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ON THE MMSE ITERATIVE EQUALIZATION FOR TDMA PACKET SYSTEMS

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Abstract – Original turbo equalization using a trellis-based channel equalizer and channel decoder improves significantly the bit error rate performance. However, a large alphabet modulation employed in the systems with multipath channels requires an excessive high number of states in such equalizer, so the optimal maximum *a posteriori* probability (MAP) becomes prohibitively complex. Therefore, sub-optimum equalizers with *a priori* information from the channel decoder have to be considered in order to enhance its performance. In this paper we investigate the performances of minimum mean square error (MMSE) filter based iterative equalization for the Enhanced General Packet Radio Service (EGPRS) radio link. The simulation results demonstrate that MMSE turbo equalization constitutes an attractive candidate for single-carrier wireless transmissions with multilevel modulation, in long delay-spread environments.

Keywords: turbo equalization, MMSE filter, TDMA systems

1. INTRODUCTION

Turbo-equalization [1], [2] is a powerful mean to perform joint equalization and decoding, when considering coded data transmission over time dispersive channels. The association of the code and the discrete-time equivalent channel (separated by an interleaver) is seen as the serial concatenation of two codes. The turbo principle, the iterative exchange of extrinsic information between a soft-in/soft-out (SISO) equalizer and a SISO decoder, may then be used at the receiver for improve its performances.

Classically, these SISO modules are implemented using conventional a posteriori probability (APP) algorithms [3]. This leads to a complexity which evolves in $O(M^L)$ for the equalization; M, L being respectively the modulation alphabet size and the discrete-time equivalent channel length. Obviously, the optimal equalization process needed at the receiver becomes rapidly untractable for long impulse response channels and for high constellations order.

In this paper we study the performances of low complexity SISO equalizer, based on minimum mean square error (MMSE) equalization, proposed in [4] and [5], in context of packet transmission TDMA systems.

A joint design of the equalization and demodulation parts, based on a gaussian assumption, allows producing good estimates of the symbol extrinsic probabilities. Special care is then taken to the computation of the bit extrinsic log-likelihood ratios (LLRs), in order to fully exploit the mutual information between the bits associated with a given complex symbol, capitalizing on methods presented in [6], [7], [8].

2. SYSTEM DESCRIPTION

A. Signal model

The transmission scheme is represented in Fig. 1. A frame of information bits b_k is encoded by a rate-r convolutional encoder. The resulting encoded bits c_m are interleaved using a random permutation function to give the interleaved coded bits x_m. The q bits $x_n^p = x_{(n-1)q+p}$ $(p = 1,...,q)$ are grouped and mapped to a complex symbol d_n, among the $M = 2^q$ possible symbols of the considered constellation. The resulting complex symbols are transmitted over the channel, which is assumed static over a frame and perfectly known. At the receiver, we assume matched filtering to the whole transmission chain, symbol-rate sampling and discrete-time noise whitening. Thus, the channel may be represented by its equivalent discrete-time white noise filter model, i.e. a causal discrete-time filter with coefficients h_j $(j = 0,...,L)$ corrupted by white gaussian noise samples w_n of variance s_w^2.

The symbols r_n at the output of the channel may thus be expressed as

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$$r_n = \sum_{l=0}^{L} h_l d_{n-l} + w_n \qquad (1)$$

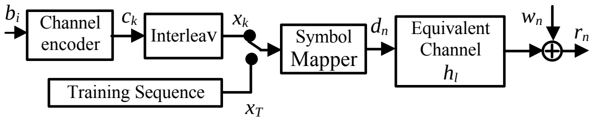

Fig. 1. Transmitter scheme

B. Iterative receiver

A global view of the proposed receiver scheme is given in Fig. 2. Seen at this level of generality, it is similar to a classical turbo-equalizer [1]. It consists of two stages: a SISO equalizer/demapper and a SISO decoder separated by bit-deinterleavers and bit-interleavers. Those two stages exchange extrinsic information, on an iterative fashion, in order to improve the performances. The decoder is implemented using an optimal APP algorithm. We focus in this paper on the presentation of the proposed equalizer/demapper. Note that the a priori inputs and extrinsic outputs of the equalizer/demapper are bit LLRs. It has thus to deal with all the bit/symbol conversion aspects associated with the considered modulation.

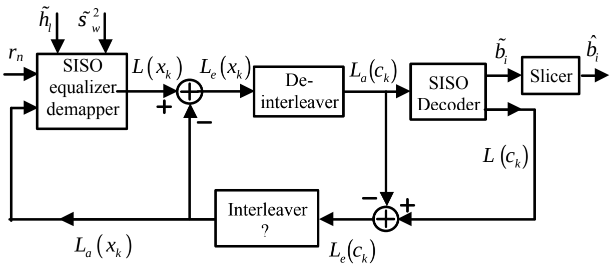

Fig. 2. Turbo equalization receiver

3. SISO EQUALIZER

A. General description

A more detailed scheme of the proposed equalizer is given in Fig. 3. Using the bit a priori LLRs $L_a(x_m)$ produced by the decoder, it begins by computing the first and second order statistics of the symbols d_n. Using those statistics and the received samples r_n, a symbol equalizer produces estimates $\hat{d}_n$ in order to minimize $E\left\{\left|d_n - \hat{d}_n\right|^2\right\}$. The M corresponding symbol extrinsic probabilities are then approximated as $\Pr\left(\hat{d}_n | d_n\right)$ using an equivalent gaussian channel assumption at the output of the symbol equalizer. The parameters of this equivalent channel are calculated for each estimate $\hat{d}_n$ on the basis of the equalizer structure and of the symbols statistics. Based on the obtained symbol extrinsic probabilities, we finally evaluate, for $p = 1,...,q$, the bit extrinsic probabilities $\Pr_e\left(x_i^p\right)$ using the a priori information about the other bits $x_n^r \ (r \neq p)$ associated with the considered symbol d_n, and output an extrinsic LLR $L_e\left(x_n^p\right)$. To keep the analogy with an optimal APP equalizer, we only use the a priori information available about symbols d_i with $i \neq n$ when computing the estimate $\hat{d}_n$ and the symbol extrinsic probabilities $\Pr\left(\hat{d}_n | d_n\right)$.

B. Symbols statistics using the a priori information

We first calculate an estimation of the mean and variance of each symbol on the basis of the a priori information available. Noting S the set of all possible symbols, with $\text{card}(\mathcal{S}) = M$, for each transmitted symbol d_n, we compute first the symbols a priori probabilities $\Pr_a\left(d_n = s_j\right) \ \forall s_j \in \mathcal{S}$. Assuming independence between the interleaved coded bits and basing on the corresponding bits a priori probabilities this probability can be written as:

$$\Pr_{a,n}\left(s_j\right) = \Pr_a\left(d_n = s_j\right) = \prod_{p=1,...,q} \Pr_a\left(x_n^p\right) \qquad (2)$$

where the q bits $x_n^p \ \left(p = 1,...,q\right)$ takes values in $\{0,1\}$ as a function of the considered symbols s_j. The probabilities $\Pr_a\left(x_n^p\right)$ are classically calculated from $L_a\left(x_n^p\right) = \ln \dfrac{\Pr_a\left(x_n^p = 1\right)}{\Pr_a\left(x_n^p = 0\right)}$, the a priori LLR of the bit x_n^p. The mean value $\bar{d}_n$ of symbol d_n on the basis of the a priori information at time n is then:

$$\bar{d}_n = E\left\{d_n \middle| L_{a,n}\right\} = \sum_{s_j \in \mathcal{S}} s_j \times \Pr_{a,n}\left(s_j\right) \qquad (3)$$

Similarly, the variance u_n^2 of symbol at time *n,* on the basis of the a priori information, is:

$$u_n^2 = E\left\{\left|d_n - \bar{d}_n\right|^2 \middle| L_{a,n}\right\} = \sum_{s_j \in \mathcal{S}} |s|^2 \times \Pr_{a,n}\left(s_j\right) - \left|\bar{d}_n\right|^2 \qquad (4)$$

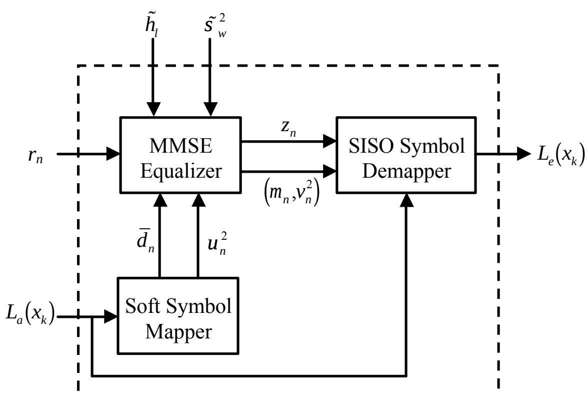

Fig. 3. Block diagram of generic
SISO MMSE equalizer with demapper

The expression of u_n^2 depends on the considered modulation. If $|s| = \text{const} \quad \forall s \in \mathcal{S}$ (i.e. M-PSK), the expectation $E\left\{|d_n|^2\right\}$ is constant and equals s_d^2, that is the variance of a symbol without a priori information. If not (i.e. M-QAM), it has to be calculated explicitly as in (4).

C. Symbol equalizer

Defining the equalizer length as $N = N_1 + N_2 + 1$, we first introduce a sliding-window model using the vectors

$$\mathbf{r}_n = \left[r_{n-N_1} \ldots r_n \ldots r_{n+N_2} \right]^T \qquad (5.a)$$

$$\mathbf{d}_n = \left[d_{n-N_1-L} \ldots d_n \ldots d_{n+N_2} \right]^T \qquad (5.b)$$

$$\mathbf{w}_n = \left[w_{n-N_1} \ldots w_n \ldots w_{n+N_2} \right]^T \qquad (5.c)$$

and the $\left(N \times (N+L)\right)$-channel matrix

$$\mathbf{H} = \begin{bmatrix} h_L & \cdots & h_0 & 0 & \cdots & \cdots & 0 \\ 0 & h_L & \cdots & h_0 & 0 & \cdots & 0 \\ \vdots & \ddots & \ddots & \ddots & \ddots & \ddots & \vdots \\ 0 & \cdots & \cdots & 0 & h_L & \cdots & h_0 \end{bmatrix} \qquad (6)$$

At each time step *n*, we may then write:

$$\mathbf{r}_n = \mathbf{H}\mathbf{d}_n + \mathbf{w}_n \qquad (7)$$

where $\mathbf{w}_n$ is a complex gaussian noise vector, i.e. $\mathbf{w}_n \sim \mathcal{N}\left(\mathbf{0}, s_w^2 \mathbf{I}\right)$; $\mathbf{I}$ being the $N \times N$ identity matrix. Considering this sliding-window channel model, we present a low-complexity equalizer [5] in order to produce estimates $\hat{d}_n$, assuming the knowledge of the noise and symbols first and second order statistics.

An MMSE estimator may be expressed in this context as:

$$\hat{d}_n = E\{d_n\} + \mathbf{p}_n^H \left[\mathbf{r}_n - E\{\mathbf{r}_n\} \right] \qquad (8)$$

with the length-*N* complex vector $\mathbf{p}_n$ given by:

$$\mathbf{p}_n = \text{cov}\{\mathbf{r}_n, \mathbf{r}_n\}^{-1} \cdot \text{cov}\{\mathbf{r}_n, \mathbf{d}_n\} \qquad (9)$$

where ^{H} denotes the conjugate transpose operator and $\text{cov}\{\mathbf{x}, \mathbf{y}\} = E\left\{\left[\mathbf{x} - E\{\mathbf{x}\}\right]\left[\mathbf{y} - E\{\mathbf{y}\}\right]^H\right\}$

In conformity with turbodecoding and turboequalization principle [5],[6], the a priori information about symbol d_n should not be used in the evaluation of its estimate $\hat{d}_n$. In other words, at time *n*, for symbols d_i with $i \neq n$, we can use the mean $\bar{d}_i$ and the variance u_i^2 computed on the basis of the a priori information in (2) and (3). On the contrary, the mean and variance of symbol d_n is computed without using the corresponding a priori information, which leads to 0 and s_d^2 respectively. The expectation in (8) can be computed as follow:

$$E\{d_n\} = 0 \text{ and } E\{d_n\} = E\{\mathbf{H}\mathbf{d}_n + \mathbf{w}_n\} = \mathbf{H}\bar{\mathbf{d}}_n \qquad (10)$$

where we have defined $\bar{\mathbf{d}}_n = E\{\mathbf{d}_n\}$ as:

$$\bar{\mathbf{d}}_n = \left[\bar{d}_{n-N_1-L} \ldots \bar{d}_{n-1} \; 0 \; \bar{d}_{n+1} \ldots \bar{d}_{n+N_2} \right]^T \qquad (11)$$

The first factor in (9) may be calculated as follows:

$$\begin{aligned} \text{cov}\{\mathbf{r}_n, \mathbf{r}_n\} &= \mathbf{H} \cdot \text{cov}\{\mathbf{d}_n, \mathbf{d}_n\} \cdot \mathbf{H}^H + s_w^2 \mathbf{I} \\ &= \mathbf{H}\mathbf{R}_{dd,n}\mathbf{H}^H + s_w^2 \mathbf{I} \end{aligned} \qquad (12)$$

where

$$\mathbf{R}_{dd,n} = \text{cov}\{\mathbf{d}_n, \mathbf{d}_n\} = E\left\{\left[\mathbf{d}_n - \bar{\mathbf{d}}_n\right]\left[\mathbf{d}_n - \bar{\mathbf{d}}_n\right]^H\right\} \qquad (13)$$

Using the definition given in (3) and once again avoiding using the a priori information available about symbol d_n at time step *n*, this matrix can be expressed as:

$$\begin{aligned} \mathbf{R}_{dd,n} = \text{diag}[\, &\text{var}\left(d_{n-N_1-L}\right) \ldots \text{var}\left(d_{n-1}\right) \; s_d^2 \\ &\text{var}\left(d_{n+1}\right) \ldots \text{var}\left(d_{n+N_2}\right)] \end{aligned} \qquad (14)$$

where we used the independence assumption between the coded bits, so that $\text{cov}\left(d_n, d_i\right) = 0$ for $n \neq i$

The second factor in (9) may be calculated as follows:

$$\begin{aligned} \text{cov}\{\mathbf{r}_n, \mathbf{d}_n\} &= \mathbf{H}\text{cov}\left(\mathbf{d}_n, \mathbf{d}_n\right) = \mathbf{H}E\left\{\left[\mathbf{d}_n - \bar{\mathbf{d}}_n\right] d_n^*\right\} \\ &= \mathbf{H}\,\text{cov}\left(\mathbf{d}_n, d_n^*\right) = \mathbf{H}\mathbf{e}s_d^2 \end{aligned} \qquad (15)$$

where $\mathbf{e}$ denotes a length-(N+L) vector of all zeros, except for the (N_1+L+1)th element, which is 1, and $\mathbf{h} = \mathbf{H}\mathbf{e}$.

Using (9), (12) and (15), the complex vector $\mathbf{p}_n$, which is seen as a time-varying equalization filter, becomes:

$$\mathbf{p}_n = s_d^2 \left[\mathbf{H}\mathbf{R}_{dd,n}\mathbf{H}^H + s_w^2 \mathbf{I} \right]^{-1} \mathbf{h} \qquad (16)$$

Finally, using (8), (10) and (16), we obtain the following expression of the estimate symbol $\hat{d}_n$:

$$\begin{aligned} \hat{d}_n &= \mathbf{p}_n^H \left[\mathbf{r}_n - \mathbf{H}\bar{\mathbf{d}}_i \right] = \\ &= s_d^2 \mathbf{h}^H \left[\mathbf{H}\mathbf{R}_{dd,n}\mathbf{H}^H + s_w^2 \mathbf{I} \right]^{-1} \left[\mathbf{r}_n - \mathbf{H}\bar{\mathbf{d}}_n \right] \end{aligned} \qquad (17)$$

The generalized MMSE equalizer reduces to classical MMSE equalization for the first iteration of the iterative process when a priori information is not available. This scheme may also be seen as an improved interference canceler taking the statistical nature of the soft values into account. It reduces to a classical soft-interference canceler [9], when perfect a priori information is available.

D. Equivalent AWGN channel assumption

At the output of the equalizer, in order to be able to demodulate the symbols, we assume that the estimate $\hat{d}_n$ is the output of an equivalent AWGN channel having d_n as its input:

$$\hat{d}_n = m_n d_n + h_n \qquad (18)$$

m_n is the equivalent amplitude of the signal at the output and h_n is a complex white gaussian noise with zero mean and variance n_n^2. This is equivalent to say that the estimates are complex gaussian distributed, i.e. $\hat{d}_n \sim \mathcal{N}_c\left(m_n d_n, n_n^2\right)$. The parameters m_n and n_n^2 are calculated at each time step *n* as a function of the equalizer structure, and thus also of the symbols statistics.

The mean m_n is calculated by first evaluating:

$$E\left\{\hat{d}_n, d_n^*\right\} = m_n E\left\{\left|d_n\right|^2\right\} = m_n s_d^2 \quad (19)$$

which can be expressed as:

$$E\left\{\hat{d}_n, d_n^*\right\} = \mathbf{p}_n^H \mathbf{h} s_d^2 \quad (20)$$

and, from (16) and (20), we finally obtain:

$$m_n = \mathbf{p}_n^H \mathbf{h} = \mathbf{h}^H \left[\mathbf{H}\mathbf{R}_{dd,n}\mathbf{H}^H + s_w^2 \mathbf{I}\right]^{-1} \mathbf{h} s_d^2 \quad (21)$$

The variance n_n^2 may be expressed as:

$$n_n^2 = E\left\{\left|h_n\right|^2\right\} = E\left\{\left|\hat{d}_n - m_n d_n\right|^2\right\} = E\left\{\left|\hat{d}_n\right|^2\right\} - m_n^2 s_d^2 \quad (22)$$

We have thus:

$$n_n^2 = \mathbf{p}_n^H \left[\mathbf{H}\mathbf{R}_{dd,n}\mathbf{H}^H + s_w^2 \mathbf{I}\right]\mathbf{p}_n - m_n^2 s_d^2 \quad (23)$$

Using (16), (21) and (23), we find the following expression:

$$n_n^2 = m_n s_d^2 - m_n^2 s_d^2 \quad (24)$$

E. Symbol extrinsic probabilities computation

In order to compute the bit extrinsic probabilities, we have first to approximate the symbol extrinsic probabilities. We use therefore the gaussian equivalent channel assumption given in (18) and estimate the symbol posterior probabilities $\Pr\left(d_i\right)$ as:

$$\Pr\left(d_n\right) = \Pr\left(d_n \left|\mathbf{r}\right.\right) \approx \Pr\left(d_n \left|\hat{\mathbf{d}}\right.\right) \approx \Pr\left(d_n \left|\hat{d}_n\right.\right) = \frac{\Pr\left(d_n \left|\hat{d}_n\right.\right) \cdot \Pr_a\left(d_n\right)}{\Pr\left(\hat{d}_n\right)} \quad (25)$$

where $\mathbf{r}$ and $\hat{\mathbf{d}}$ are the sequences of the received symbols and of the estimates respectively and we used the Bayes rule and the equivalent AWGN channel for the frequency selective channel. From (25), the symbol extrinsic probabilities $\Pr_e\left(d_n\right)$ may then be written as follows:

$$\Pr_e\left(d_n\right) = \mathcal{K}_d \frac{\Pr\left(d_n\right)}{\Pr_a\left(d_n\right)} \sim \frac{\Pr\left(d_n \left|\hat{d}_n\right.\right)}{\Pr\left(\hat{d}_n\right)} \sim \Pr\left(\hat{d}_n \left|d_n\right.\right) \quad (26)$$

where $\mathcal{K}_d$ is a normalization constant and where the last equivalence is obtained when omitting the terms common to all hypotheses. Using the parameters m_n and n_n^2 of the equivalent AWGN channel computed for the estimate $\hat{d}_n$, the *M* symbol extrinsic probabilities at time *n* may finally be approximated as follows:

$$\Pr\left(\hat{d}_n \left|d_n\right.\right) = \frac{1}{n_n^2 p} \exp\left(-\frac{\left|\hat{d}_n - m_n d_n\right|^2}{n_n^2}\right) \quad (27)$$

Note that these probabilities were obtained without using the a priori information available about symbol d_n, which is consequent with the optimal algorithm. Note also, that the equivalent channel assumption allowed taking the symbols statistics into account

F. Bit extrinsic LLR computation

The extrinsic probabilities of a given coded bit x_n^p may be expressed as a function of the symbol extrinsic probabilities $\Pr_e\left(d_n\right)$ [7], as follows:

$$\Pr_e\left(x_n^p\right) = \mathcal{K}_x \frac{\Pr\left(x_n^p\right)}{\Pr_a\left(x_n^p\right)} \approx \sum_{s_i : x_n^p} \Pr_e\left(d_n\right) \left[\prod_{\substack{r=1,\dots,q \\ r \neq p}} \Pr_a\left(x_n^r\right)\right] \quad (28)$$

where $\mathcal{K}_x$ is a normalization constant, $\Pr\left(x_n^p\right) = \Pr\left(x_n^p \left|\mathbf{r}\right.\right)$ are the posterior probabilities of the bit x_n^p and the notation $s_i : x_n^p$ represents the subset of the symbols $s_i \in \mathcal{S}$ with a given value of x^p. $\Pr_e\left(d_n\right) = \Pr_e\left(d_n = s_i\right)$ denotes the extrinsec probability that the emitted symbol at time *n* is the symbol s_i from $\mathcal{S}$ set of possible symbols. Equation (28) allows taking the a priori probabilities of the other bits x_n^r $\left(r = 1,\dots,q\ ;\ r \neq p\right)$ associated with the considered symbol d_n in order to evaluate the extrinsic probabilities of the bit x_n^p. So, the mutual influence between encoded bits is used for a better demapping. Using (27) and (28), we can then approximate $\Pr_e\left(x_n^p\right)$ as:

$$\Pr_e\left(x_n^p\right) \approx \sum_{s_i : x_n^p} \Pr_e\left(\hat{d}_n \left|d_n\right.\right) \left[\prod_{\substack{r=1,\dots,q \\ r \neq p}} \Pr_a\left(x_n^r\right)\right] \quad (29)$$

The finally form of bit extrisec LLR $L_e\left(x_n^p\right) = \ln \frac{\Pr_e\left(x_n^p = 1\right)}{\Pr_e\left(x_n^p = 0\right)}$ at the output of the equalizer is:

$$L_e\left(x_n^p\right) \approx \ln \frac{\sum_{s_i : x_n^p = 1} \Pr\left(\hat{d}_n \left|d_n\right.\right) \left[\prod_{\substack{r=1,\dots,q \\ r \neq p}} \Pr_a\left(x_n^r\right)\right]}{\sum_{s_i : x_n^p = 0} \Pr\left(\hat{d}_n \left|d_n\right.\right) \left[\prod_{\substack{r=1,\dots,q \\ r \neq p}} \Pr_a\left(x_n^r\right)\right]} \quad (30)$$

As shown in [6], we can obtain a more robust implementation in the logarithmic domain, using the well-known generalized maximum function and taking the considered constellation into account to get further simplifications

G. Asymptotic performances

In the iterative process at the receiver, asymptotic performances are reached when perfect a priori information is available at the equalizer/demapper. In this case, it can be shown that the proposed scheme manages to totally suppress ISI and reaches the matched filter bound (MFB). The only difference remaining while using the equivalent AWGN channel model at the output of the equalizer is the noise correlation. However, this noise correlation is not taken into account in the demodulation process and, from the point of view of the decoder, it is broken due to the presence of the deinterleaver. The asymptotic performances of the scheme are thus logically identical to those of iterative demodulation and decoding on an AWGN channel [7]. They can be obtained by simulation considering iterative demodulation and decoding on an AWGN channel with perfect a priori information at the demapper, considering the same code and the same mapping. There is no optimal solution; it depends on the considered E_b / N_0, code, mapping, channel and on the number of iterations allowed at the receiver

4. SIMULATION RESULTS

Simulations of the MMSE turbo equalization were performed in the Enhanced Data for GSM Evolution (EDGE) radio access scheme. The burst structure is described in [10]. A burst carries 2x58 playload data symbols, includes a middle training sequence of 26 symbols and 8.25 guard symbols at the end. We performed the simulation for MCS-5 coding scheme [10], which employs 8-PSK modulation. So 3 interleaved encoded bits are mapped, using Gray labelling, into one burst symbol. For the sake of simplicity, the performance results will be examined only over user data, which are encoded by a rate 1/3 non-recursive non-systematic convolutional encoder with constraint length 7 and octal generator polynomials 133,171,145. In MCS-5 coding scheme, to obtain a user data code rate R=0,37 [10], the CPS=20 puncturing pattern are used. The coded punctured user data are interleaved with deterministic interleaver from EDGE technical recommendation, combined with the header part and the flags, forming a block of 1392 bits. This bits are partitioned over 4 data blocks of 348 bits, i.e. 2x58 8-PSK symbols, which are finally mapped onto 4 different bursts. We evaluated the bit error rate (BER) for the transmission over channel A, a channel with 11 taps [0.04, -0.05, 0.07, -0.21, -0.5, 0.72, 0.36, 0.21, 0.03, 0.07], and over channel B, a channel of length five [2-0.4j, 1.5+1.8j, 1, 1.2-1.3j, 0.8+1.6j]. The real or complex path gains were normalized such that $\sum_{l=0}^{L-1} |h_l|^2 = 1$. In all simulations, we totalized 50 frame errors for each E_s / N_0 and performed six iterations. The reference curve, represented by a dashed line in both figures, corresponds to the performance of the coded transmission scheme over an ISI-free AWGN channel. In figure 4 are plotted the receiver performance (BER after decoding) for transmission over channel A. The filter length is 10, N_1=0 and N_2=10. We can see that, for medium and high SNR, the receiver improves its performance from an iteration to another and after five-six iterations it achieves the ISI free transmission performance, for E_s / N_0 greater than three.

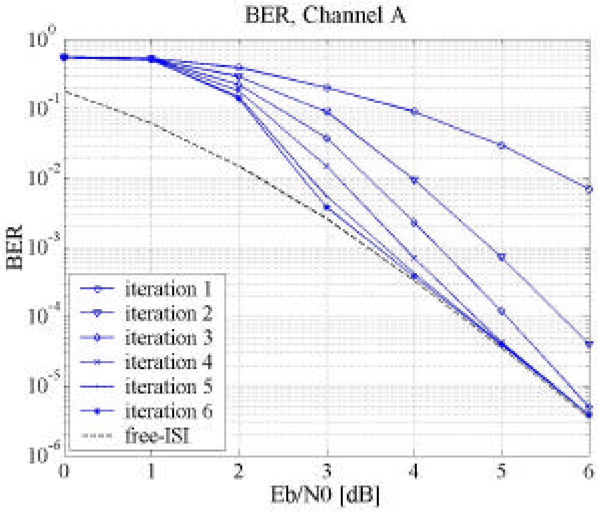

Fig. 4. BER-curves for MCS-5 over channel A.

In figure 5 the results of the transmision over channel B and N_1=3 , N_2=7 are reproduced. Because this channel is harder to equalize, the receiver improves its performances by iteration technique for higher threshold of E_s / N_0. We can draw the same conclusion: the ISI is totaly supressed after an iteration number, depending on amount of E_s / N_0 for the medium and high SNR.

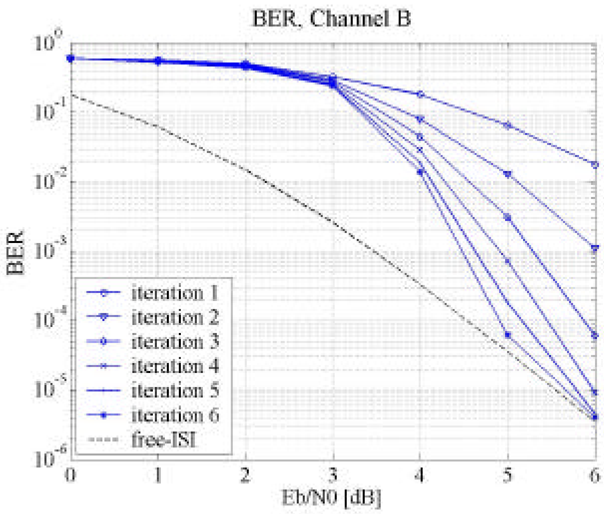

Fig. 5. BER-curves for MCS-5 over channel B.

This turboequalization scheme preserve the performance previous reported in [5], even in a dereministic interleaver context of EDGE system.

5. CONCLUSIONS

The 8-PSK modulation scheme used in EDGE requires a sub-optimum equalizer. The MMSE filter based equalizers has the lowest complexity, which is independent of the size of the symbol alphabet. Their moderate performance can be increased using iterative principle in so called turbo-equalization scheme.

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