Improving MIMO Detection by L-Value Analysis and Adaptive Threshold-Based Cancellation

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***Abstract*— We present an improved algorithm for detection of the received signal of a convolutionally encoded wireless MIMO system. The method is based on BLAST (Bell-Labs Layered Space-Time) algorithm with successive interference cancellation. Compared to existing solutions, cancellation is done adaptively. The receiver uses an MAP soft output demapper combined with an iterative (Turbo-) decoder. Adaptive symbol cancellation is done using a threshold which takes reliability of a symbol decision into account. Cancellations of wrongly decided symbols are reduced to a large extend. As a result, we obtain an SNR gain of 1 to 1.5 dB for Rayleigh fading MIMO channels with additive white Gaussian noise.**

I. Introduction

In the initial publications [1], [2] the BLAST (Bell-Labs Layered Space-Time) algorithm was mainly presented as zero-forcing (ZF) with ordered successive interference cancellation (OSIC). The minimum mean square error (MMSE) solution was mentioned there, but presented in detail e. g. in [3]. It is well known that OSIC reduces the bit error ratio (BER) of an *uncoded* BLAST system [3]. However, the behaviour of a *coded* BLAST system is quite different. A simple and straightforward implementation of a coded BLAST transmitter is shown in Fig. 1. The binary output of the data source (information bits) $u(t_u)$ is encoded to yield $x'(t_x)$ (coded bits), interleaved resulting in $x(t_x)$, and mapped to $\mathbf{s}^{\mathrm{T}}(t_s)$ so that each of the M transmitters carries an independent data stream $s_i(t_s)$ where the index i denotes the number of the transmitter ($i = 1, \ldots, M$). t_u, t_x, and t_s are the discrete time instances of information bits, coded bits, and symbols, respectively, which are dropped in the following to simplify the notation. A random interleaver of size $S = 96000$ and a systematic convolutional (SC) code of rate $R_c = 1/2$, memory $\nu = 2$, and feedforward polynomial $G_{\text{oct}} = 7$ are used. As usual, the polynomial is given in octal numbers. This structure was called *vertical coding* [4].

A convenient way to describe a flat fading MIMO model is the use of the matrix notation

$$\mathbf{r} = H\mathbf{s} + \mathbf{n} \tag{1}$$

where $\mathbf{s} = (s_1, \ldots, s_M)^{\mathrm{T}}$ is the transmit symbol column vector. Each component s_i is a complex QAM symbol to be sent by antenna i ($i = 1, 2, \ldots, M$). $\mathbf{n} = (n_1, \ldots, n_N)^{\mathrm{T}}$ is an additive noise column vector with components n_j, which are complex AWGN at receive antenna j, each with zero mean and variance $2\sigma_{\mathrm{n}}^2$ ($j = 1, 2, \ldots, N$). This means that real and imaginary part of n_j are both Gaussian with variance $\sigma_{\mathrm{n}}^2 = N_0/2$. N is the number of receive antennas. We consider $\mathbf{n}$ to be uncorrelated with expected value $\mathsf{E}\{\mathbf{n}\mathbf{n}^*\} = 2\sigma_{\mathrm{n}}^2 I_N$. I_N is the identity matrix of size N. Furthermore, it is assumed that the vectors $\mathbf{s}$ and $\mathbf{n}$ are uncorrelated, i. e. $\mathsf{E}\{\mathbf{s}\mathbf{n}^*\} = 0$. $\mathbf{r} = (r_1, \ldots, r_N)^{\mathrm{T}}$ is the received symbol column vector, where r_j is received by antenna j ($j = 1, 2, \ldots, N$). The channel matrix H is given by $H = (\mathbf{h}_1 \ldots \mathbf{h}_M)$ with $\mathbf{h}_i = (h_{1i}, \ldots, h_{Ni})^{\mathrm{T}}$ is a column vector of H. The impulse response h_{ji} from transmitter i to receiver j is modeled as a zero-mean, complex Gaussian random variable satisfying $\mathsf{E}\{|h_{ij}|^2\} = 1$ (i. e. the channel is passive). All entries of H are i.i.d. $(\cdot)^{\mathrm{T}}$ denotes the transpose and $(\cdot)^*$ the conjugate transpose. If not stated otherwise, bold characters are used for vectors in our text. (Complex) scalars are denoted by small and matrices by capital letters.

Assuming that we use the same mapping for all transmit antennas and mean energy per symbol E_s, the averaged total transmit power equals ME_s. We define the signal to noise ratio (SNR) as $E_b/N_0 = \frac{E_s \cdot N}{R_c \cdot Q \cdot 2\sigma_{\mathrm{n}}^2}$ where $\frac{E_s \cdot N}{R_c \cdot Q}$ is the transmitted energy per information bit at the receiver. Using the column vectors of H we can write Eq. (1) as

$$\mathbf{r} = \sum_{i=1}^{M} \mathbf{h}_i s_i + \mathbf{n} \tag{2}$$

The BLAST system needs to have at least as much receive antennas N as transmit antennas M. While the transmitter does not need to have any knowledge about the channel, we assume that the receiver has full knowledge of H.

The paper is organized as follows. In Section II we describe the receiver. The new detection scheme is presented in Section III. In Section IV simulation results are given and discussed follows by complexity considerations in Section V. Section VI concludes this paper.

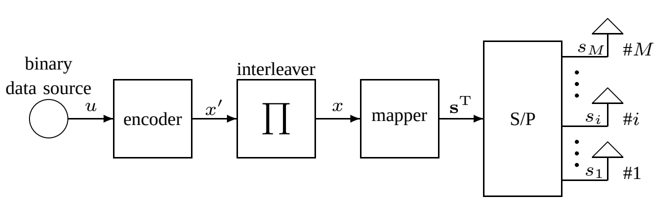

Fig. 1: Transmitter

II. The Receiver

Fig. 2 shows the block diagram of the receiver. The MIMO detector takes the N received samples r_j and performs the so called minimum mean squared error ordered successive interference cancellation (MMSE-OSIC) or just the MMSE algorithm. The latter operates without symbol cancellation [1]. Alternatively, the detector performs ZF (zero forcing) or ZF-OSIC. Further, in the following we will introduce adaptive cancellation to improve BER. As explained later, in our solution the MAP soft-output demapper is part of the detector rather than a separate block after detection.

Assume, that the detection order is given by the ordered sequence $\mathcal{D} = \{d_1, d_2, \ldots, d_M\}$. i. e. symbol s_{d_1} is detected first out of the received vector $\mathbf{r}$, then s_{d_2} etc. until s_{d_M}. The BLAST algorithm generates $\mathcal{D}$ upon the evaluation of the signal to noise ratio during processing. The detection of the symbol s_{d_1} is based on its equalization row vector

$$\mathbf{w}_{d_1} = (w_{d_1,1}, w_{d_1,2}, \ldots, w_{d_1,N}) \tag{3}$$

where $\mathbf{w}_{d_1}$ is the d_1^{th} row of the MMSE inverse of the channel matrix given by

$$W = \left(\alpha I_M + H^* H\right)^{-1} H^* \tag{4}$$

with $\alpha = (2\sigma_{\text{n}}^2)/E_s$ and $w_{i,j}$ is in general the j^{th} element of $\mathbf{w}_i$. The detection process for the first symbol s_{d_1} is given by

$$\begin{aligned} y_{d_1} &= \mathbf{w}_{d_1}\mathbf{r} \\ &= \mathbf{w}_{d_1}\left(H\mathbf{s} + \mathbf{n}\right) \\ &= \mathbf{w}_{d_1} H\mathbf{s} + \mathbf{w}_{d_1}\mathbf{n} \ . \end{aligned} \tag{5}$$

Introducing a so called *Post Detection Channel* (PDC) $\widetilde{\mathbf{h}}_{d_1} = \mathbf{w}_{d_1} H$ and a *Post Detection Noise* (PDN) $\widetilde{n}_{d_1} = \mathbf{w}_{d_1}\mathbf{n}$ results in

$$\begin{aligned} y_{d_1} &= \widetilde{\mathbf{h}}_{d_1}\mathbf{s} + \widetilde{n}_{d_1} \\ &= \underbrace{\widetilde{h}_{d_1,d_1} s_{d_1}}_{(A)} + \underbrace{\left(\sum_{\substack{j=1 \\ j\neq d_1}}^{M} \widetilde{h}_{d_1,j}\cdot s_j\right)}_{(B)} + \underbrace{\widetilde{n}_{d_1}}_{(C)} \end{aligned} \tag{6}$$

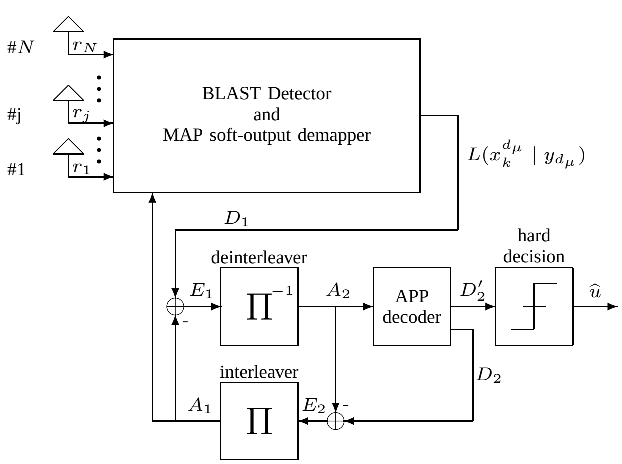

Fig. 2: Receiver structure

where $\widetilde{h}_{i,j}$ is in general the j^{th} element of $\widetilde{\mathbf{h}}_i$. Note, that $\widetilde{\mathbf{h}}_i$ is a row vector whereas $\mathbf{h}_i$ is a column vector. Term (A) is the desired symbol s_{d_1} weighted with its PDC coefficient. Term (B) is the *Post Detection Inter Symbol Interference* (PD-ISI) which consists of all other symbols weighted with their corresponding PDC coefficients. Term (C) is the PDN for symbol s_{d_1}.

Since $\mathbf{n}$ is Gaussian (i. e. its entries have Gaussian pdf) $\widetilde{n}_{d_1}$ can be seen as a realization of a Gaussian process with variance $\widetilde{\sigma}_{\text{n},d_1}^2 = \sigma_{\text{n}}^2 \sum_{i=1}^{N} |w_{d_1,i}|^2$. We can also compute the variance of the PD-ISI: $\sigma_{\text{PD-ISI},d_1}^2 = \sum_{\substack{i=1\\ i\neq d_1}}^{M} |\widetilde{h}_{d_1,i}|^2 \mathsf{E}\{|s_i|^2\}$. We assume that each antenna transmits symbols with the same energy E_s, i. e. $\mathsf{E}\{|s_i|^2\} = E_s$ $(i = 1, \ldots, M)$. The distribution of the PD-ISI is not Gaussian. However, to simplify the scheme, we summarize terms (B) and (C) into one single realization of a random process which we assume to be Gaussian as in [5] with variance $\widetilde{\sigma}_{d_1}^2 = \widetilde{\sigma}_{\text{n},d_1}^2 + \sigma_{\text{PD-ISI},d_1}^2$.

The MAP demapper in Fig. 2 performs bit-wise soft-output demapping described in [6]. Thus, the L-value of the k^{th} bit $x_k^{d_1}$ of symbol y_{d_1} is given by

$$L(x_k^{d_1} \mid y_{d_1}) = L_a\left(x_k^{d_1}\right) + \ln \frac{\sum_{\nu=0}^{2^{Q-1}-1} p\Big(y_{d_1} | x_k^{d_1}=1, x_{j,j=0,\ldots,Q-1,j\neq k}^{d_1}\Big)\cdot \mathrm{e}^{\sum_{\substack{j=0\\ j\neq k}}^{Q-1} x_j^{d_1}\cdot L_a\left(x_j^{d_1}\right)}}{\sum_{\nu=0}^{2^{Q-1}-1} p\Big(y_{d_1} | x_k^{d_1}=0, x_{j,j=0,\ldots,Q-1,j\neq k}^{d_1}\Big)\cdot \mathrm{e}^{\sum_{\substack{j=0\\ j\neq k}}^{Q-1} x_j^{d_1}\cdot L_a\left(x_j^{d_1}\right)}} \tag{7}$$

The values of the bits $x_j^{d_1}$ in (7) accomplish the following equation [6]: $\sum_{j=0}^{k-1} x_j^{d_1}\cdot 2^j + \sum_{j=k+1}^{Q-1} x_j^{d_1}\cdot 2^{j-1} = \nu$. Eq. (7) means that we set bit $x_k^{d_1}$ to 1 in the numerator and to 0 in the denominator and permute over all other bits within the symbol. Q is the number of bits per symbol. The BLAST algorithm provides no a priori information on the coded or information bits. So the complete a priori information for the very first pass is set to zero: $L_a\left(x_i^{d_1}\right) = 0 \ \forall i$. Note, that the MAP demapper performs a 1x1 demapping as for systems with only 1 transmitter and 1 receiver. Thus, its complexity is independent of the number of transmit or receive antennas and depend only on the number of bits per symbol.

To detect the next symbol s_{d_2}, in case of no cancellation, the detector goes back to (3) and replaces index d_1 with d_2. This procedure is done until all M transmitted symbols are detected. With cancellation one has to compute a reduced receive vector $\mathbf{r}_{\{d_1\}} = \mathbf{r} - \mathbf{h}_{d_1}\cdot \hat{s}_{d_1}$ where $\hat{s}_{d_1}$ is the estimate of s_{d_1} based on y_{d_1} and the index $\{d_1\}$ indicates that symbol s_{d_1} is deleted in $\mathbf{r}$. In addition, one has to null out the corresponding d_1^{th} column of H to compute the detection row vector $\mathbf{w}_{d_2}$. Now, in (5), d_1 and $\mathbf{r}$ have to be replaced by d_2 and $\mathbf{r}_{\{d_1\}}$, respectively, to compute y_{d_2}. Following this detection principle [1] equation (6) rewrites for an arbitrary detection step d_μ ($1 \le \mu \le M$) as

$$y_{d_\mu} = \underbrace{\widetilde{h}_{d_\mu,d_\mu} s_{d_\mu}}_{(A)} + \underbrace{\left(\sum_{j \in \overline{\mathcal{D}_{d_\mu}}} \widetilde{h}_{d_\mu,j} \cdot s_j \right)}_{(B)} + \underbrace{\widetilde{n}_{d_\mu}}_{(C)} + \underbrace{\sum_{j \in \mathcal{D}_{d_\mu}} \mathbf{w}_{d_\mu} \mathbf{h}_j \left(s_j - \hat{s}_j \right)}_{(D)} \; . \tag{8}$$

$\mathcal{D}_{d_\mu} = \{d_1, d_2, \ldots, d_{\mu-1}\}$ is the set of all yet detected symbols and $\overline{\mathcal{D}_{d_\mu}} = \{d_{\mu+1}, d_{\mu+2}, \ldots, d_M\}$ are the remaining symbols. Thus, $\mathcal{D}_{d_\mu} + \{d_\mu\} + \overline{\mathcal{D}_{d_\mu}} = \mathcal{D}$. Equation (8) differs from (6) mainly in terms (B) and (D). In fact, the sums in these two terms are disjoint in the sense that they have no summing indices in common. Term (B) still is the PD-ISI of all not yet detected symbols whereas term (D) represents the *Post Cancellation Interference* (PCI). In fact, term (D) is the summation of the PCI over all symbols that have been detected and cancelled and is the reason for error propagation. From (8) it can be seen that all the transmitted symbols s_j ($j \neq d_\mu$) interfere with each other either as PD-ISI or PCI if $s_j \neq \hat{s}_j$. Term (D) vanishes for the correct decided symbols with $s_j = \hat{s}_j$.

In [7] the enormous potential of *genie* cancellation was shown. With genie cancellation the order of diversity increases with each detection step. But it was also shown that the error propagation counters this potential and the overall performance is reduced. To mitigate the negative effect of error propagation some proposals were made, e. g. [7]–[14], which include grouping, iterative re-ordering, and iterative cancellation. All these approaches have in common that there is an additional computational effort and in most cases some iterations are required. Furthermore, the improvement in terms of BER is demonstrated for the uncoded system. With a coded system the decoding process should know, whether cancellation is perfect or not, i. e. whether term (D) in (8) is zero or not. However, this information is not present during demapping and decoding. As a consequence the soft output information of the demapper may be wrong and the decoder runs into trouble. The method presented in the next Section does not have this drawback and is computationally inexpensive. Simulation results are presented in Section IV.

III. BLAST WITH ADAPTIVE CANCELLATION

A. Principle Idea

A solution is, only to cancel out those symbols that are recognized as correct with a high reliability. As a consequence we can avoid error propagation and gain diversity if the cancellation is correct. But the problem is how to find out those symbols? The decoding and error correction process starts after all symbols are soft output demapped. However, this is too late, because the BLAST cancellation is already done. Horizontal decoding [4] could help, but an efficient decoder needs to have independent and thus interleaved input. But a per-layer-interleaver in combination with symbol cancellation introduces detection time delay and complexity.

For the sake of completeness, the scheme in Fig. 2 is already prepared for iterative processing. The a priori information A_1 is subtracted from the a posteriori information D_1 given by the L-values in (7). After deinterleaving of E_1, the result A_2 is passed on as new a priori information to the soft input soft output APP decoder. The decoder calculates the new a posteriori information D_2. The extrinsic information E_2 is interleaved resulting in new a priori information A_1 for the detector and demapper. The feedback information A_1 can be used for iterative schemes as e. g. presented in [5], [6], [15]. After some iterations the output L-values D_2' of the decoder are hard decided to obtain estimates $\widehat{u}$ on the information bits. In this paper we focus on the improvement of the very first detection step (0^{th} iteration). Note, that for ZF $\alpha = 0$ and $\widetilde{H} = I_M$ resulting in $(B) = 0$ and $\widetilde{h}_{d_\mu,d_\mu} = 1$ for symbol s_{d_μ} ($\mu = 1, 2, \ldots, M$), but still $(D) \neq 0$.

B. Reliability measure for symbol decision

A measure for the reliability of a bit b_k is the magnitude of its likelihood value (L-value) defined as $L(b_k) = \ln \frac{P[b_k=1]}{P[b_k=0]}$, where $P[\cdot]$ is the probability. As an extension, the reliability value $\mathcal{R}(s)$ of a symbol s consisting of Q bits can be based upon the L-values of these underlying bits. Two simple (different) measures are proposed as

$$\mathcal{R}(s) = \sum_{j=1}^{Q} |L(b_j)| \tag{9}$$

$$\text{or} \qquad \mathcal{R}(s) = \min_{j=1,\ldots,Q} |L(b_j)| \tag{10}$$

During our simulations we found that (10) is a conservative approach and suited to achieve low BER. So we have taken this metric in our algorithm. Suppose we have just detected the symbol y_{d_μ} and have computed the L-values $L(x_k^{d_\mu})$ ($k = 1, \ldots, Q$). Then we compute $\mathcal{R}\left(y_{d_\mu}\right)$ as in (10). The probability density functions (pdf) of $\mathcal{R}\left(y_{d_\mu}\right)$ is $p\left(\mathcal{R}\left(y_{d_\mu}\right)\right)$. The key is, that we split this pdf into two parts and define

$$\begin{aligned} p_{\mathrm{c},y_{d_\mu}} &= p\left(\mathcal{R}\left(y_{d_\mu}\right) | \hat{s}_{d_\mu} = s_{d_\mu}\right) \\ \text{and} \qquad p_{\mathrm{w},y_{d_\mu}} &= p\left(\mathcal{R}\left(y_{d_\mu}\right) | \hat{s}_{d_\mu} \neq s_{d_\mu}\right) \end{aligned}$$

where the indices c and w stand for correct and wrong decision, respectively. The receiver of course does not know whether $\hat{s}_{d_\mu} = s_{d_\mu}$ or $\hat{s}_{d_\mu} \neq s_{d_\mu}$. But as can be seen from Fig. 3 the two distributions are quite different. In this figure we show the simulated distributions of the first symbol y_{d_1} to be detected in a system with $M = N = 2$, QPSK, Gray mapping at an operation point of $E_b/N_0 = 10\,\mathrm{dB}$. During our investigation we observed that, if we change the number of bits per symbol, use anti-Gray mappings or operate at different E_b/N_0, the distributions obtained are basically of the same shape. Thus, the depicted distributions in Fig. 3 are typical to explain the scheme. The pdfs $p_{\mathrm{c},y_{d_1}}$ are significantly larger than $p_{\mathrm{w},y_{d_1}}$. In case of wrong estimation the underlying histogram has almost no entries for $\mathcal{R}(y_{d_1}) > 14$. The

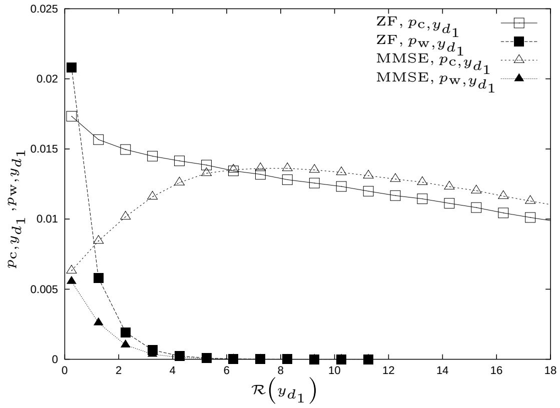

Fig. 3: pdfs of minimum absolute L-values for QPSK, Gray mapping, $E_b/N_0 = 10\,\mathrm{dB}$, $M = N = 2$, both ZF and MMSE

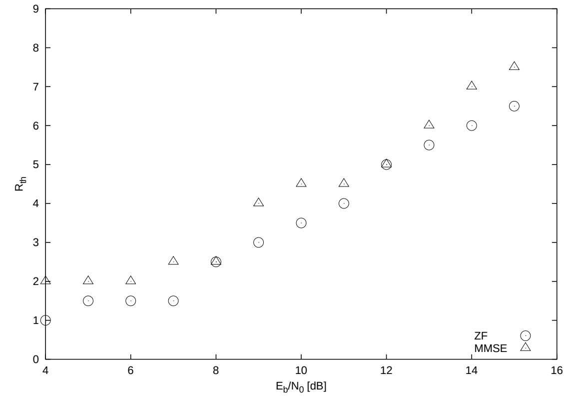

Fig. 4: Optimum thresholds for ZF and MMSE, QPSK, Gray mapping, and $M = N = 2$

simulated pdfs of the MMSE case differ a little bit from the ZF case, especially for low $\mathcal{R}\left(y_{d_1}\right)$. In particular, for MMSE the two distributions $p_{\mathrm{c},y_{d_1}}$ and $p_{\mathrm{w},y_{d_1}}$ are better 'separated', i. e.:

$$\int_0^\infty p_\mathrm{c}\left[y_{d_1}\right] p_\mathrm{w}\left[y_{d_1}\right] d\mathcal{R}\left(y_{d_1}\right)\Big|_{\mathrm{MMSE}} < \int_0^\infty p_\mathrm{c}\left[y_{d_1}\right] p_\mathrm{w}\left[y_{d_1}\right] d\mathcal{R}\left(y_{d_1}\right)\Big|_{\mathrm{ZF}} \quad (11)$$

This leads us to the conclusion that our presented scheme is more advantageous for MMSE than for ZF algorithm.

In fact, if $\mathcal{R}\left(y_{d_\mu}\right)$ is high (e. g. $\mathcal{R}\left(y_{d_\mu}\right) = 15$) then it is very likely that $\hat{s}_{d_\mu} = s_{d_\mu}$ because $p\left[\mathcal{R}\left(y_{d_\mu}\right)|\hat{s}_{d_\mu} \neq s_{d_\mu}\right] \approx 0$. On the other hand, if $\mathcal{R}\left(y_{d_\mu}\right)$ is low we are not sure whether $\hat{s}_{d_\mu} = s_{d_\mu}$ or not. But with this knowledge we can set a threshold $\mathcal{R}_\mathrm{th}$ and apply the following rule:

- cancel the detected symbol $\hat{s}_{d_\mu}$ if $\mathcal{R}\left(y_{d_\mu}\right) > \mathcal{R}_\mathrm{th}$
- do not cancel otherwise

C. Definition of threshold

The higher we choose the threshold the more conservative the cancellation process is. That means that if we only want to cancel those symbols which are extremely save, we have to choose a high threshold. For $\mathcal{R}_\mathrm{th} \to \infty$ we do not cancel at all, whereas for $\mathcal{R}_\mathrm{th} = 0$ we have permanent cancellation. We need the demapper providing the L-values to be part of the detector to be able to decide whether to cancel or not.

It remains to find the optimal threshold that yields low BER. For this reason we have to find the optimum between two cases: If the threshold is too high we hardly cancel anything and cannot benefit from the gain of genie cancellation. Contrary, if the threshold is too low we run the risk of cancelling wrong symbols resulting in a worse performance. Fig. 4 shows the optimum threshold points $\mathcal{R}_{th}$ over E_b/N_0 found by simulation for QPSK, Gray mapping, and $M = N = 2$. In contrast to OSIC it is now very likely to happen that there are already detected symbols which will be deleted and others will not. This modifies the BLAST algorithm. In the calculation of the equalization row vectors $\mathbf{w}_{d_\mu}$ we only have to null out those columns of the channel matrix H whose corresponding symbols are actually detected *and* deleted. Following this principle, (8) becomes

$$y_{d_\mu} = \underbrace{\widetilde{h}_{d_\mu,d_\mu} s_{d_\mu}}_{(A)} + \underbrace{\left(\sum_{j \in \overline{\mathcal{C}_{d_\mu}}} \widetilde{h}_{d_\mu,j} \cdot s_j\right)}_{(B)} + \underbrace{\widetilde{n}_{d_\mu}}_{(C)} + \underbrace{\sum_{j \in \mathcal{C}_{d_\mu}} \mathbf{w}_{d_\mu} \mathbf{h}_j \left(s_j - \hat{s}_j\right)}_{(D)} . \quad (12)$$

where $\mathcal{C}_{d_\mu}$ is the set of all *cancelled* symbols until this detection step and $\overline{\mathcal{C}_{d_\mu}}$ are all not yet detected symbols plus the symbols that have been detected but not cancelled excluding the symbol to be detected in this step. Note that $\mathcal{C}_{d_\mu}$ is a subset of $\mathcal{D}_{d_\mu}$.

IV. Simulation results

Fig. 5 shows the simulation results for a system with $M = N = 2$ antennas. We included both, MMSE and ZF, with (a)

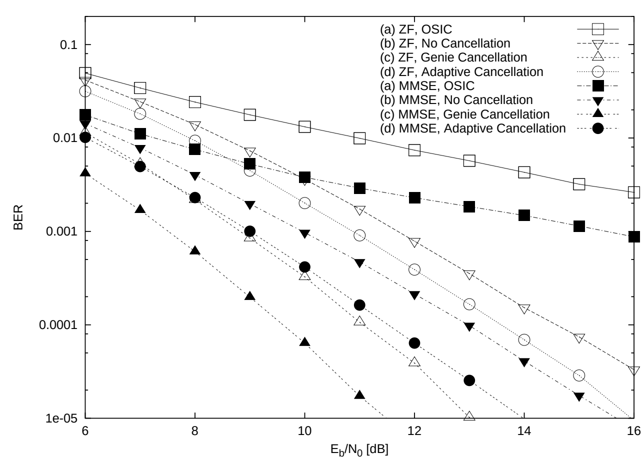

Fig. 5: BER for $M = N = 2$, QPSK, Gray mapping, both ZF and MMSE

OSIC, (b) no cancellation at all, (c) genie cancellation, and (d) adaptive threshold-based cancellation. It can be seen that OSIC the performs worse than all other schemes: the advantage of OSIC in an uncoded BLAST system is no longer visible due to too optimistic L-values as input signal to the decoder. We also recognize that at a BER of 10^{-4} there is a gain of about 3.5 dB between the genie cancellation scheme and the scheme where no cancellation is done. Our proposed adaptive threshold-based cancellation method at BER of 10^{-4} outperforms the no-cancellation-case by approximately 1 dB for ZF and 1.5 dB for MMSE. This is achieved without reordering the detection order and without any iteration. Thus, further iteration steps have got an improved starting point. For MMSE our proposal almost reaches the performacne of the ZF genie cancellation. Note, that the presented scheme can easily be extended to systems with more than 2 antennas.

V. Remarks on Computational Complexity

The basic BLAST algorithm and the demapper are the same for all schemes. But the new proposals takes the demapper as part of the detector. This is not necessary for the regular BLAST architecture. The L-values are computed anyway to be the input of the decoder. So the additional complexity of our proposal is just the execution of (10), which is almost negligible. If we omit cancellation due to a low minimum absolute value of $\mathcal{R}\left(y_{d_\mu}\right)$ the complexity reduces compared to full OSIC. It has to be pointed out that there is no iteration, reordering or additional interference cancellation required. To use the optimal threshold by applying a lookup table, the knowledge of the SNR is required. However, with MMSE equalization according to (4), α is also required. Thus, it has to be estimated in both cases.

VI. Conclusion

The BLAST algorithm is an effective way to detect output signals of a MIMO transmission system. BLAST can be operated with zero forcing (ZF) or minimum mean square error (MMSE). Both schemes can be combined with ordered successive interference cancellation (OSIC). In this paper we have presented an improved method to reduce the BER of a convolutionally encoded MIMO system. The problem of conventional BLAST with OSIC is that it works fine, if cancellation of correctly decided symbols is done. A relatively high SNR is a prerequisite of that. If the SNR is low, symbol decision errors occur which give rise to error propagation, which reduces the performance. In our proposal we follow the idea to cancel only symbols, which are decided correctly. As a consequence, we reduce the error propagation effect of the conventional OSIC. To do so, we define a reliability measure, which indicates the correctness of the symbol decision. We propose the minimal L-value $\mathcal{R}(s)$ of the underlying bits of the symbol s as a measure. From statistical evaluations of the conditional probabilities of $\mathcal{R}(s)$ given in Fig. 3 we derive a threshold $\mathcal{R}_{th}$ for cancellation, which depends on the SNR of the channel (Fig. 4). The proposed receiver is composed of a conventional BLAST detector with an adaptive threshold for cancellation, an MAP soft output demapper and an iterative (Turbo-) decoder. We have tested our algorithm by extensive computer simulations using Rayleigh fading MIMO channels with additive white Gaussian noise. As a result, we get an SNR gain of about 1 dB with zero forcing and 1.5 dB with minimum mean square error, both compared to the conventional BLAST algorithm. We have also shown that the hardware overhead of our proposal is very small.

References

[1] P. W. Wolniansky, G. J. Foschini, G. D. Golden, and R. A. Valenzuela, “V-BLAST: An architecture for realizing very high data rates over the rich-scattering wireless channel,” in *1998 International Symposium on Signals, Systems, and Electronics ISSSE’98*, Pisa, Italy, September 30, 1998.

[2] G. D. Golden, G. J. Foschini, R. A. Valenzuela, and P. W. Wolniansky, “Detection algorithm and initial laboratory results using V-BLAST space-time communication architecture,” *ELECTRONIC LETTERS*, vol. 35, no. 1, pp. 14–15, January 7, 1999.

[3] A. van Zelst, “Extending the capacity of next generation wireless LANs using space division multiplexing combined with OFDM,” Master’s thesis, Technische Universiteit Eindhoven, Faculty of Electrical Engineering, Section of Telecommunication Technology and Electromagnetics, Netherlands, 1999.

[4] X. Li, H. Huang, G. J. Foschini, and R. A. Valenzuela, “Effects of iterative detection and decoding on the performance of BLAST,” in *IEEE Global Telecommunications Conference 2000 (Globecom)*, San Francisco, USA, November 27 - December 1, 2000, pp. 1061–1066.

[5] W.-J. Choi, K.-W. Cheong, and J. M. Cioffi, “Iterative soft interference cancellation for multiple antenna systems,” in *IEEE Wireless Communications and Networking Conference WCNC 2002*, 2002.

[6] S. ten Brink, J. Speidel, and R.-H. Yan, “Iterative demapping and decoding for multilevel modulation,” in *IEEE Global Telecommunications Conference 1998 (Globecom)*, November 1998, pp. 579–584.

[7] S. Bäro, G. Bauch, A. Pavlic, and A. Semmler, “Improving BLAST performance using space-time block codes and turbo decoding,” in *IEEE Global Telecommunications Conference 2000 (Globecom)*, November 2000, pp. 1067–1071.

[8] A. Benjebbour, H. Murata, and S. Yoshida, “Performance of iterative successive detection algorithm for space-time transmission,” in *IEEE Vehicular Technology Conference VTC 2001 Spring*, Rhodes, Greece, May 2001, pp. 1287–1291.

[9] X. Li, H. C. Huang, A. Lozano, and G. J. Foschini, “Reduced-complexity detection algorithms for systems using multi-element arrays,” in *IEEE Global Telecommunications Conference 2000 (Globecom)*, San Francisco, USA, November 27 - December 1, 2000.

[10] J. W. Kang and K. B. Lee, “Layered structure ML detection scheme for MIMO systems,” in *The 12th Joint Conference on Communications and Information JCCI 2002*, 2002.

[11] W.-J. Choi, R. Negi, and J. M. Cioffi, “Combined ML and DFE decoding for the V-BLAST system,” in *IEEE International Conference on Communications ICC 2000*, New Orleans, LA, USA, June 18-22, 2000, pp. 1243–1248.

[12] A. Bhargave, R. J. P. de Figueiredo, and T. Eltoft, “A detection algorithm for the V-BLAST system,” in *IEEE Global Telecommunications Conference 2001 (Globecom)*, San Antonio, USA, November 2001, pp. 494–498.

[13] W. Zha and S. D. Blostein, “Modified decorrelation decision-feedback detection of BLAST space-time system,” in *IEEE International Conference on Communications ICC 2002*, New York, NY, USA, April 28 - May 2, 2002, pp. 335–339.

[14] C. M. Degen, C. M. Walke, and B. Rembold, “Comparative study of efficient decision-feedback equalization schemes for MIMO systems,” in *European Wireless 2002*, Florence, Italy, February 25-28, 2002.

[15] M. Witzke, S. Bäro, F. Schreckenbach, and J. Hagenauer, “Iterative detection of MIMO signals with linear detectors,” in *36th Asilomar Conference On Signals, Systems And Computers ACSSC 2002*, Pacific Grove, CA, USA, November 3-6, 2002.