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Iterative Multiuser Detection and Decoding for DMT VDSL Systems

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***Abstract* – In recent years, iterative processing techniques with soft-in/soft-out (SISO) components have received considerable attention. Such techniques, based on the so-called turbo principle, are exemplified through turbo decoding, turbo equalization and turbo multiuser detection. In this paper, turbo multiuser detection is applied to a discrete multitone (DMT) very-high-rate digital subscriber line (VDSL) system to combat crosstalk signals and to obtain substantial coding gain. The proposed iterative DMT receiver is shown to achieve an overall 7.0 dB gain over the uncoded optimum receiver at a bit error rate of 10^{-7} for a channel with severe intersymbol interference and additive white Gaussian noise and with one dominant crosstalk signal. Impulse noise is detrimental to the proposed scheme but can be overcome through erasure decoding techniques, as is shown by example.**

I. INTRODUCTION

Digital Subscriber Line (DSL) technology provides transport of high-bit-rate digital information over telephone subscriber lines. The latest in DSL technology is Very-high-rate DSL (VDSL), which provides tens of megabits per second to those customers who desire broadband entertainment or data services. Intersymbol interference (ISI) is one of the major obstacles to high-data-rate, bandwidth efficient communications. Multicarrier modulation (MCM), following Shannon's optimum transmission suggestion, achieves the highest performance in channels with ISI. A particular form of MCM, known as discrete multitone (DMT), has been found to be well suited for DSL application and is adopted in ANSI T1.413 ADSL standards [12]. Synchronized DMT (SDMT), which exploits time-division duplexing (TDD) for simple programmability of downstream-to-upstream data rates and for reducing the impact of self near-end crosstalk (NEXT), has been proposed as a candidate for the VDSL standard. Noise on a phone line usually occurs because of imperfect balance of the twisting pair. There are many types of noises that couple into the phone line, the most common of which are crosstalk noise, radio noise and impulse noise. While the radio noise problem can be solved or at least alleviated by restricting VDSL transmission within radio bands, crosstalk and impulse noise are two principal sources of degradation in VDSL transmission systems. The traditional single-user detector (SUD) for such systems merges crosstalk into the background noise, which is assumed to be white and Gaussian. Recent research has explored the nature of crosstalk signals and has shown the potential benefits of robust multiuser detection (MUD) to jointly mitigate crosstalk and impulsive noise for contaminated VDSL signals [3], [4].

Coding is a common way to reduce the gap in channel capacity experienced by uncoded systems. A concatenated coding scheme consisting of an inner trellis code (a 4-D Wei code) and an outer Reed-Solomon (RS) code was proposed for ADSL DMT systems to provide a 5 dB coding gain at bit-error-rate (BER) 10^{-7} without bandwidth expansion [15]. There are two problems with this approach. First, since the constellation size varies from tone to tone, a time-varying trellis-coded modulation (TCM) encoder is required. Second, further improvement is very difficult from a practical implementation perspective because of the complexity of Viterbi decoding for the multidimensional TCM. Recently, turbo coding has been proposed for DMT systems [6], [9]. One typical turbo code is used to code across all subchannels. The coded bits are then interleaved and allocated to various tones for quadrature amplitude modulation (QAM). Thus, a single standard binary decoder can be employed at the receiver and further improvement in turbo coding is easily incorporated. A coding gain of 6.0 dB for bandwidth efficiency of 2 bits/s/Hz and 4.1 dB for 3 bits/s/Hz was reported for a channel with severe ISI at BER 10^{-5} [9].

In recent years, iterative processing techniques with soft-in/soft-out (SISO) components have received considerable attention. The basic idea is to break up optimum joint signal processing, e.g. concatenated decoding, joint equalization and decoding, or joint decoding and multiuser detection (MUD), which is typically very complex and require large amounts of memory, into separate components, iterating between them with the exchange of probabilities or "soft" information. This approach typically results in almost no loss of information. This so-called turbo principle is exemplified through turbo decoding [7], turbo equalization [5] and turbo multiuser detection [10], [14]. In this paper, we consider the application of turbo multiuser detection in a coded DMT VDSL system to combat crosstalk and to obtain substantial coding gain. We also consider the effect of impulse noise, which has been found to greatly impact the performance of the channel decoding stage, and an erasure decoding technique is proposed as a remedy.

The paper is organized as follows. In Section II a signal model for the DMT VDSL communication system is described, together with the iterative receiver structure for multiuser detection and channel decoding, and the impulse noise model. In Section III, we describe MUD-based schemes for

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DMT VDSL signal detection, while in Section IV details of the iterative decoding process are given. Simulation results are given in Section V, and Section VI concludes the paper.

II. SYSTEM DESCRIPTION

We consider a convolutionally encoded DMT system with crosstalk as shown in Fig. 1. The information bits **d** are first encoded into coded bits **b** 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 bits are optimally allocated to $\overline{N}$ subchannels and mapped to QAM signals of various constellation sizes. Then the conjugate-symmetric vector of length $N = 2\overline{N}$ is transformed using the inverse fast Fourier transform (IFFT) to get a real time-domain vector. After serial-to-parallel and digital-to-analog conversion, the DMT VDSL signal $x(t)$ is transmitted into the channel, where it is corrupted by additive coupled crosstalk signals and background noise. At the receiver end, after inverse 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

$$Y_n = H_n \cdot X_n + \sum_{k=2}^{K} F_{n,k} \cdot C_{n,k} + N_n \,, \quad n = 1,\ldots,\overline{N} \,, \qquad (1)$$

where for the nth subchannel, H_n is the channel gain, X_n is the transmitted (complex) DMT symbol, $C_{n,k}$ is the kth crosstalk signal, $k = 2, \ \ldots, \ K$, $F_{n,k}$ is the corresponding crosstalk coupling function, and N_n is the background noise. Output values of the FFT are fed into the demodulator and decoder for further processing.

Figure 2 shows the turbo structure for iterative demodulation and decoding. It consists of two stages: a soft metric calculator (the demodulation stage) and a SISO channel decoder (the decoding stage). The two stages are separated by an interleaver and a de-interleaver. The crosstalk signals are first detected via a multiuser detection technique, discussed in Section III. Then a channel log-likelihood ratio (LLR) for the kth bit carried by the ith subchannel symbol is calculated as follows:

$$\Lambda_1(b_{k,i}) = \log \frac{P(b_{k,i} = 1 \,|\, r(t))}{P(b_{k,i} = -1 \,|\, r(t))} \,, \qquad (2)$$

where $\{r(t)\}$ is the received waveform as shown in Fig. 1. Using Bayes' formula and discarding the common term $p(r(t))$, (2) can be written as

$$\Lambda_1(b_{k,i}) = \underbrace{\log \frac{p(r(t) \,|\, b_{k,i} = 1)}{p(r(t) \,|\, b_{k,i} = -1)}}_{\lambda_1(b_{k,i})} + \underbrace{\log \frac{P(b_{k,i} = 1)}{P(b_{k,i} = -1)}}_{\lambda_2{}^p(b_{k,i})} \,, \qquad (3)$$

where the second term $\lambda_2^p(b_{k,i})$ 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 $\lambda_1(b_{k,i})$, denoting the extrinsic information obtained from the demodulation stage about the bit $b_{k,i}$, is then de-interleaved and sent to the channel decoder as *a priori* information. Similarly, the SISO channel decoder computes the *a posteriori* LLR of each code bit and then excludes the influence of *a priori* knowledge to get extrinsic information from the decoding stage about the bit b_j as follows:

$$\begin{aligned} \lambda_2(b_j) &= \Lambda_2(b_j) - \lambda_1^p(b_j) \\ &= \log \frac{P(b_j = 1 \,|\, \text{decoding})}{P(b_j = -1 \,|\, \text{decoding})} - \lambda_1^p(b_j) \end{aligned} \,. \qquad (4)$$

The above factorization is derived in Section IV. 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* LLR for information bits, which is used to make final decisions.

Impulse noise is a severe impairment to DSL transmission, especially after long loop attenuation (at a residential location) and at high frequencies (where the DSL signal is more severely attenuated). To model the behavior of the impulse noise we use a two-term Gaussian mixture model in the frequency domain as proposed in [4]. The first-order probability density function of the noise model has the form

$$(1-\varepsilon) N(0,\sigma^2) + \varepsilon N(0,\kappa\sigma^2) \qquad (5)$$

with $\sigma > 0$, $0 \le \varepsilon \le 1$, and $\kappa \ge 1$. Here, the $N(0,\sigma^2)$ term represents the nominal background noise, and the $N(0,\kappa\sigma^2)$ term represents an impulse component, with ε representing the probability that impulses occur in a given subchannel. It is assumed that impulse noise samples in disjoint frequency bins are independent.

III. MITIGATION OF CROSSTALK VIA MULTIUSER DETECTION

Let us consider the detection problem for the data model given in (1). The traditional single user detector demodulates QAM symbols tone-by-tone independently. On the other hand, joint maximum-likelihood detection of both VDSL and crosstalk signals selects a set of $\overline{N}$ inputs $\{X_n\}$ and the crosstalk sequence $\mathbf{C}_k^{(i)} = \left\{C_{1,k}{}^{(i)}, C_{2,k}{}^{(i)}, \ldots, C_{\overline{N},k}{}^{(i)}\right\}, \quad k = 2,\ldots,K$, to satisfy

$$X_n = \arg\left\{ \min_{\{X_n\},\{C_{n,k}{}^{(i)}\}} \sum_{n=1}^{\overline{N}} \left| Y_n - H_n \cdot X_n - \sum_{k=2}^{K} F_{n,k} \cdot C_{n,k}{}^{(i)} \right|^2 \right\},$$

$$n = 1,\ldots,\overline{N} \,, \qquad (6)$$

where the minimization is taken over the DMT signaling alphabet and all possible crosstalk sequences $\mathbf{C}_k = \left[\mathbf{C}_k^{(i)}, i = 1,\ldots,\left|\mathbf{C}_k\right|\right], \ k = 2,\ldots,K$, that can occur within the VDSL symbol period of interest. The size $\left|\mathbf{C}_k\right|, \ k = 2,\ldots,K,$ of the set of all possible crosstalk sequences can be large but is always finite when all the crosstalkers are digital signals or are derived from digital signals.

Just as its counterpart in wireless CDMA, the maximum-likelihood multiuser detector (ML-MUD) achieves optimum performance but suffers from very high complexity. An alternative approach is to employ interference cancellation (IC) [13], i.e., to attempt excision of the crosstalk from the received signal before applying the traditional DMT VDSL signal detection. The interference cancellation approach is based on a natural idea: if decisions have been made about an interfering user's information bits, the interfering signal can be reconstructed and subtracted at the receiver. This will achieve perfect interference elimination if the decisions were correct; otherwise, things could be worse than without the canceller. Nevertheless, in VDSL applications, incorrect detection may not be as bad as one might think. In SDMT, self-NEXT is almost completely avoided by TDD, so NEXT from other communication systems dominates. Usually these crosstalk signals are detected in the time domain and a DMT symbol interval will contain more than one crosstalk bit. Often distortion effects in the time (resp. frequency) domain will diffuse in the frequency (resp. time) domain, in which further gain can be obtained by making hard decisions in the high-SNR scenario. These phenomena make IC a good choice in this application.

Our interference cancellation multiuser detection (IC-MUD) scheme is described as follows:

1) A hard decision is made on the VDSL signal in the frequency domain. This decision can be obtained through the received signal $\mathbf{Y}=(Y_1,\ldots,Y_{\overline{N}})^T$ directly (the first iteration) or through soft LLR values λ_2 from a SISO decoder.
2) The DMT symbols $\hat{\mathbf{X}}=(X_1,\ldots,X_{\overline{N}})^T$ are reconstructed based on these detected bits.
3) The desired signal is subtracted and the known crosstalk coupling function is applied to get a frequency domain estimate of the entire crosstalk sequence $\tilde{\mathbf{C}}_k=(\tilde{C}_{1,k},\ldots,\tilde{C}_{\tilde{N},k})^T$ via
$$\tilde{\mathbf{C}}_k=\mathbf{F}_k^{-1}\circ(\mathbf{Y}-\mathbf{H}\circ\hat{\mathbf{X}}-\sum_{i<k}\mathbf{F}_i\circ\hat{\mathbf{C}}_i)\,, \qquad (7)$$
where $\mathbf{H}=(H_1,\ldots,H_{\overline{N}})^T$, $\mathbf{F}_i=(F_{1,i},\ldots,F_{\overline{N},i})^T$, $\mathbf{F}_k^{-1}=(1/F_{1,k},\ldots,1/F_{\overline{N},k})^T$, $\hat{\mathbf{C}}_i$ are formerly detected and recreated crosstalk signals, and "$\circ$" denotes Kronecker (elementwise) product.
4) A time-domain sequence is obtained through the IFFT, $\tilde{\mathbf{c}}_k=\mathrm{IFFT}(\tilde{\mathbf{C}}_k)$, after which hard decisions are made and the crosstalk signal is recreated and transformed to the frequency domain to get $\hat{\mathbf{C}}_k=(\hat{C}_{1,k},\ldots,\hat{C}_{\tilde{N},k})^T$.
5) The above process is repeated until all crosstalk signals $\hat{\mathbf{C}}=\left[\hat{\mathbf{C}}_k, k=2,\ldots,K\right]$ are estimated and reconstructed.
6) Finally, SUD is used for DMT signal detection, i.e.,
$$X_n=\arg\left\{\min_{\{X_n\}}\left|Y_n-H_n\cdot X_n-\sum_{k=2}^{K}F_{n,k}\hat{C}_{n,k}\right|^2\right\},$$
$$n=1,\ldots,\overline{N}\,. \qquad (8)$$

IV. ITERATIVE DECODING FOR CODED DMT SYSTEM

In contrast to trellis-coded modulation where coding and modulation are considered jointly to obtain an optimal signal constellation in the sense of maximum Euclidean distance in the signal space, the iterative decoding process we propose separates demodulation and channel decoding stages for ease of implementation. The greatest benefits for this turbo process are the very large increase in code memory due to the interleaver, and almost no loss of optimality due to the iterative exchange of extrinsic soft information.

Assuming the independence of the background noise between subchannels and between the two dimensions of each subchannel, the soft metric delivered by the demodulation stage (see (3)) can be expressed as

$$\lambda_1(b_{k,i})=\log\frac{\sum_{X_i:b_{k,i}=1}p\left(Y_i\left|X_i,\sum_{m=2}^{K}\hat{C}_{i,m}\right.\right)\prod_{l\neq k}P(b_{l,i})}{\sum_{X_i:b_{k,i}=0}p\left(Y_i\left|X_i,\sum_{m=2}^{K}\hat{C}_{i,m}\right.\right)\prod_{l\neq k}P(b_{l,i})}\,, \qquad (9)$$

where the summations are over all possible DMT symbol realizations with the indicated conditions. In (9),

$$P(b_j)=\frac{1}{2}[1+b_j\tanh(\lambda_2{}^p(b_j))]\,, \qquad (10)$$

where $\lambda_2^p(b_j)$ represents the interleaved *a priori* LLR delivered from the channel decoding stage in the previous iteration. Invoking the Gaussian assumption on the background noise, we can write

$$p\left(Y_i\left|X_i,\sum_{m=2}^{K}\hat{C}_{i,m}\right.\right)=\frac{1}{2\pi\sigma^2}e^{-\frac{\left|Y_i-H_iX_i-\sum_{m=2}^{K}F_{i,m}\hat{C}_{i,m}\right|^2}{2\sigma^2}}\,, \qquad (11)$$

where σ^2 is the noise variance per dimension. The crosstalk signals are detected using the MUD schemes discussed in Section III. For ML-MUD, the crosstalk signals are detected and subtracted from the received signal before iteration begins. For IC-MUD, after each iteration, the LLR from the SISO decoder is used to get refined estimates of the VDSL signals and thus to help better detect the crosstalk signals. Due to the independence of the in-phase and quadrature data, the above formulas (9) through (11) can be treated for each dimension separately.

For the channel decoding stage, we consider a binary rate-1/*n* convolutional code with constraint length υ. The BCJR decoding algorithm [1] is known to yield the optimal symbol estimate with minimum symbol-error-rate. What is more important for turbo processing is the BCJR algorithm's ability to yield soft information in the form of *a posteriori* LLRs for coded and information bits. Using the same notation as in [1] and [14], for stage *t* of the code trellis transiting from state $S_{t-1}=s'$ to $S_t=s$ associated with input d_t and output $\mathbf{b}_t=(b_t^1,\ldots,b_t^n)$, we have

$$\Lambda(b_t^k) = \log \frac{\sum_{S_k^+} \alpha_{t-1}(s')\gamma_t(s',s)\beta_t(s)}{\sum_{S_k^-} \alpha_{t-1}(s')\gamma_t(s',s)\beta_t(s)}, \tag{12}$$

where S_k^+ is the set of state pairs (s',s) such that the *k*th coded bit at stage *t* is 1 and S_k^- the corresponding set for -1; and

$$\Lambda(d_t) = \log \frac{\sum_{U_k^+} \alpha_{t-1}(s')\gamma_t(s',s)\beta_t(s)}{\sum_{U_k^-} \alpha_{t-1}(s')\gamma_t(s',s)\beta_t(s)}, \tag{13}$$

where U_k^+ is the set of state pairs (s',s) such that the information bit at stage *t* is 1 and U_k^- is the corresponding set for -1.

As no channel outputs are available for the outer code of a concatenated coding system, we have

$$\begin{aligned} \gamma_t(s',s) &= P(s \mid s') \\ &= \prod_{l=1}^{n} P(b_t^l) = \prod_{l=1}^{n} \frac{1}{2}[1 + b_t^l \tanh(\lambda_1^p(b_t^l))], \end{aligned} \tag{14}$$

where $\lambda_1^p(b_t^j)$ represents the de-interleaved *a priori* LLR delivered from the demodulation stage that is related to the state transition of $s' \to s$. The α_t and β_t terms are defined with forward and backward recursions as

$$\alpha_t(s) = \sum_{s'} \alpha_{t-1}(s')\gamma_t(s',s), \tag{15}$$

and

$$\beta_t(s) = \sum_{s'} \beta_{t+1}(s')\gamma_{t+1}(s,s'), \tag{16}$$

with boundary conditions

$$\alpha_0(0) = 1, \text{ and } \alpha_0(s) = 0, \text{ for } s \neq 0, \tag{17}$$

and

$$\beta_\tau(0) = 1, \text{ and } \beta_\tau(s) = 0, \text{ for } s \neq 0, \tag{18}$$

where τ denoted the frame length of the information bits. The summations of (15) and (16) are over all states s' for which the transition $s' \leftrightarrow s$ is possible. So the extrinsic information (4) produced by the SISO channel decoder can be written as

$$\lambda_2(b_t^k) = \log \frac{\sum_{S_k^+} \alpha_{t-1}(s')\beta_t(s) \prod_{l=1, l\neq k}^{n} P(b_t^l)}{\sum_{S_k^-} \alpha_{t-1}(s')\beta_t(s) \prod_{l=1, l\neq k}^{n} P(b_t^l)} = \Lambda_2(b_t^k) - \lambda_1^p(b_t^k) \tag{19}$$

where $\{b_j\} \to \{b_t^k\}$ with $j = (t-1)n + k$ for a rate-$1/n$ convolutional code.

The BCJR algorithm is known to have numerical problems associated with the representations of probabilities due to the large dynamic range of α_t and β_t. Thus it is preferable that operations be processed in the logarithmic domain. Other well known simplified SISO decoding algorithms include Max-Log-MAP and SOVA, whose implementations are omitted here due to space limitations (see, e.g., [7], [11]).

V. SIMULATION RESULTS

In this section we examine the behavior and the performance of the proposed turbo multiuser detection receivers for coded DMT-VDSL signals with crosstalk via computer simulations. The main results are for an AWGN channel, although impulse noise issues are also addressed briefly. Bit-error-rate is adopted as the performance measure with respect to the geometric signal-to-noise ratio (SNR), which is defined as

$$\mathrm{SNR}_{geo} = \Gamma \cdot \left[\left(\prod_{i=1}^{\bar{N}} \left(1 + \frac{\mathrm{SNR}_i}{\Gamma}\right) \right)^{1/\bar{N}} - 1 \right], \tag{20}$$

where SNR_i is the SNR on the *i*th subchannel, and Γ is the SNR gap to capacity [12]

In the simulations, the DMT VDSL signal is assumed to occupy 0-25.6 MHz with 256 subchannels in an FDM design. The symbol rate for each VDSL subchannel is 100 kilosymbols-per-second. A rate-1/2 [23, 35] convolutional code with constraint length 5 is used for channel coding. The number of coded bits per data frame is set at 1024, indicating an average bit rate of 512 bits per DMT block or 2 bits/s/Hz. A random interleaver of length 1024 is used for interleaving and de-interleaving. The coded DMT system is applied to a channel with severe ISI, the transfer function of which is taken from [9] to be

$$H(\omega) = 1 + 0.9\cos(\omega T). \tag{21}$$

This channel is not necessarily typical of a VDSL line transfer function, however it serves as a useful example for illustration purposes. Campello's margin-adaptive bit-loading algorithm [2] is used to allocate bits to DMT subchannels. We assume a square QAM constellation for simplicity, so the granularity of bit loading equals 2. Also the first two tones (up to 200 KHz) are not used for compatibility with POTS/ISDN service. We see that typical constellations are 64-, 16- and 4-QAM. Subchannels not able to transmit 2 bits reliably are not used.

We assume one NEXT crosstalk signal with a known coupling function. The crosstalk signal is binary phase-shift keying (BPSK) modulated, carried on a 12.8 MHz central frequency with a 0.8M symbol-per-second rate. Such a situation would arise, for example, due to the coexistence of HomePhone LANs and asymmetric DMT VDSL signals in the same cable in the customer premises [4]. Thus, there are 2^8 possible crosstalk sequences in one VDSL symbol. This number is chosen for simulation simplicity. In reality, this number could be much larger. The average PSD levels of the crosstalk signal and background noise floor are fixed while that of the desired signal is varied, corresponding to different line lengths (the signal attenuation is increasing with the line length). In our simulation, the average PSD of the crosstalk is 30 dB higher than that of the background noise floor, and the peak PSD of the crosstalk is 48 dB higher. These settings seem to agree roughly with empirical results [12].

To get an overall idea of how much gain is obtained through turbo multiuser detection in coded DMT systems, we compare the performance of the proposed schemes with their SUD, uncoded and non-iterative counterparts in Fig. 3. Here, SUD means application of traditional demodulation to an uncoded DMT system, while IC-MUD and ML-MUD mean application of the corresponding multiuser detection scheme on an uncoded system. Clearly, MUD greatly outperforms SUD. For BER greater than 10^{-3}, IC-MUD performs identically with ML-MUD. But when the desired signal level increases, the performance of IC diminishes. (It is well known that IC performs worst when the powers of the different users are equal or comparable.) IC-MUD+VA and ML-MUD+VA refer to detection of a coded DMT VDSL signal in a non-iterative way: after multiuser detection, hard decisions are made on coded bits, then the Viterbi algorithm is used for decoding. The coded DMT system has a smaller $d_{\min}$ compared with its uncoded counterpart for the same $\frac{E_b}{N_0}$, but the coding gain more than offsets this disadvantage. We see that at BER 10^{-7}, ML-MUD+VA provides 2.5 dB gain over ML-MUD. Again, the IC-MUD+VA scheme deteriorates for higher DMT signal levels. The ML-MUD+MAP and IC-MUD+SOVA denote the performance of our turbo iterative algorithms, which adopt ML-MUD and IC-MUD for multiuser detection, and adopt MAP [11] and SOVA [7] for channel decoding respectively, at the fifth iteration. At BER 10^{-7}, we see that an additional 4.5 dB gain is achieved over ML-MUD+VA.

Finally, we would like to examine the performance of this proposed iterative DMT receiver with impulse noise. The impulse noise is assumed to have parameters $\varepsilon = 0.01$ and $\kappa = 100$ (see (5)), which means the impulse spike is 20 dB higher than the background noise floor with occurrence probabillity of 1% per frequency bin. We do not include crosstalk signals, and MAP is used in the SISO decoder. Figure 4 shows that the performance of the proposed receiver is greatly degraded with impulse noise. The use of erasure decoding can remedy this. In the demodulation stage, for those bits associated with impulse-contaminated symbols, instead of calculating soft metrics for them, the *a priori* information is used as a substitute, i.e., $\lambda_1(b_{k,i}) = \lambda_2^p(b_{k,i})$. (For the first iteration, these are set to zero.) In DMT systems, the erasure positions, where impulse spikes appear, can possibly be detected in advance through pilot tones. In Fig. 5, we see that the performance of the proposed receiver experiences almost no performance loss with impulse noise with the aid of erasure decoding. The reader is referred to [4] for an alternative approach for combating impulse noise when crosstalk signals are present.

VI CONCLUSIONS

In this paper, a new coded DMT VDSL receiver structure using the idea of turbo multiuser detection is proposed and is shown to achieve an overall 7.0 dB gain over the uncoded optimum receiver at BER 10^{-7} for a channel with severe ISI, AWGN, and one dominant crosstalk signal. The effect of impulse noise is detrimental to the proposed scheme but can be overcome through an erasure decoding technique.

In this paper, we have assumed knowledge of the line transfer function and crosstalk coupling functions. In reality, however, channel identification is needed, and the effects of channel estimation error should be taken into consideration. These issues are of interest for further study. The problem of detecting impulse spike positions also deserves further study.

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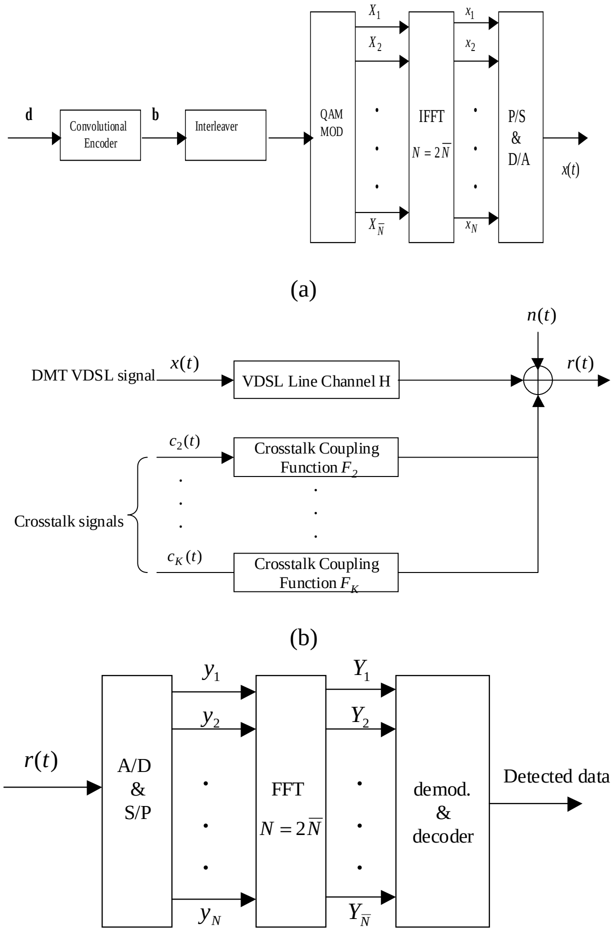

(a)

(b)

(c)

Fig. 1 VDSL DMT System Configuration: (a) Transmitter; (b) Channel; (c) Receiver

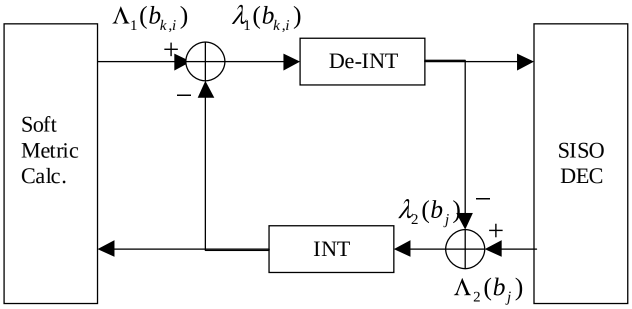

Fig. 2 Turbo structure for iterative demodulation and decoding

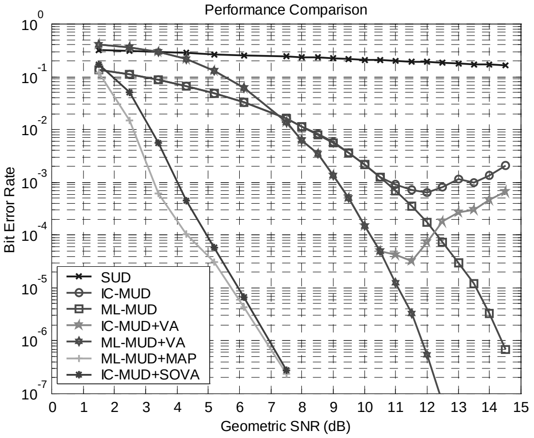

Fig. 3 Performance comparison of various DMT VDSL receivers

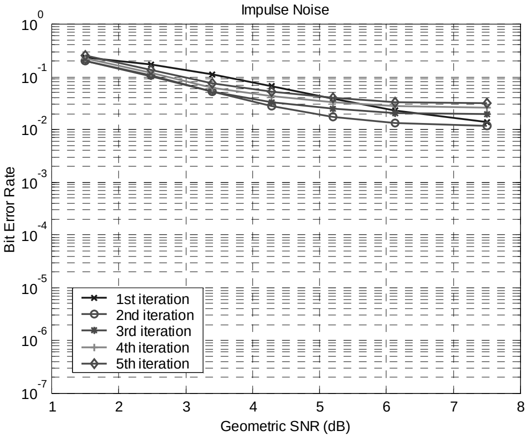

Fig. 4 Performance of the iterative DMT receiver with impulse noise

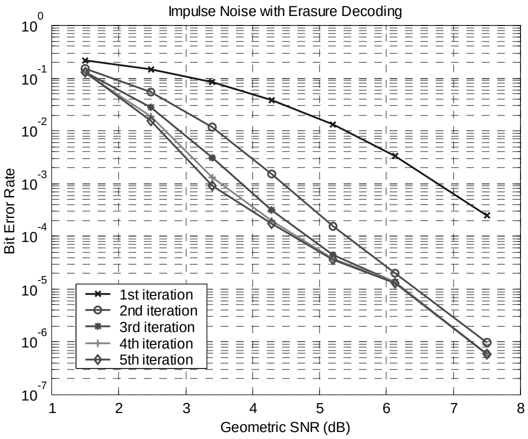

Fig. 5 Performance of the iterative DMT receiver with erasure decoding with impulse noise