Successive Detection Algorithm for Frequency Selective Layered Space-Time Receiver

Dirk Wübben, Volker Kühn, and Karl-Dirk Kammeyer
Department of Communications Engineering
University of Bremen
Otto-Hahn-Alle
D-28359 Bremen, Germany
Email: {wuebben,kuehn,kammeyer}@ant.uni-bremen.de

Abstract— **The use of multiple antenna systems provides a great performance advantage in Rayleigh fading environments. To exploit this advantage several schemes like the layered space-time architecture have been proposed, but mostly restricted to narrowband transmission. Recently, a generalization of the well known V-BLAST detection algorithm for frequency-selective fading channels has been presented by Lozano and Papadias. In order to improve the performance of this receiver, we propose an iterative detection algorithm in this paper.**

I. Introduction

In a Rayleigh fading environment multiple antenna systems provide an enormous increase in capacity compared to single antenna systems [1]. Consequently, multiple-input multiple-output (MIMO) systems are predestined for high data rate wireless communications. To exploit this potential, several schemes like Space-Time Block Codes and Space-Time Trellis Codes have been presented. Foschini proposed another MIMO system containing a diagonally layered coding scheme named D-BLAST (*Diagonal Bell Labs Layered Space-Time*) [2]. A simplified scheme was proposed in [3] and is known as V-BLAST (*Vertical* BLAST), which associates each layer with a specific transmit antenna. These layers are detected by a successive interference cancellation technique which nulls the interferer by linearly weighting the received signal vector with a zero-forcing nulling vector. The extension to MMSE nulling was presented in [4] and a computational efficient detection algorithm utilizing the QR decomposition was proposed in [5]. In [6] an enhanced iterative detection scheme called Backward Iterative Cancellation (BIC) was presented. It improves the system performance by cancelling high diversity streams at each iteration and thereby increases the system diversity.

Most MIMO algorithms presented up to now have been restricted to narrowband transmission. For real applications these algorithms should be extended to channels, where the signal bandwidth exceeds the coherence bandwidth of the channel. Recently, such a generalization of the V-BLAST detection algorithm for frequency-selective fading channels has been presented by Lozano and Papadias. It employs a multiple-input single-output (MISO) decision feedback structure to detect the distinct layers in a successive way and we call it frequency-selective BLAST (FS-BLAST) here. In this paper, we combine the basic ideas of FS-BLAST with BIC to significantly improve the performance of frequency-selective MIMO systems.

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The remainder of this paper is organized as follows. The MIMO system is described in Section II and the FS-BLAST algorithm is reviewed in Section III. The new approach of extending the iterative BIC algorithm to FS-BLAST is introduced in Section IV and the performance improvement is investigated. A summary and conclusion marks can be found in Section V.

II. System description

We consider the frequency-selective (FS) multiple antenna system with n_T transmit and $n_R \geq n_T$ receive antennas shown in **Fig. 1**. The data is demultiplexed in n_T data substreams of equal length (called layers). These substreams are mapped into M-PSK or M-QAM symbols, organized in frames and transmitted over the n_T antennas at the same time.

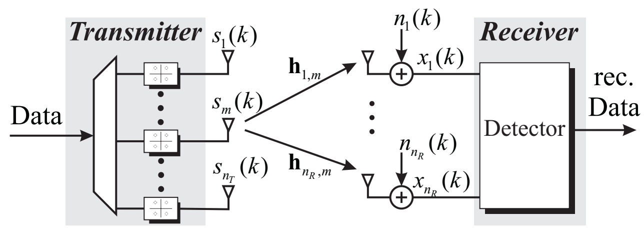

Fig. 1. Model of a frequency-selective MIMO system with n_T transmit and n_R receive antennas.

In order to describe the FS-MIMO system, we denote the symbol transmitted by antenna m at time instant k by $s_m(k)$ with average symbol energy σ_s^2 and likewise the signal received at antenna n is labeled by $x_n(k)$. The frequency-selective single-input single-output (SISO) channel impulse response from transmit antenna m to receive antenna n is given by the $(L+1)\times 1$ vector[1] $\mathbf{h}_{n,m} = [h_{n,m}(0)\ h_{n,m}(1)\ \ldots\ h_{n,m}(L)]^T$. For simplicity, we assume the same channel order L for all SISO channels $\mathbf{h}_{n,m}$. It is further assumed that the channel is constant over the frame length, but may change from frame to frame (*block fading channel*) and is perfectly known by the receiver. We use the power normalization ($\mathrm{E}\{\|\mathbf{h}_{n,m}\|^2\} = 1$) for the i.i.d. complex fading gains of each SISO channel to obtain a passive channel.

To describe the input-output relation between transmit antenna m and all n_R receive antennas the $n_R{\times}(L{+}1)$ frequency-selective SIMO channel matrix

$$\mathbf{H}_m = \begin{bmatrix} \mathbf{h}_{1,m}^T \\ \vdots \\ \mathbf{h}_{n_R,m}^T \end{bmatrix} \tag{1}$$

is introduced. With the sequence of $L+1$ symbols transmitted by antenna m

$$\mathbf{s}_m(k) = \begin{bmatrix} s_m(k) \\ \vdots \\ s_m(k-L) \end{bmatrix} \tag{2}$$

the received signal vector $\mathbf{x}(k) = [x_1(k)\ \ldots\ x_{n_R}(k)]^T$ of dimension $n_R \times 1$ becomes a linear superposition of n_T frequency-selective SIMO transmissions and noise:

$$\mathbf{x}(k) = \sum_{m=1}^{n_T} \mathbf{H}_m\, \mathbf{s}_m(k) + \mathbf{n}(k)\ . \tag{3}$$

In (3) vector $\mathbf{n}(k)$ depicts the noise at the n_R receive antennas at symbol time k, assuming uncorrelated white gaussian noise and spatial covariance $\mathrm{E}\left\{\mathbf{n}(k)\,\mathbf{n}^H(k)\right\} = \sigma^2\,\mathbf{I}_{n_R}$.

III. Space-Time Receiver

A. *Principle of FS-BLAST*

To detect the transmitted signals, appropriate schemes to mitigate the influence of CCI (co-channel interference) and ISI (intersymbol interference) for each transmit signal have to be investigated. In this paper, we use the frequency-selective extension of V-BLAST as proposed in [7], [8] and call it FS-BLAST. Similar to V-BLAST, in each detection step one layer is treated as target layer and all other layers are treated as interferers. The target layer is detected by suppressing the co-channel interferers, the estimated signals are subtracted from the received signal, before the remaining layers are detected in the same successive way.

Therefore a decision feedback equalizer (DFE) structure consisting of a MISO feedforward filter (FFF) and a SISO feedbackward filter (FBF) is employed in each detection stage as shown in **Fig. 2**.

The sequence of received signals is filtered by the FFF in space and time domain. The aim of this filter is to extract the target layer m, but suppress the other layers and also part of the own ISI. The FBF filters the sequence of decisions on previously detected symbols delayed by k_0 time steps and its output is subtracted from the FFF output. Thus, the feedbackward filter is used to remove that part of ISI from the present estimate caused by previously detected symbols.

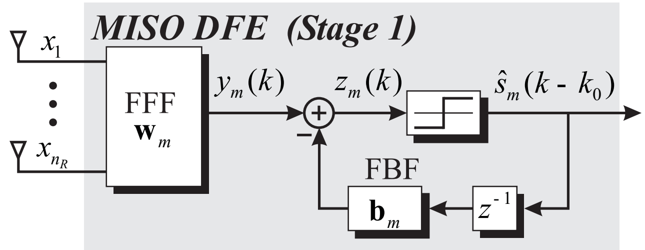

Fig. 2. Initial MISO-DFE stage of FS-BLAST to detect layer m.

B. *MISO-Filter Design*

To calculate the FFF and the FBF, the input-output relation in (3) has to be extended to describe a *sequence* of received signals. With regard to the FFF order N, we denote the sequence of $N+1$ received signals by the $n_R\,(N+1)\times 1$ vector

$$\underline{\mathbf{x}}(k) = \begin{bmatrix} \mathbf{x}(k) \\ \vdots \\ \mathbf{x}(k-N) \end{bmatrix} \tag{4}$$

and the transmit signal vector (2) is extended to the $(N+L+1)\times 1$ sequence vector

$$\underline{\mathbf{s}}_m(k) = \begin{bmatrix} s_m(k) \\ \vdots \\ s_m(k-N-L) \end{bmatrix}\ . \tag{5}$$

By utilizing the $n_R\,(N{+}1){\times}(N{+}L{+}1)$ block Toeplitz matrix

$$\underline{\mathbf{H}}_m = \begin{bmatrix} \boxed{\mathbf{H}_m} & & & \\ & \boxed{\mathbf{H}_m} & & \\ & & \ddots & \\ & & & \boxed{\mathbf{H}_m} \end{bmatrix} \tag{6}$$

and the noise sequence $\underline{\mathbf{n}}(k)$ the input-output relation for a sequence of received signals (4) is given by

$$\underline{\mathbf{x}}(k) = \sum_{m=1}^{n_T} \underline{\mathbf{H}}_m\, \underline{\mathbf{s}}_m(k) + \underline{\mathbf{n}}(k)\ . \tag{7}$$

At the first stage of detection, this sequence of received signals is filtered by the FFF $\mathbf{w}_m$ and the filter output becomes

$$y_m(k) = \sum_{\tau=0}^{N} \mathbf{w}_m(\tau)\, \mathbf{x}(k-\tau) = \mathbf{w}_m\, \underline{\mathbf{x}}(k)\ . \tag{8}$$

The row vector $\mathbf{w}_m = [\mathbf{w}_m(0)\,\mathbf{w}_m(1)\,\ldots\,\mathbf{w}_m(N)]$ of dimension $1{\times}n_R\,(N{+}1)$ denotes the FFF and consists of $N{+}1$ space-only filter taps $\mathbf{w}_m(\tau) = [w_{m,1}(\tau)\ \ldots\ w_{m,n_R}(\tau)]$ each of dimension $1{\times}n_R$.

[1]Throughout the remainder, $(\cdot)^*$, $(\cdot)^T$ and $(\cdot)^H$ denote the conjugation, the matrix transposition and the hermitian transposition, respectively. Furthermore $\mathbf{I}_\alpha$ indicates the $\alpha\times\alpha$ identity matrix and $\mathbf{0}_{\alpha,\beta}$ denotes the $\alpha\times\beta$ all zero matrix.

To get an ISI-free representation for $s_m(k-k_0)$, the sequence of L_S most recent decisions of target layer m

$$\hat{\mathbf{s}}_m(k-k_0-1) = \begin{bmatrix} \hat{s}_m(k-k_0-1) \\ \vdots \\ \hat{s}_m(k-k_0-L_S) \end{bmatrix}$$

is filtered by the $1 \times L_S$ feebackward filter (FBF) $\mathbf{b}_m = [b_m(1)\; b_m(2)\; \ldots\; b_m(L_S)]$, with decision delay k_0 to be optimized in the range $0 \le k_0 \le N+L-L_S$. Finally the output of the FBF is subtracted from the FFF output

$$\begin{aligned} z_m(k) &= \mathbf{w}_m\, \underline{\mathbf{x}}(k) - \mathbf{b}_m\, \hat{\mathbf{s}}_m(k-k_0-1) \\ &= [\mathbf{w}_m\; \mathbf{b}_m] \cdot \begin{bmatrix} \underline{\mathbf{x}}(k) \\ -\hat{\mathbf{s}}_m(k-k_0-1) \end{bmatrix} \\ &= \mathbf{f}_m\, \mathbf{a}_m \end{aligned} \tag{9}$$

and decided by utilizing an appropriate quantization device to form the estimation $\hat{s}_m(k-k_0)$ (see **Fig. 2**). In (9) the $1 \times n_R\,(N+1) + L_S$ row vector $\mathbf{f}_m = [\mathbf{w}_m\; \mathbf{b}_m]$ contains the FFF and FBF coefficients for layer m, whereas column vector $\mathbf{a}_m$ denotes the currently received sequence and (under the assumption of correct previous decisions $\hat{s}_m(k-\tau) = s_m(k-\tau)$ for $k_0+1 \le \tau \le k_0+L_S$) the negative of the transmitted sequence from antenna m delayed by k_0+1 taps. The MSE solution for the filter design is found by minimizing the cost-function [8]

$$\begin{aligned} J_{MSE} &= \mathrm{E}\left\{|z_m(k) - s_m(k-k_0)|^2\right\} \\ &= \mathrm{E}\left\{|\mathbf{f}_m\, \mathbf{a}_m - s_m(k-k_0)|^2\right\} \\ &= \mathbf{f}_m\, \mathbf{Q}_m\, \mathbf{f}_m^H - \mathbf{f}_m\, \mathbf{p}_m - \mathbf{p}_m^H\, \mathbf{f}_m^H + \sigma_s^2 \end{aligned} \tag{10}$$

with the $n_R\,(N+1) + L_S \times 1$ vector[2]

$$\mathbf{p}_m = \mathrm{E}\left\{\mathbf{a}_m s_m^*(k-k_0)\right\} = \sigma_s^2 \begin{bmatrix} (\underline{\mathbf{H}}_m)_{k_0+1} \\ \mathbf{0}_{N+L+1,1} \end{bmatrix} \tag{11}$$

and the $n_R\,(N+1) + L_S \times n_R\,(N+1) + L_S$ matrix

$$\mathbf{Q}_m = \mathrm{E}\left\{\mathbf{a}_m \mathbf{a}_m^H\right\} = \begin{bmatrix} \mathbf{R}_{xx} & -\mathbf{R}_{s_m x}^H \\ -\mathbf{R}_{s_m x} & \mathbf{R}_{s_m s_m} \end{bmatrix}. \tag{12}$$

The $n_R(N+1) \times n_R(N+1)$ covariance matrix $\mathbf{R}_{xx}$ in (12) is defined by

$$\begin{aligned} \mathbf{R}_{xx} &= \mathrm{E}\left\{\underline{\mathbf{x}}(k)\, \underline{\mathbf{x}}^H(k)\right\} \\ &= \sigma_s^2 \sum_{m=1}^{n_T} \underline{\mathbf{H}}_m\, \underline{\mathbf{H}}_m^H + \sigma^2\, \mathbf{I}_{n_R(N+1)}\,, \end{aligned} \tag{13}$$

the cross correlation between signal vector $\hat{\mathbf{s}}_m(k-k_0-1)$ and receive sequence $\underline{\mathbf{x}}(k)$ is calculated according to

$$\begin{aligned} \mathbf{R}_{s_m x} &= \mathrm{E}\left\{\hat{\mathbf{s}}_m(k-k_0-1)\, \underline{\mathbf{x}}^H(k)\right\} \\ &= \sigma_s^2\, (\underline{\mathbf{H}}_m)^H_{k_0+2\ldots k_0+L_S+1} \end{aligned} \tag{14}$$

and the $L_S \times L_S$ correlation matrix of $\hat{\mathbf{s}}_m(k)$ is given by

$$\mathbf{R}_{s_m s_m} = \mathrm{E}\left\{\hat{\mathbf{s}}_m(k)\, \hat{\mathbf{s}}_m^H(k)\right\} = \sigma_s^2 \mathbf{I}_{L_S}\,. \tag{15}$$

[2] $(\underline{\mathbf{H}}_m)_\alpha$ denotes column α of matrix $\underline{\mathbf{H}}_m$.

By expanding the cost function in (10) to quadratic form $J_{MSE} = \left(\mathbf{f}_m \mathbf{Q}_m - \mathbf{p}_m^H\right) \mathbf{Q}_m^{-1} \left(\mathbf{Q}_m \mathbf{f}_m^H - \mathbf{p}_m\right) - \mathbf{p}_m^H \mathbf{Q}_m^{-1} \mathbf{p}_m + \sigma_s^2$, the minimum of J_{MSE} is obviously donated by the filter vector $\mathbf{f}_m = \mathbf{p}_m^H \mathbf{Q}_m^{-1}$ and the mean square error is given by $\mathrm{MSE}_m = \sigma_s^2 - \mathbf{p}_m^H \mathbf{Q}_m^{-1} \mathbf{p}_m$.

According to [8] in each detection step l the layer m with the smallest MSE_m is selected and detected with an optimized delay k_0 taken into account. After detecting the layer, its interference can be removed from the received signal $\mathbf{x}(k)$ similar to V-BLAST [2] and the corresponding entries in the channel matrix are cancelled. The remaining layers are detected in the same way and the output of detection stage l is depicted by $\hat{s}_{(l)}(k)$ (**Fig. 3**).

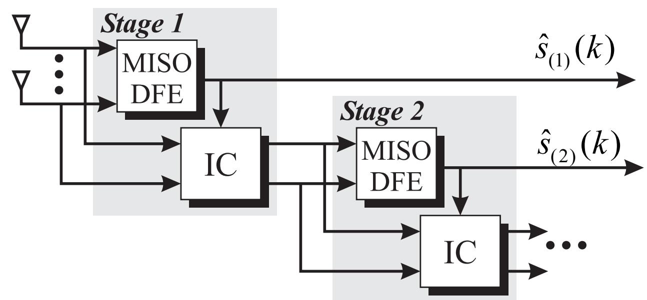

Fig. 3. First two MISO-DFE stages of the successive interference suppression and cancellation structure.

C. ML-Detection

By regarding the MISO-DFE stage in (**Fig. 2**) in terms of channel impulse shortening [9], the MISO filter creates a SISO channel of order L_S between transmit antenna m and the output $y_m(k)$ of the FFF. The target impulse response (TIR) of this SISO channel is given by the $1 \times (L_S+1)$ row vector $\tilde{\mathbf{b}}_m = [b_m(0)\; \mathbf{b}_m] = [b_m(0)\; b_m(1)\; \ldots\; b_m(L_S)]$ and consequently a standard Viterbi algorithm can be used for equalizing the FFF output stream with a (reduced) channel order L_S. Under ideal conditions the filter tap $b_m(0)$ should be equal to one, whereas in the present of noise it is not. Its value is determined by $b_m(0) = \mathbf{w}_m\, (\underline{\mathbf{H}}_m)_{k_0+1}$.

In the remainder of this paper, we always assume MLSE of the FFF output, which outperforms a time domain DFE in general. Therefore, the design criteria for a FS-BLAST receiver are given by the feedforward filter order N, the decision delay k_0 and the order L_S of the target impulse response. By reducing the order L_S, the equalizer effort is decreased enormously at the expense of performance degradation.

D. Simulation Result

In this section we investigate the performance of FS-BLAST with Viterbi equalization in each stage. **Fig. 4** shows the bit error rate (BER) for a system with $n_T = 4$ and $n_R = 6$ antennas for a frequency-selective channel of order $L = 6$, a FFF-order of $N = 10$ and a varying number of TIR-taps $L_S + 1$. As reference, the BER of V-BLAST within a non-frequency-selective (NFS) channel is also plotted.

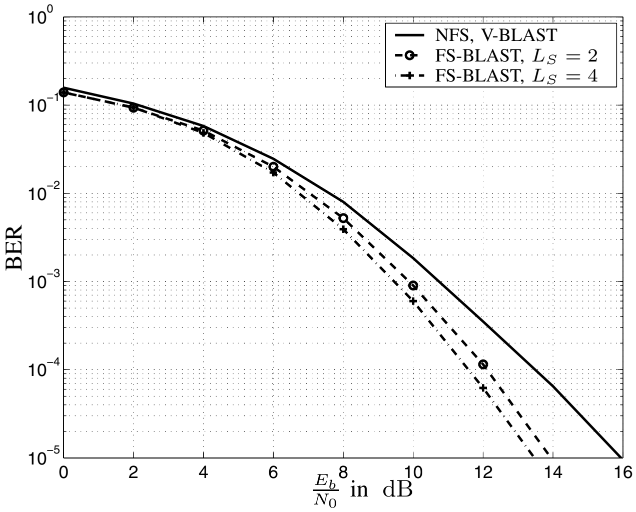

Fig. 4. Simulation with $n_T = 4$ and $n_R = 6$ antennas, QPSK modulation, frequency-selective channel of order $L = 6$, FS-BLAST with FFF order $N = 10$.

Obviously, the FS system outperforms the system with flat fading, due to the increase in diversity. Depending on the order L_S of the target impulse response, the bit error performance improves, but with the cost of an increasing computational complexity.

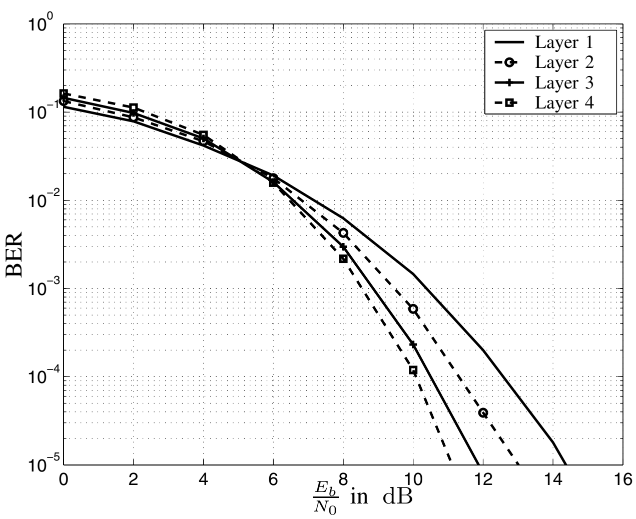

Fig. 5. BER per layer for system with $n_T = 4$ and $n_R = 6$ antennas, QPSK modulation, frequency-selective channel of order $L = 6$, FS-BLAST with FFF order $N = 10$ and TIR order $L_S = 4$.

Fig. 5 shows the BER for each layer in order of detection. In the lower SNR region error propagation has a large influence, hence the layer detected first performs slightly better than the layers detected in the subsequent. The BER curves traverse at a SNR of approximately $5\,\mathrm{dB}$, which presents the decreasing influence of error propagation. Consequently, the performance of the later detected layers improves due to their higher diversity degree. Based on this observation, we propose an iterative detection scheme in the sequence, which is a generalization of the Backward Iterative Cancellation algorithm proposed in [6].

IV. Backward Iterative Cancellation

To enhance the performance of V-BLAST, Benjebbour et al. proposed an iterative improvement of the successive detection scheme for flat fading channels [6]. In this paper we apply this Backward Iterative Cancellation (BIC) detection scheme to frequency-selective channels and therefore call it FS-BIC.

In the first iteration ($i = 1Gute$) of FS-BIC, the common FS-BLAST algorithm as proposed in Section III is applied (denoted by FS_BLAST($\mathbf{x}$)). The estimates of the first detected layer are denoted by[3] $\hat{s}^{(1)}_{(1)}$, whereas $\hat{s}^{(1)}_{(m)}$ depicts the output of stage m. After detecting all layers with the FS-BLAST algorithm, the interference of the layer detected last ($\hat{s}^{(1)}_{(n_T)}$) is subtracted from the received signal $\mathbf{x}$ to get a modified received signal vector

$$\acute{\mathbf{x}} = \mathbf{x} - \mathbf{H}_{(n_T)}\hat{\mathbf{s}}^{(1)}_{(n_T)} \tag{16}$$

and the corresponding coefficients of the channel matrix are set to zero. The detection of the remaining $n_T - 1$ layer denoted by FS_BLAST($\acute{\mathbf{x}}$) is repeated with the FS-BLAST algorithm and the output is denoted by $\hat{s}^{(2)}_{(1)}, \ldots, \hat{s}^{(2)}_{(n_T-1)}$. To renew the detection of the high-diversity layer (n_T), the influence of these new estimates are subtracted from the receive signal $\mathbf{x}$ and by equalizing

$$\mathbf{x}_{(n_T)} = \mathbf{x} - \sum_{m=1}^{n_T-1} \mathbf{H}_{(m)}\,\hat{\mathbf{s}}^{(2)}_{(m)} \tag{17}$$

we obtain $\hat{s}^{(2)}_{(n_T)}$.

In the third iteration step, the interference of $\hat{s}^{(2)}_{(n_T)}$ and $\hat{s}^{(2)}_{(n_T-1)}$ is subtracted from the received signal and consequently only the first $n_T - 2$ layers are detected by FS-BLAST. With these new replicas, the estimates for layer (n_T) and $(n_T - 1)$ are renewed, again. The whole iterative detection algorithm contains n_T iteration steps and is summarized in **Table I**.

A. Simulation Result

In this section we investigate the performance of the iterative FS-BIC algorithm for the system determined in Paragraph III-D with $n_T = 4$ and $n_R = 6$ antennas, QPSK modulation, a channel order of $L = 6$ and a FFF order of $N = 10$. **Fig. 6** shows the BER in each iteration for a receiver with a TIR order of $L_S = 4$. An improvement of about $2.3\,\mathrm{dB}$ is visible for the second iteration step compared to the initial detection stage. The improvement in the third iteration becomes smaller, but is still remarkable, whereas the fourth iteration leads only to a slight enhancement.

Fig. 7 shows the BER per iteration for the same system, but with a TIR order of only $L_S = 2$, which results in a decrease in computational complexity due to the smaller

[3]To simplify the description, we omit the time index k in the remainder of this paragraph.

TABLE I. FS-BIC ALGORITHM

- Initial FS_BLAST($\mathbf{x}$) of all layers to get estimates $\hat{s}^{(1)}_{(1)}, \ldots, \hat{s}^{(1)}_{(n_T)}$
- for $i = 2, \ldots, n_T$
  - Remove high diversity estimates $\hat{\mathbf{s}}^{(i-1)}_{(m)}$ from received signal $\mathbf{x}$ to achieve modified received signal for remaining layer $(1), \ldots, (n_T - i + 1)$

$$\hat{\mathbf{x}} = \mathbf{x} - \sum_{m=n_T-(i-2)}^{n_T} \mathbf{H}_{(m)}\, \hat{\mathbf{s}}^{(i-1)}_{(m)}$$

  - Apply FS_BLAST($\hat{\mathbf{x}}$) to renew estimates $\hat{s}^{(i)}_{(1)}, \ldots, \hat{s}^{(i)}_{(n_T-i+1)}$
  - for $l = n_T - (i-2), \ldots, n_T$
    - To improve detection of layer (l), remove renewed estimates of lower diversity layers and past estimates of higher diversity layers

$$\mathbf{x}_l = \mathbf{x} - \sum_{m=1}^{l-1} \mathbf{H}_{(m)}\, \hat{\mathbf{s}}^{(i)}_{(m)} - \sum_{m=l+1}^{n_T} \mathbf{H}_{(m)}\, \hat{\mathbf{s}}^{(i-1)}_{(m)}$$

    - Equalize $\mathbf{x}_l$ to achieve estimate $\hat{s}^{(i)}_{(l)}$
  - end
- end

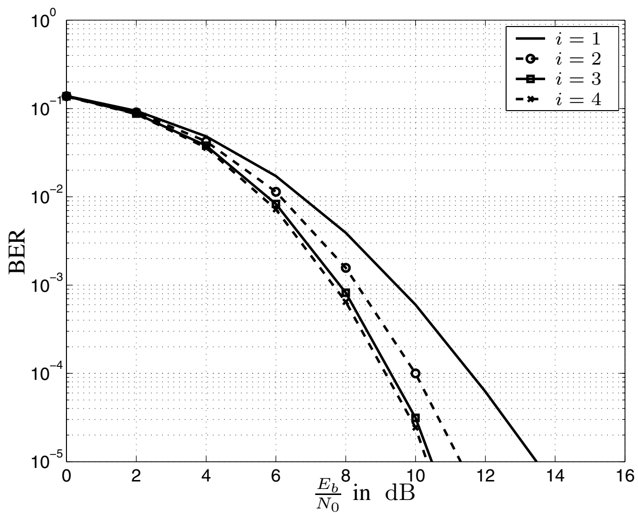

Fig. 6. BER per iteration i for channel order $L = 6$, FFF order $N = 10$ and target impulse response (TIR) order $L_S = 4$.

trellis. The performance is again improved by the iterative detection algorithm and only a difference of approximately $0.5\,\text{dB}$ related to the detector with $L_S = 4$ is visible for a BER of 10^{-5}.

It is worth to note, that the order of detection and the corresponding decision delay k_0 from the first iteration step can be retained for the subsequent steps. Consequently, the computational effort for these iteration steps is enormously smaller compared to the first step.

V. SUMMARY AND CONCLUSIONS

In this paper we described an extension of the backward iterative cancellation scheme for frequency-selective layered

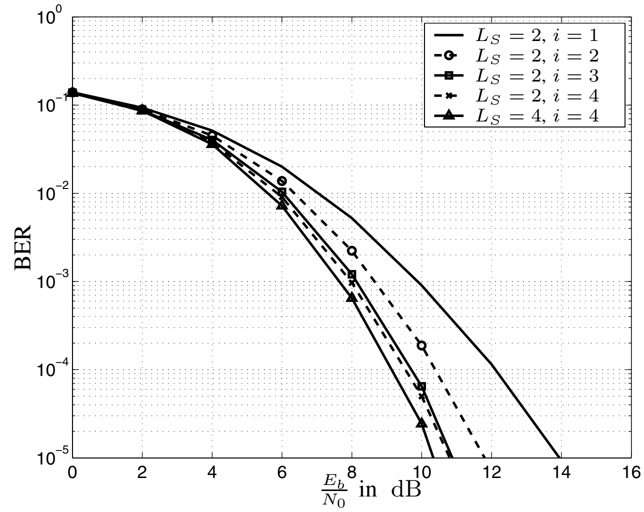

Fig. 7. BER per iteration for TIR order $L_S = 2$ and $L_S = 4$.

space-time systems. Subsequent to an initial detection, the performance is improved by successively cancelling high-diversity estimates and therefore enhancing the diversity degree. The performance enhancement of the proposed successive algorithms compared to FS-BLAST was discussed by simulation results.

Further improvements can be reached by implementing forward error correction codes and soft-input-soft-output equalizer in a turbo-kind scheme.

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