

Iterative Detection and Decoding with an Improved V-BLAST for MIMO-OFDM systems

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Abstract—This paper proposes an improved vertical Bell Labs Layered Space-Time (V-BLAST) with iterative detection and decoding (IDD) scheme for coded layered space-time architectures in MIMO-OFDM systems. For the iterative process, a low-complexity demapper is developed by making use of both nonlinear interference cancellation and linear filtering. Also a simple cancellation method based on hard decision is presented to reduce the overall complexity. Simulation results demonstrate that the proposed V-BLAST with IDD scheme offers the performance close to the optimal turbo-MIMO approach, while providing tremendous savings in computational complexity.

I. INTRODUCTION

As fourth generation (4G) wireless systems are being designed for offering high-quality multimedia services, the demand for higher bit rates will increase substantially compared to existing services. In order to satisfy this growing demand, considerable research attentions have been focused on improving the spectral efficiency in wireless channels. As increasing demand for higher bit rate leads to wideband communications, wireless channels become frequency selective. Multicarrier modulation realized by orthogonal frequency division multiplexing (OFDM) is well suited for such broadband applications.

The layered space-time architecture suggested in [1] has promised extremely high spectral efficiency by employing multiple antennas in multi-input multi-output (MIMO) systems. Among spatial-division multiplexing (SDM) techniques, V-BLAST [2] exhibits the best trade-off between performance and complexity. In order to further enhance the link performance, channel coding is usually employed for MIMO systems. However, traditional methods of symbol detection adopted in the V-BLAST do not work well when the channel coding is applied, since the decoder performance substantially suffers from the error propagation inherent in the decision feedback process. Thus, the receiver needs to compensate for the error propagation prior to the channel decoder.

Berrou *et al* developed the revolutionary iterative turbo receiver for decoding concatenated convolutional codes, which are capable of approaching the Shannon capacity in an additive white Gaussian noise (AWGN) channel [3]. Since then, the turbo decoding algorithm has been successfully extended to the turbo equalization by considering the intersymbol interference (ISI) channel as a rate-1 inner code [4]. The original

system introduced therein leveraged the ideas of the turbo decoding algorithms to the related problem of concatenation of equalization and decoding [5].

In parallel, by applying the turbo processing principle into the design of MIMO systems, Tonello [6] suggested an approach based on the serial concatenation of a convolutional encoder and a space-time signal constellation mapper, and showed that the structure in [6] approaches the optimal performance. We will refer to this design as a *turbo-MIMO* system in this paper. One of major drawbacks of such turbo-MIMO concepts is that demapping and decoding complexities increase exponentially with the number of antennas, and/or the number of bits per modulation symbol.

To reduce the complexity, several suboptimal MIMO detectors were proposed by making use of both nonlinear interference cancellation and linear minimum mean-square error (MMSE) filtering [7][8][9]. These suboptimal equalization and decoding processes utilize soft-input error control decoding by exchanging soft information between an equalizer and decoding algorithm.

In this paper, we propose a reduced complexity iterative detection and decoding (IDD) approach using decoder output and combined with decision error compensation for MIMO-OFDM systems. A similar IDD approach was proposed by Li *et al* [10]. In their work, it was shown that the IDD significantly improves the performance of V-BLAST. However, their study did not consider decision errors in the canceling process, therefore failed to fully utilize the potential of V-BLAST. In contrast, based on our strategy of the decision error compensation, we will show that a significant performance gain can be obtained using only hard decisions from decoder.

Through simulation results, we demonstrate that the performance of the proposed iterative scheme combined with the enhanced V-BLAST is very close to optimum turbo-MIMO systems, while providing tremendous savings in computational complexity.

The organization of the paper is as follows: Section II reviews a near optimal turbo-MIMO system based on bit-interleaved coded modulation. In Section III, we describe enhanced V-BLAST with detection error compensation. We introduce a new iterative detection and decoding scheme in Section IV. In Section V, the simulation results are presented comparing the proposed method with the conventional V-BLAST and turbo-MIMO. Finally, the paper is terminated

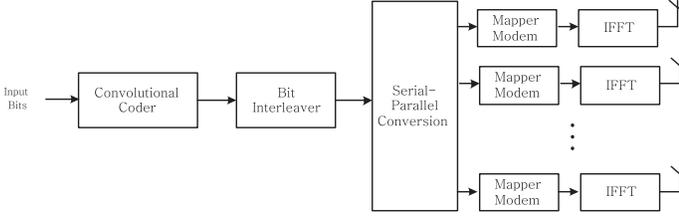


Fig. 1. Coded Layered Space-Time OFDM System Transmitter Model

with conclusions in Section VI.

II. SYSTEM DESCRIPTION

In this section, we consider a coded layered space-time OFDM system with N_t transmit antennas and N_r receive antennas. Fig. 1 shows the coded layered space-time transmitter architecture.

For a channel model, we make the following assumptions. Considering the time domain channel impulse response between the i th transmit and j th receive antenna, the frequency selective channel can be modeled as

$$h^{ji}(\tau) = \sum_{n=1}^L \bar{h}^{ji}(n) \delta(\tau - \tau_n)$$

where the channel coefficient $\bar{h}^{ji}(n)$ is the time domain channel impulse response at the n th time slot, $\delta(\cdot)$ is the Dirac delta function, and L denotes the number of channel taps. Here the channel coefficients $\bar{h}^{ji}(n)$ are independent complex Gaussian with zero mean. It follows that the channel frequency response can be expressed by

$$h_k^{ji} = \sum_{n=1}^L \bar{h}^{ji}(n) e^{-j2\pi k \tau_n / N_c T_s}$$

where N_c indicates the number of subchannels and T_s represents the sampling period.

Let us define the N_t -dimensional complex transmitted signal vector \mathbf{x}_k , and the N_r -dimensional complex received signal vector \mathbf{y}_k . Assuming proper cyclic prefix operation and the discrete Fourier transform (DFT), the received signal vector at the k th subcarrier can be written as

$$\mathbf{y}_k = \mathbf{H}_k \mathbf{x}_k + \mathbf{n}_k, \quad \text{for } k = 1, 2, \dots, N_c \quad (1)$$

where

$$\mathbf{H}_k = [\mathbf{h}_k^1 \ \mathbf{h}_k^2 \ \dots \ \mathbf{h}_k^{N_t}] = \begin{bmatrix} h_k^{11} & \dots & h_k^{1N_t} \\ \vdots & \ddots & \vdots \\ h_k^{N_r 1} & \dots & h_k^{N_r N_t} \end{bmatrix}. \quad (2)$$

Here \mathbf{h}_k^n denotes the n th column of the N_r by N_t channel matrix \mathbf{H}_k . The additive noise terms in \mathbf{n}_k in (1) are independent and identically-distributed complex Gaussian with variance σ_n^2 .

It is assumed that the covariance matrix of \mathbf{x}_k equals $E[\mathbf{x}_k \mathbf{x}_k^\dagger] = \sigma_s^2 \mathbf{I}_{N_t} = \frac{P}{N_t} \mathbf{I}_{N_t}$ where \mathbf{I}_{N_t} and P denotes

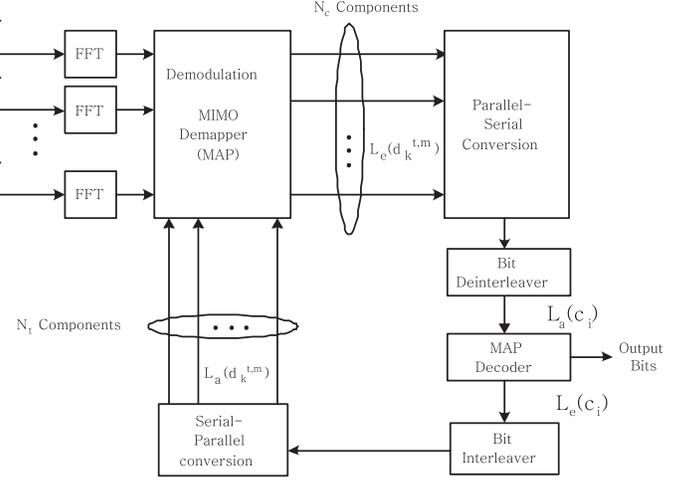


Fig. 2. The Receiver Structure of Near Optimum Turbo-MIMO in OFDM Systems

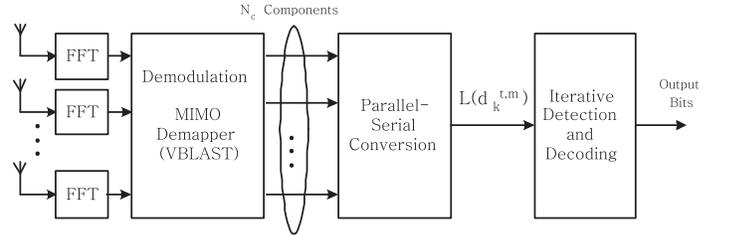


Fig. 3. The Receiver Structure of the Proposed Iterative Scheme

an identity matrix of size N_t and the total power of \mathbf{x}_k , respectively. And $E[\cdot]$ and $(\cdot)^\dagger$ indicate expectation and the complex-conjugate transpose, respectively.

Fig. 2 shows the receiver structure of the turbo-MIMO system [6][11]¹. The optimal MIMO demapper relying on the *a priori* information on the transmitted data from the maximum *a posteriori* probability (MAP) decoder produces the MAP estimates of the demapped bits for each subcarrier in the N_t transmitter streams. The estimation method for MAP as MIMO demapper in Fig. 2 is illustrated in [6][11].

The main problem with this approach is that when computing the log-likelihood ratio (LLR) values in the MIMO demapper, the search candidate number grows exponentially with the number of transmit antennas and/or bits per symbol. In the following sections, we propose an efficient IDD scheme for coded layered space-time architecture.

III. ENHANCED V-BLAST WITH DETECTION ERROR COMPENSATION

Fig. 3 shows the receiver structure for the proposed IDD scheme. The soft outputs are generated by the V-BLAST detector and sent to the IDD block for subsequent iterations. The detailed description of the IDD block in Fig. 3 will be

¹We refer to the receiver in Fig. 2 as "turbo-MIMO" instead of ST-BICM where N_t may be larger than N_r .

presented in Section IV. In this section, we briefly describe an enhanced V-BLAST operation [12] in Fig. 3.

A. Filtering in V-BLAST

In this section, we describe an enhanced receiver algorithm which takes decision errors into account. Let us define \hat{x}_k^n as the detected symbol for layer n . For simplicity, we assume that the ordering of the decisions $\{\hat{x}_k^1, \hat{x}_k^2, \dots, \hat{x}_k^{i-1}\}$ have been made according to the optimal detection order. Also, we define $\mathbf{x}_k^i \triangleq [x_k^i, x_k^{i+1}, \dots, x_k^{N_t}]^T$, $\mathbf{H}_k^{i:j} \triangleq [\mathbf{h}_k^i, \mathbf{h}_k^{i+1}, \dots, \mathbf{h}_k^j]$ and $\hat{\mathbf{x}}_k^{i-1} \triangleq [\hat{x}_k^1, \hat{x}_k^2, \dots, \hat{x}_k^{i-1}]^T$. In the conventional V-BLAST algorithm, the pre-detected symbol vector $\hat{\mathbf{x}}_k^{i-1}$ until step $i-1$ is canceled out from the received vector signal at step i , resulting in the modified received vector \mathbf{y}_k^i

$$\mathbf{y}_k^i = \mathbf{y} - \mathbf{H}_k^{1:i-1} \hat{\mathbf{x}}_k^{i-1} = \mathbf{H}_k^{i:N_t} \mathbf{x}_k^i + \mathbf{n}_k. \quad (3)$$

Here we assume that all the previous decisions are correct ($\hat{x}_k^n = x_k^n$ for $n = 1, 2, \dots, i-1$). However, considering the presence of decision errors, Equation (3) becomes

$$\mathbf{y}_k^i = \mathbf{H}_k^{i:N_t} \mathbf{x}_k^i + \mathbf{H}_k^{1:i-1} \hat{\mathbf{e}}_k^{i-1} + \mathbf{n}_k \quad (4)$$

where $\hat{\mathbf{e}}_k^{i-1} = [e_k^1, \dots, e_k^{i-1}]^T$ is defined with $e_k^n = x_k^n - \hat{x}_k^n$.

Using the new signal model in (4) and the MMSE criterion, the equalizer matrix \mathbf{G} is formulated to minimize the mean square value of the error defined as $\mathbf{e} = \mathbf{x}_k^i - \mathbf{G}\mathbf{y}_k^i$.

Finally, denoting $\alpha = \frac{\sigma_n^2}{\sigma_s^2}$, \mathbf{G} is obtained as

$$\mathbf{G} = \mathbf{H}_k^{i:N_t} \left(\mathbf{H}_k^{i:N_t} \mathbf{H}_k^{i:N_t} \dagger + \frac{1}{\sigma_s^2} \mathbf{H}_k^{1:i-1} \mathbf{Q}_{\hat{\mathbf{e}}_k^{i-1}} \mathbf{H}_k^{1:i-1} \dagger + \alpha \mathbf{I}_{N_r} \right)^{-1}$$

More details for the computation of the decision error covariance matrix $\mathbf{Q}_{\hat{\mathbf{e}}_k^{i-1}}$ is described in [12].

B. SOFT-OUTPUT DEMAPPER

Here, after making some assumptions on the output of MMSE equalization, we will derive the optimal soft bit metric which takes the detection errors into account. Let us define \mathbf{g}_t as the t th row of \mathbf{G} which corresponds to the equalizer for \hat{x}_k^t . Applying this equalizer vector into Equation (4) yields

$$\begin{aligned} \hat{x}_k^t &= \mathbf{g}_t \mathbf{H}_k^{i:N_t} \mathbf{x}_k^i + \mathbf{g}_t \mathbf{H}_k^{1:i-1} \hat{\mathbf{e}}_k^{i-1} + \mathbf{g}_t \mathbf{n}_k \\ &= \beta x_k^t + w \end{aligned} \quad (5)$$

where β and w are defined as $\beta = \mathbf{g}_t \mathbf{h}_k^t$ and $w = \sum_{\substack{j=i \\ j \neq t}}^N \mathbf{g}_t \mathbf{h}_k^j x_k^j + \mathbf{g}_t \mathbf{H}_k^{1:i-1} \hat{\mathbf{e}}_k^{i-1} + \mathbf{g}_t \mathbf{n}_k$, respectively.

Since those terms in w are assumed to be independent with each other, it can be shown in (5) that the variance of w is computed as

$$\begin{aligned} \sigma_w^2 &= \sigma_s^2 (\beta - \beta^2) \\ &= \sum_{\substack{j=i \\ j \neq t}}^N |\mathbf{g}_t \mathbf{h}_k^j|^2 \sigma_s^2 + \sum_{j=1}^{i-1} |\mathbf{g}_t \mathbf{h}_k^j|^2 E[|e_k^j|^2 | \hat{x}_k^j] + \sigma_n^2 \|\mathbf{g}_t\|^2. \end{aligned} \quad (6)$$

Note that the second term in (6) corresponds to decision errors until step $i-1$ and this term has been neglected in the conventional V-BLAST. The detailed computation for the LLR values is presented in [12].

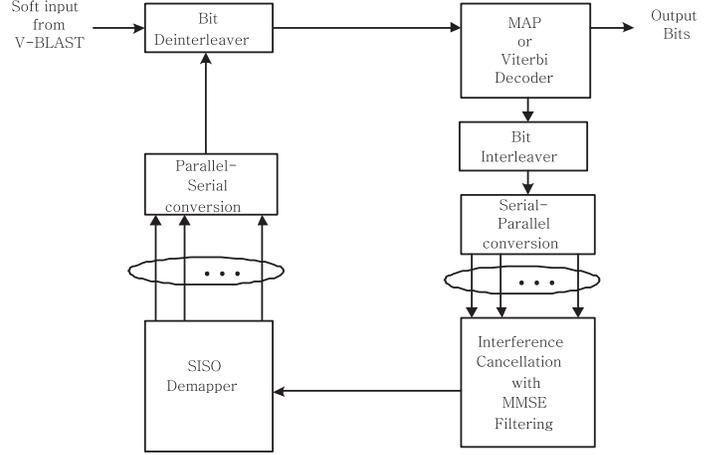


Fig. 4. The Proposed Iterative Detection and Decoding (IDD) Block

IV. ITERATIVE DETECTION AND DECODING

We now propose an IDD scheme combined with the enhanced V-BLAST for MIMO-OFDM systems. In the preceding section, the effect of error propagation is minimized by quantifying the decision error before decoding. In this section, we exploit the channel coding gain to further improve the performance. Fig. 4 illustrates the IDD block in Fig. 3 in detail. As shown in this diagram, the IDD block starts with LLR values generated by the improved V-BLAST described in Section III. Now we describe the IDD scheme based on the BCJR algorithm or Viterbi algorithm.

Compared with the turbo-MIMO receiver in Fig. 2, one big difference in the proposed IDD block is that MIMO demapper block is replaced by single-input single-output (SISO) demapper. The SISO demapper is employed to simplify the computation of the MIMO demapper, as this block dominates the complexity of the turbo-MIMO concept. Different from other suboptimal algorithms that iteratively exchange extrinsic information between each other, our proposed IDD scheme can be processed using only hard decisions. Thus, a complex BCJR decoder can be replaced by much simpler Viterbi decoder to reduce the computational complexity further.

As tentative decisions are available from the decoder output, those information can be used in the interference cancellation in the demapper block. In order to detect x_k^t , the hard decisions for all the other symbols $x_k^1, \dots, x_k^{t-1}, x_k^{t+1}, \dots, x_k^{N_t}$ are used to cancel the interference form \mathbf{y}_k in (1). Let \hat{x}_k^i be the hard decision symbol obtained from the decoder. Then, defining $\hat{\mathbf{x}}_k^t = [\hat{x}_k^1, \dots, \hat{x}_k^{t-1}, 0, \hat{x}_k^{t+1}, \dots, \hat{x}_k^{N_t}]$, the received signal \mathbf{y}_k is modified to

$$\underline{\mathbf{y}}_k^t = \mathbf{y}_k - \mathbf{H}_k \hat{\mathbf{x}}_k^t = \mathbf{H}_k \underline{\mathbf{x}}_k^t + \mathbf{n}_k \quad (7)$$

where $\underline{\mathbf{x}}_k^t = [e_k^1, \dots, e_k^{t-1}, x_k^t, e_k^{t+1}, \dots, e_k^{N_t}]$ with $e_k^i = x_k^i - \hat{x}_k^i$.

We apply an MMSE estimation filter \mathbf{w}_t^t to the modified received vector $\underline{\mathbf{y}}_k^t$ to get an estimation of the transmitted symbol x_k^t . Thus, we need to minimize the estimation error

defined as $e = x_k^t - \mathbf{w}_t \mathbf{y}_k^t$. By invoking the orthogonality principle, the MMSE weight \mathbf{w}_t can be obtained as

$$\mathbf{w}_t = \sigma_s^2 \mathbf{h}_k^t \left(\mathbf{H}_k \mathbf{Q}_{\mathbf{x}_k^t} \mathbf{H}_k^\dagger + \sigma_n^2 \mathbf{I}_{N_r} \right)^{-1} \quad (8)$$

where $\mathbf{Q}_{\mathbf{x}_k^t} = E[\mathbf{x}_k^t \mathbf{x}_k^{t\dagger}]$ is the covariance matrix of \mathbf{x}_k^t .

To simplify the computation, we neglect off-diagonal terms in the covariance matrix $\mathbf{Q}_{\mathbf{x}_k^t}$ since the elements in \mathbf{x}_k^t are independent under the assumption that an interleaver between the MIMO demapper and the decoder makes the bit information independent. Thus the computation for the covariance matrix reduces to

$$\mathbf{Q}_{\mathbf{x}_k^t} = \text{diag} \left[E[|e_k^1|^2], \dots, E[|e_k^{t-1}|^2], \sigma_s^2, E[|e_k^{t+1}|^2], \dots, E[|e_k^{N_t}|^2] \right].$$

The computation of $\mathbf{Q}_{\mathbf{x}_k^t}$ depends on whether the IDD employs Viterbi decoder or BCJR decoder. If Viterbi decoder is applied, we simply set $E[|e_k^i|^2]$ to 0 for $i \neq t$, so that the MMSE weight \mathbf{w}_t reduces to a simple matched filter.

On the other hand, when BCJR decoder is employed for better performance, we can compute $E[|e_k^i|^2]$ as

$$E[|e_k^i|^2] \triangleq E[|x_k^i - \hat{x}_k^i|^2 | L(d_k^{i,1}), \dots, L(d_k^{i, \log_2 M})] \text{ for } i \neq t \\ = \sum_{s \in \mathcal{S}} |s - \hat{x}_k^i|^2 \mathbf{P}(x_k^i = s | L(d_k^{i,1}), \dots, L(d_k^{i, \log_2 M}))$$

where $L(d_k^{i,m})$ is the LLR value for the bit $d_k^{i,m}$, which is the m th bit ($m = 1, 2, \dots, \log_2 M$) of the constellation symbol at the i th transmit antenna ($i = 1, 2, \dots, N_t$) at the k th subcarrier. Here M is the constellation size.

The probability of each symbol $\mathbf{P}(x_k^i = s | L(d_k^{i,1}), \dots, L(d_k^{i, \log_2 M}))$ can be computed by assuming that the output interference-plus-noise is Gaussian. Thus, the symbol probability can be directly obtained using the LLRs of the coded bits at the output of the decoder. Then it is straightforward to show that

$$\mathbf{P}(x_k^i = s \mid L(d_k^{i,1}), \dots, L(d_k^{i, \log_2 M})) \\ = \prod_{m=1}^{\log_2 M} \mathbf{P}(d_k^{i,m} = d_s^m | L(d_k^{i,m})) \quad (9)$$

where d_s^m denotes the m th bit of s . From the definition of the LLR, we have

$$\mathbf{P}(d_k^{i,m} = d_s^m | L(d_k^{i,m})) = \begin{cases} \frac{\exp(L(d_k^{i,m}))}{1 + \exp(L(d_k^{i,m}))} & \text{for } d_s^m = 0 \\ \frac{\exp(-L(d_k^{i,m}))}{1 + \exp(-L(d_k^{i,m}))} & \text{for } d_s^m = 1. \end{cases}$$

After computing $E[|e_k^i|^2]$, \mathbf{w}_t can be obtained in (8) with $\mathbf{Q}_{\mathbf{x}_k^t}$. Applying the MMSE filter \mathbf{w}_t to \mathbf{y}_k^t in (7) yields

$$z_k^t = \mathbf{w}_t \mathbf{y}_k^t = \alpha x_k^t + v.$$

Finally in the SISO demapper block, the soft output bit for x_k^t can be computed as

$$L(d_k^{t,m}) \triangleq \log \frac{P[d_k^{t,m} = 0 | z_k^t]}{P[d_k^{t,m} = 1 | z_k^t]} \\ = \log \frac{\sum_{s \in \mathcal{S}_0^m} \exp\left(-\frac{\|z_k^t - \alpha s\|^2}{\sigma_v^2}\right)}{\sum_{s \in \mathcal{S}_1^m} \exp\left(-\frac{\|z_k^t - \alpha s\|^2}{\sigma_v^2}\right)}$$

where $\alpha = \mathbf{w}_t \mathbf{h}_k^t$ and $\sigma_v^2 = \sigma_s^2(\alpha - \alpha^2)$. Note that we assume that symbols s are equally probable.

In this SISO demapper, the number of candidates to compute the above LLR values is only M regardless of antenna numbers, while the MIMO demapper in [11] for the turbo-MIMO requires M^{N_t} . Thus, the proposed IDD reduces the computational complexity substantially.

V. SIMULATION RESULTS

In this section, we illustrate the performance of the proposed IDD scheme combined with the enhanced V-BLAST in MIMO-OFDM systems. The OFDM modulation employs 64 point FFT. We adopt a channel model with a 5-tap power delay profile having exponentially decayed Rayleigh fading characteristics. Also, one frame is assumed to consist of one OFDM symbol for simplicity, and then the size of a random interleaver is determined by $N_c \cdot N_t \cdot \log_2 M$. In this simulation, binary convolutional codes with polynomials (23,35) is used for the code of rate 1/2.

In order to demonstrate, we compare the following systems:

- *The proposed IDD with BCJR: Applying the enhanced V-BLAST with BCJR decoder in the IDD block*
- *The proposed IDD with VA: Applying the enhanced V-BLAST with Viterbi decoder in the IDD block*
- *The conventional IDD: Employing the conventional V-BLAST with Viterbi decoder in the IDD block*

Fig. 5 shows the FER of different iterative schemes for 16QAM with $N_t = N_r = 2$. As can be shown in this plot, the proposed IDD schemes with VA and BCJR outperform the conventional IDD by 3 dB and 4 dB, respectively, at 1% FER at the fourth iteration. Comparing the proposed IDD with VA and the conventional IDD, we can see that the use of the enhanced V-BLAST provides an improvement of 3 dB in the IDD process. More interestingly, Fig. 5 indicates that the proposed IDD performs within 1-2 dB of the turbo-MIMO while achieving significant complexity reduction. Fig. 6 depicts the performance of MIMO-OFDM systems with the increased number of transmit and receive antennas. In the 4 transmit antenna case, the performance gap between the turbo-MIMO and the proposed iterative scheme is less than 1 dB.

In Fig. 7, we consider the proposed IDD scheme for a high spectral efficiency. This plot presents the performance comparison in the 4 by 4 MIMO-OFDM system with 16QAM which leads to the spectral efficiency of 8 bps/Hz. This extreme example is impracticable for the turbo-MIMO because the MIMO demapper is simply infeasible. In this case, the

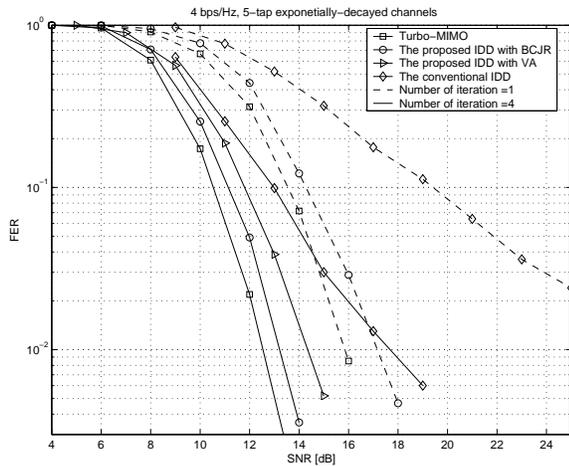


Fig. 5. Frame error probability of different iterative decoding schemes with 16QAM and $N_t = N_r = 2$

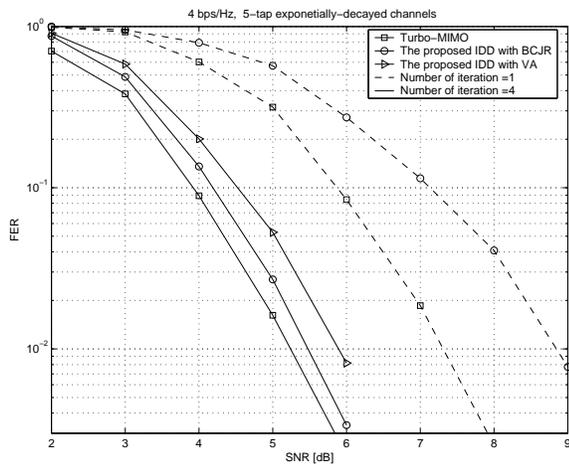


Fig. 6. Frame error probability of different iterative decoding schemes with 4QAM and $N_t = N_r = 4$

optimum MIMO demapper needs to search $16^4 = 65536$ candidates to compute the LLRs, whereas for the proposed scheme only 16 candidates are to be considered. Even with this significant reduction in the candidate number, the performance is within 1-2 dB of the turbo-MIMO case. This feature of the proposed IDD scheme is quite important, as the target spectral efficiency in future wireless systems becomes higher.

VI. CONCLUSION

In this paper, we have proposed pragmatic schemes for the layered space-time architectures in MIMO-OFDM systems. We have demonstrated the efficacy of a simple iterative MIMO detector based on the interference cancellation and MMSE filtering. Simulation results indicate that the performance of the proposed iterative scheme is just less than 1 dB away from the optimum turbo-MIMO for all the simulation configurations with remarkably reduced complexity. The simulation results

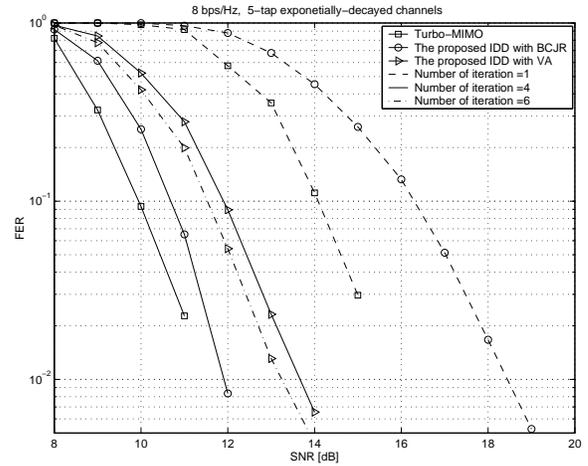


Fig. 7. Frame error probability of different iterative decoding schemes with 16QAM and $N_t = N_r = 4$

confirm that by properly treating the decision errors in interference cancellation, the detrimental effects of error propagation can be almost completely overcome by the proposed iterative processing.

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