{1\le k\le K}L_k$, $\mathbf{F}(j)=[\mathbf{f}_1(j),\cdots,\mathbf{f}_K(j)]$, $\mathbf{b}(i)=[b_1(i),\cdots,b_K(i)]^T$, and $\mathbf{r}(l)=[r_0(l),\cdots,r_{N-1}(l)]^T$, we can write

$$\mathbf{r}(l)=\sum_{j=0}^{L-1}\mathbf{F}(j)\mathbf{b}(l-j)+\mathbf{n}(l)\ . \tag{11}$$

Due to the convolutional effect of the multipath channel, L successive samples of the received data vector should be collected as a sufficient statistic for detection of the transmitted data. On denoting $\mathbf{r}=[\mathbf{r}(l)^T,\cdots\mathbf{r}(l+L-1)^T]^T$, $\mathbf{b}=[\mathbf{b}(l-L+1)^T,\cdots,\mathbf{b}(l+L-1)^T]^T$,

$$\mathbf{F}=\begin{bmatrix}\mathbf{F}(L-1)&\cdots&\mathbf{F}(0)&\cdots&\mathbf{0}\\\vdots&\ddots&\ddots&\ddots&\vdots\\\mathbf{0}&\cdots&\mathbf{F}(L-1)&\cdots&\mathbf{F}(0)\end{bmatrix},$$

and $\mathbf{n}=[\mathbf{n}(l)^T,\cdots\mathbf{n}(l+L-1)^T]^T$, we finally have the succinct form of

$$\mathbf{r}=\mathbf{F}\mathbf{b}+\mathbf{n}\ . \tag{12}$$

MC-CDMA systems are the frequency-domain duals of DS-CDMA systems, in which the spreading is carried out in the frequency domain instead of the time domain. The effects of a frequency-selective channel can be analyzed in the frequency domain as convolution is replaced by multiplication. Let us assume for simplicity that the total bandwidth is divided into N subchannels with the center frequency of each subchannel given by

$$f_{c,i} = \frac{i}{T} = \frac{i}{NT_c} = \frac{i}{N} B_T \;,\; i = 1, \cdots, N \;, \qquad (13)$$

where as before N is the processing gain, T is the symbol interval, T_c is the notional chip duration, and B_T is the total bandwidth. Each user assumes a transmitted signal in a form analogous to (2) and (3), but in the frequency domain, i.e.,

$$X_k(i) = A_k b_k s_{ki} = \frac{1}{\sqrt{N}} A_k b_k c_{ki} \;, \qquad (14)$$

where $\{X_k(i)\}$ is the discrete Fourier transform (DFT) of the sampled transmitted signal of the kth user. We assume that the subchannel bandwidth is less than the channel coherence bandwidth so as to experience a frequency-flat fading represented by corresponding gains,

$$H_{k,i} = \alpha_{k,i} e^{j\theta_{k,i}} \;. \qquad (15)$$

On denoting

$$\mathbf{s}_k = [s_{k,1}, \cdots, s_{k,N}]^T \;, \mathbf{H}_k = [H_{k,1}, \cdots, H_{k,N}]^T \;,$$

and $\mathbf{f}_k = A_k \mathbf{s}_k \circ \mathbf{H}_k$ as containing the element wise product of $A_k \mathbf{s}_k$ and $\mathbf{H}_k$, the received signal in the frequency domain is given by a form similar to (12):

$$\mathbf{r} = \mathbf{F}\mathbf{b} + \mathbf{n} \;, \qquad (16)$$

where $\mathbf{r} = [r_1, \cdots, r_N]^T$ collects the discrete received spectrum in the N subcarriers, $\mathbf{F} = [\mathbf{f}_1, \cdots, \mathbf{f}_K]$, $\mathbf{b} = [b_1, \cdots, b_K]^T$, and $\mathbf{n} = [n_1, \cdots, n_N]^T$ collects the noise. Since the received signals in DS-CDMA and MC-CDMA can be expressed in the same form, the receiver signal processing described below can be applied to either system. Consequently in the following, for simplicity, we will illustrate only MC-CDMA systems.

III. Turbo Multiuser Detection for Powerline Communications

In Fig. 1, a convolutionally encoded multiuser MC-CDMA system is shown. For each user k, $1 \le k \le K$, the information bits $\{d_k\}$ are first encoded into coded bits with a standard binary convolutional encoder with code rate R. A code-bit interleaver is used to decorrelate the noise on the coded bits at the input of the channel decoder. The interleaved coded bits $\{b_k\}$ are spread across the N subchannels and mapped to quadrature amplitude modulation (QAM) signals. Then the conjugate-symmetric vector of length $\bar{N} = 2N$ is transformed using the IFFT to get a real time-domain vector. After parallel-to-serial and digital-to-analog conversion, the signal of the kth user $x_k(t)$ is transmitted into the channel, where it is corrupted by additive multiple-access signals and background noise. At the receiver end, after analog-to-digital and serial-to-parallel conversion, the received signal is transformed back to the frequency domain using an FFT, where it can be written as in (16).

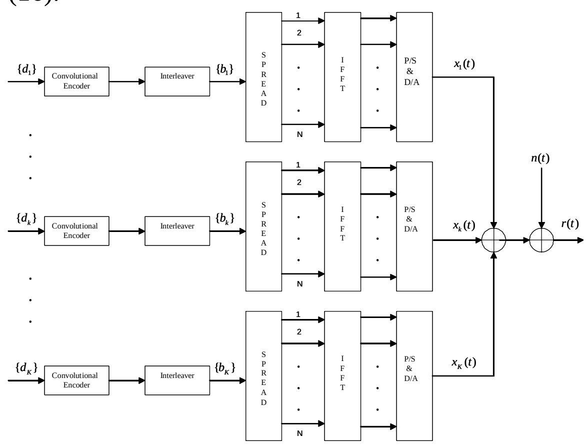

Fig. 1 A coded multiuser MC-CDMA system

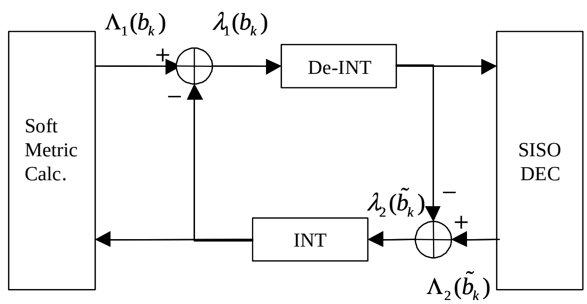

Fig. 2 Turbo structure for iterative demodulation and decoding

Figure 2 shows the turbo structure for turbo multiuser detection and decoding. It consists of iteration between two stages: a soft metric calculator (the demodulation stage) and a soft-input soft-output (SISO) channel decoder (the decoding stage). The two stages are separated by an interleaver and a de-interleaver. A channel log-likelihood ratio (LLR) for the interleaved coded bit of the kth user is calculated as follows:

$$\Lambda_1(b_k) = \underbrace{\log \frac{p(\{r(t)\} \mid b_k = 1)}{p(\{r(t)\} \mid b_k = -1)}}_{\lambda_1(b_k)} + \underbrace{\log \frac{P(b_k = 1)}{P(b_k = -1)}}_{\lambda_2^p(b_k)} \;, \qquad (17)$$

where the second term $\lambda_2^p(b_k)$ represents the *a priori* LLR delivered from the decoding stage in the previous iteration. For the first iteration, this term is set to zero if we assume equally likely coded bits. The first term $\lambda_1(b_k)$, denoting the extrinsic information obtained from the demodulation stage about the bit b_k, is then de-interleaved and sent to the channel decoder as its *a priori* information. Similarly, the SISO channel decoder computes the *a posteriori* LLR of each coded bit and then excludes the influence of *a priori* knowledge to get extrinsic information from the decoding stage about the bit $\tilde{b}_k$ as follows:

$$\lambda_2(\tilde{b}_k) = \Lambda_2(\tilde{b}_k) - \lambda_1^p(\tilde{b}_k) = \log\frac{P(\tilde{b}_k = 1 \mid \text{decoding})}{P(\tilde{b}_k = -1 \mid \text{decoding})} - \lambda_1^p(\tilde{b}_k), \quad (18)$$

where $\tilde{b}_k$ is the de-interleaved version of b_k, alternatively the coded bits before the interleaver in Fig. 1. Again, this extrinsic information is interleaved and fed back to the demodulation stage as *a priori* knowledge for the next iteration. At the last iteration, the SISO decoder also computes the *a posteriori* LLRs for information bits, which are used to make final decisions.

In the demodulation stage of (17), either the optimum ML multiuser detection or the sub-optimum MMSE parallel interference cancellation (PIC) can be used. We will show in Section IV that these two schemes achieve the same performance, owing to the turbo processing.

The demodulation stage with ML is straightforward. For each symbol interval, the extrinsic information for the code bit of the *k*th user is given by

$$\begin{aligned}\lambda_1(b_k) &= \Lambda_1(b_k) - \lambda_2^p(b_k) \\ &= \log\frac{\sum_{\mathbf{b}\in B_k^+} p(\mathbf{r}\mid\mathbf{b})p(\mathbf{b})}{\sum_{\mathbf{b}\in B_k^+} p(\mathbf{r}\mid\mathbf{b})p(\mathbf{b})} - \log\frac{P(b_k=1)}{P(b_k=-1)},\end{aligned} \quad (19)$$

where

$B_k^+ = \{(b_1, b_2, \ldots, b_K)^T : b_k = 1\}$ $B_k^- = \{(b_1, b_2, \ldots, b_K)^T : b_k = -1\}$; $p(\mathbf{r}\mid\mathbf{b})$ is a multivariate Gaussian distribution (see (16))[1]; $\lambda_2^p(b_k) = \log\left(P(b_k = 1)/P(b_k = -1)\right)$ and $p(\mathbf{b}) = \prod_k p(b_k)$ comprise *a priori* information from the decoding stage.

The demodulation stage with PIC is subtler. Suppose the received signal for some user $1 \le k \le N$ after interference cancellation is given by

$$\hat{\mathbf{r}}_k = \mathbf{F}(\mathbf{b} - \hat{\mathbf{b}}_k) + \mathbf{n}, \quad (20)$$

where $\hat{\mathbf{b}}_k = (\hat{b}_1, \hat{b}_2, \ldots \hat{b}_k = 0, \ldots, \hat{b}_K)^T$ is the estimated interference vector. First an MMSE filter is applied to $\hat{\mathbf{r}}_k$ to further suppress the residual interference plus noise, that is,

$$\mathbf{w}_k = E[\hat{\mathbf{r}}_k\hat{\mathbf{r}}_k^H]^{-1}E[\hat{\mathbf{r}}_k b_k^*] = (\mathbf{f}_k\mathbf{f}_k^H + \mathbf{F}_k\mathbf{Q}\mathbf{F}_k^H + \sigma^2\mathbf{I})^{-1}\mathbf{f}_k, \quad (21)$$

where $\mathbf{f}_k$ is the *k*th column of matrix $\mathbf{F}$, $\mathbf{F}_k$ is the complement of $\mathbf{f}_k$ in $\mathbf{F}$, and

$$\mathbf{Q} = \mathrm{diag}[1-|\hat{b}_1|^2, \ldots, 1-|\hat{b}_{k-1}|^2, 1-|\hat{b}_{k+1}|^2, \ldots, 1-|\hat{b}_K|^2],$$

which approaches 0 when estimates from the decoding stage are accurate enough for unit-modulus signals. As is shown in [6], the output of the MMSE filter $z_k = \mathbf{w}_k^H\hat{\mathbf{r}}_k$ can be well approximated as

$$z_k = \mu_k b_k + \eta_k, \quad (22)$$

where $\mu_k = E[z_k b_k^*] = \mathbf{w}_k^H\mathbf{f}_k$, and η_k is a Gaussian variable with zero mean and variance

$$\nu_k^2 = E[|z_k - \mu_k b_k|^2] = E[|z_k|^2] - |\mu_k|^2 = (\mu_k - \mu_k^2). \quad (23)$$

The extrinsic information is given in the same form as (19), but with $\mathbf{r}$ replaced by z_k and $\mathbf{b}$ with b_k, and (16) replaced with (22), and therefore with much lower complexity.

For the SISO decoding, either the optimum maximum *a posteriori* probability (MAP) algorithm or suboptimum Max-log-MAP or SOVA algorithms can be used. The reader is referred to [2] for details.

IV. Numerical Results

In this section, we simulate a multiple-access high-speed powerline communication channel with $K = 4$ users, with which the proposed advanced signal processing techniques are tested and compared with some traditional detection techniques.

For the channel model (1), we adopt the same set of parameters for the four users: $N_p = 4$, $a_0 = 0$, $a_1 = 7.8\times10^{-10}$, $\kappa = 1$ (see [11]). The users are ordered by their distance to the line termination, with user 1 being the closest. For each user, the multipath weighting factors are independent normalized complex Gaussian random variables, and the lengths of paths are uniformly distributed within a certain range. The simulated channel frequency responses are shown in Fig. 3, where the frequency-dependent attenuation and the frequency selective fading can be easily seen.

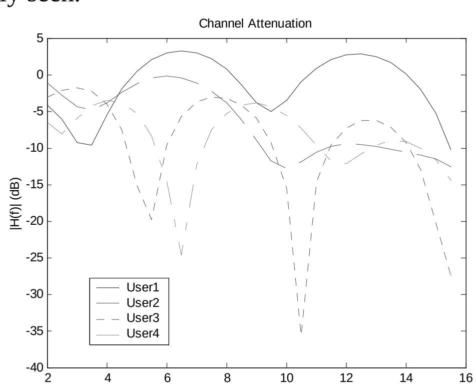

Fig. 3 Simulated channel frequency responses of four users

First we examine the immunity of single-carrier and MC-CDMA systems to frequency-selective

[1] Impulse noise will be treated separately.

fading channels and impulse noise. The single carrier system employs the carrier frequency of 3.5 MHz with a bandwidth of 0.5 MHz, with BPSK modulation. The MC-CDMA system occupies from 2 to 16 MHz with $N = 28$ subchannels, the center frequencies of which are given by $f_n = 2 + 0.5(n-1)$, $1 \le n \le 28$. For each subchannel, BPSK modulation is used. For ease of comparison, we assume a single-user uncoded system. The user of interest is user 1.

To simulate the influence of impulse noise, we adopt the commonly used two-term Gaussian mixture model as proposed in [3]. The first-order probability density function of this noise model has the form

$$(1-\varepsilon)\boldsymbol{N}(0,\sigma^2) + \varepsilon\boldsymbol{N}(0,\kappa\sigma^2), \qquad (24)$$

with $\sigma > 0$, $0 \le \varepsilon \le 1$, and $\kappa \ge 1$. Here, the $\boldsymbol{N}(0,\sigma^2)$ term represents the nominal background noise (Gaussian with zero mean and variance σ^2), and the $\boldsymbol{N}(0,\kappa\sigma^2)$ term represents an impulse component (Gaussian with zero mean and variance $\kappa\sigma^2$), with ε representing the probability that impulses occur. In our simulation we choose parameters $\varepsilon = 0.01$, which means the impulse occurs with 1% disturbance ratio [10]. According to the observation in [10], when an impulse occurs, the noise power spectral density (PSD) is colored and the overall power level is raised. Usually the spectral power of the impulse noise is concentrated in particular frequency ranges, due to the oscillating behavior of the impulse noise. In our simulation, we increases the noise PSD to 20dB higher for the frequency range of 3-6 MHz when an impulse occurs, and we set the impulse width to be of 100 μs, lasting as long as 50 symbol intervals.

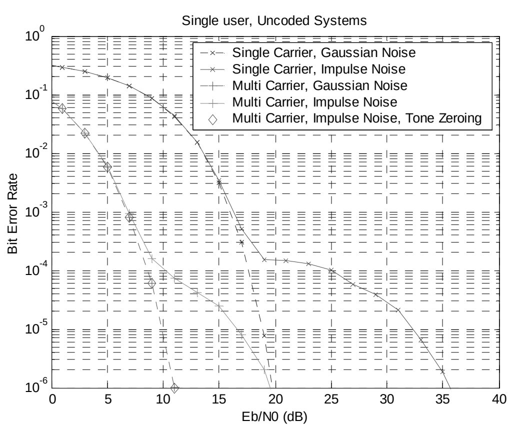

Fig. 4 Performance comparison of single-carrier and MC-CDMA systems with Gaussian and impulse noise

From Fig. 4, we can see that, with the Gaussian background noise, MC-CDMA offers almost 10dB gain over the single carrier system at a bit-error-rate (BER) of 10^{-6}. We have normalized the transmitted power of the parallel subchannels of the MC-CDMA system, so that they are compared for the same E_b / N_0. From Fig. 3 we see that there is a fading notch in the band of 3.5 MHz for user 1, which results in the poor performance of the single carrier system, while the inherent spectral diversity of the MC-CDMA system significantly improves the system performance. The MC-CDMA system is also more robust to the influence of the impulse noise, as we can see from the Fig. 4, for the same reasons. Furthermore, for the MC-CDMA system, the infected tones can be easily zeroed, resulting in almost no loss in system performance in the presence of impulse noise. The main challenge of tone zeroing (A form of erasure decoding) is in finding practical methods of obtaining fairly reliable channel state information. One may argue that the single carrier system can choose a favorable band for data communication, but this will add complexity to the transmitter and the protocols, and even may not be possible due to the rapid time-varying nature of powerline channels.

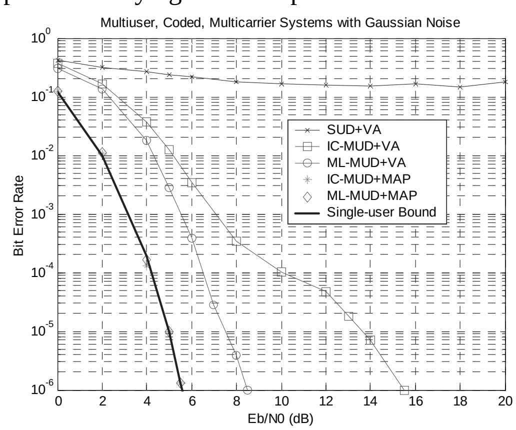

Fig. 5 Performance Comparison of various detectors for coded MC-CDMA systems with multiple users and Gaussian noise

In Fig. 5, our proposed turbo multiuser detectors are tested with a coded multiuser MC-CDMA system as shown in Fig. 1, with the Gaussian background noise. A rate-1/2 convolutional code with constraint length 5 and generator polynomials $[23, 35]_8$ is used for channel coding. The number of information bits per block per user is set as 996. Each user uses a different random interleaver of length 2000 for interleaving and de-interleaving. For simplicity, the spreading gain is set as $N = 8$, and the used subchannels are {2, 3.5, 5, 6.5, 8, 9.5, 11. 12.5} MHz. These so-called comb spread carriers are commonly used for multiple access purposes to improve the frequency diversity. The spreading code for each user is independently and randomly generated. The channel responses of the four users are given in Fig. 3. The user of interest is user 4, the weakest one. For ease of comparison and reference, the channel of user 4 has been normalized, and the other channels have been adjusted accordingly.

There are six detectors of interest in Fig. 5. IC-MUD+MAP and ML-MUD+MAP are our proposed turbo multiuser receivers, with MMSE parallel interference cancellation or maximum likelihood demodulation stages, respectively, and a MAP decoding stage. IC-MUD+VA and ML-MUD+VA are their non-iterative counterparts: after multiuser detection, hard decisions are made on coded bits, then the Viterbi algorithm (VA) is used for decoding. Also shown in the figure is the traditional detection method, SUD+VA, which ignores the multiple-access interference, and the single user bound, which assumes no multiple-access interference.

From Fig. 5 we can see that there is a tremendous performance gap between the traditional single user detector and the single user bound. Optimum ML multiuser detection significantly narrows this gap down to 3dB at a BER of 10^{-6}, but it suffers from a complexity exponentially increasing with the number of users. The suboptimum interference cancellation method, even though a good tradeoff between performance and complexity, suffers an extra 7dB loss. Both of our proposed turbo multiuser receivers, however, approaches the single user bound. It is worth noting that IC-MUD+MAP achieves this excellent performance with reasonable computational complexity, making it very appealing for practical systems.

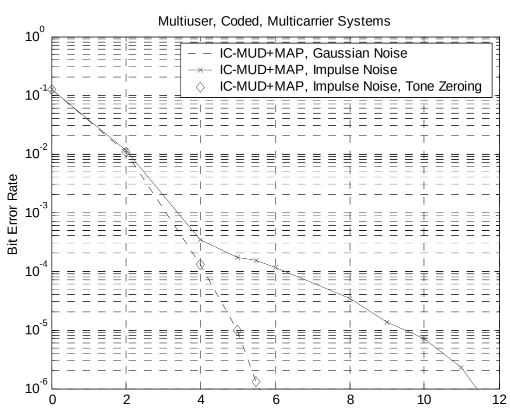

Fig. 6 Performance comparison of a proposed turbo multiuser receiver with Gaussian and impulse noise

When the same impulse noise setting in Fig. 4 is introduced, the performance of IC-MUD+MAP deteriorates about 5dB at a BER of 10^{-6}, which is easily recovered by tone zeroing or erasure decoding, as seen in Fig. 6.

V. Conclusions

In this paper, advanced signal processing techniques well developed for DSL and wireless communications have been applied to high speed powerline communications and have been seen to achieve satisfactory results therein. To be specific, coded MC-CDMA systems have been employed for data transmission, and multiuser detection and turbo decoding have been used for data detection. The proposed communication systems achieve obvious advantages over single carrier systems with respect to time varying channel attenuation, multipath frequency-selective fading, and impulse noise. The proposed turbo multiuser receivers effectively mitigate the multiple-access interference and approach the single user bound. The detrimental effects of impulse noise to the proposed scheme are remedied through erasure decoding techniques.

In this study, we have assumed that the receiver has knowledge of the channel. In reality, however, channel identification is needed, and the effects of channel estimation errors should be taken into consideration. The problem of detecting impulse spike positions (for erasure decoding purposes), both in time and frequency, also deserves further study.

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