Iterative Detection and Decoding With an Improved V-BLAST for MIMO-OFDM Systems

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***Abstract*—Multiple-input–multiple-output (MIMO) systems provide a very promising means to increase the spectral efficiency for wireless systems. By using orthogonal frequency-division multiplexing (OFDM), wideband transmission can be achieved over frequency-selective fading radio channels. First, in this paper, we introduce an improved vertical Bell Labs layered space–time (V-BLAST) receiver which takes the decision errors into account. Second, we propose an 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 minimum mean-square error filtering. Also, a simple cancellation method based on hard decision is presented to reduce the overall complexity. Simulation results demonstrate that the proposed IDD scheme combined with the improved V-BLAST performs almost as well as the optimal turbo-MIMO approach, while providing tremendous savings in computational complexity.**

***Index Terms*—Iterative detection and decoding (IDD), multiple-input–multiple-output (MIMO) systems, orthogonal frequency-division multiplexing (OFDM), vertical Bell Labs layered space–time (V-BLAST).**

I. Introduction

As FOURTH-GENERATION (4G) wireless systems are being designed for offering high-quality multimedia services, which include voice, data, and video, the required data rates will increase substantially compared with existing services. In order to satisfy this growing demand, considerable research attentions have been focused on improving the spectral efficiency in wireless channels. Reaching this goal requires an efficient usage of the limited wireless bandwidth resource. Normally, wireless channels exhibit a number of severe impairments such as intersymbol interference (ISI) and fading, which lead to great challenges in next-generation wireless communications.

As higher bit rates lead to wideband communications, wireless channels become frequency selective. Multicarrier modulation realized by orthogonal frequency-division multiplexing (OFDM) is well suited for such broadband applications [1], [2]. The OFDM modulation technique divides the total available bandwidth into a number of equally spaced frequency bands. By applying a proper cyclic prefix, individual subchannels are transformed to exhibit flat fading channel characteristics.

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The layered space–time architecture suggested in [3] has promised extremely high spectral efficiency in multiple-input–multiple-output (MIMO) systems. Among such spatial-division multiplexing (SDM) techniques, vertical Bell Labs layered space–time (V-BLAST) [4] exhibits the best tradeoff between performance and complexity. The V-BLAST uses a combination of linear and nonlinear detection techniques: first nulling out the interference from yet undetected signals, and then canceling out the interference using already detected signals [4]. However, the 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 [5]. Since then, the turbo decoding algorithm has been successfully extended to the turbo equalization by considering the ISI channel as a rate-1 inner code [6]. The original system introduced therein leveraged the ideas of the turbo decoding algorithms to the related problem of concatenation of equalization and decoding [7]. In this case, the maximum *a posteriori* probability (MAP)-based techniques are used for both equalization and decoding process. Since the MAP-based solutions often suffer from high computational load, simpler approaches have been proposed to replace the MAP receiver by combining equalization based on the linear filtering with decoding [8], [9].

In parallel, considerable research interests have been focused on techniques and algorithms which realize various benefits of turbo principle for MIMO systems. Applying the turbo processing principle, Tonnelo [10] suggested a space–time coding approach based on the serial concatenation of a convolutional encoder and a signal constellation mapper. The near-optimum receiver consists of a MIMO demapper followed by MAP decoder. We will refer to this design as a *turbo-MIMO* system in this paper. Similarly, Haykin *et al.* described the essence of turbo-MIMO wireless systems with an emphasis on SDM [11]. Also, several versions of MIMO schemes combined with turbo principle have been proposed in the literatures [12], [13]. One of major drawbacks of such turbo-MIMO concepts is that demapping and decoding complexities increase exponentially with the number of transmit/receive antennas, the number of bits per symbol and/or the code constraint length.

To reduce the complexity, several suboptimal MIMO detectors were proposed by making use of nonlinear interference cancellation and linear minimum mean-square error (MMSE) filtering. Properties of such a nonlinear interference suppressor are presented in [8] for code-division multiple-access (CDMA) channels. In [14] and [15], low-complexity turbo equalization algorithms have been proposed for frequency-selective MIMO channels. These suboptimal equalization and decoding processes utilize soft-input error control decoding by exchanging soft information between an equalizer and decoding algorithm. Also, the list sphere decoder has been applied to reduce the computational complexity in the demapper [16]. Similarly, a simplified receiver with tentative decisions is proposed for a turbo-MIMO [17].

In this paper, we consider MIMO-OFDM systems with iterative process, which combine all the schemes mentioned above. First, we introduce an enhanced V-BLAST detection algorithm which takes the error propagation effect into account. By including the decision errors into the filtering formulation, an improved detection performance is achieved. Also, we derive the optimal soft bit demapper for channel decoding by considering the error propagation.

Second, employing the enhanced V-BLAST as a front-end receiver for MIMO-OFDM systems, we propose an iterative detection and decoding (IDD) approach which further improves the detection performance by utilizing decoder output. A number of schemes have been proposed to reduce the effect of error propagation for ISI channels using decoder output [18], [19]. Interestingly, it has been shown in [19] that for the single carrier systems, the joint convolutional coding and decision feedback equalization (DFE) technique can perform within 1–2 dB of an ideal DFE without error propagation. A similar IDD approach was proposed by Li *et al.* [20] for V-BLAST. In their work, it was shown that the IDD significantly improves the performance of V-BLAST with so-called "horizontal coding," where separate channel encoders are connected to each transmit antenna. However, their study did not consider decision errors in the canceling process, therefore, failed to fully utilize the potential of V-BLAST. In contrast, we focus on "vertical coding" structure where a single channel encoder supports all transmit antennas. In this paper, 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 a decoder.

Through simulation results, we demonstrate that the performance of the proposed IDD scheme combined with the enhanced V-BLAST is very close to near-optimum turbo-MIMO systems, while providing tremendous savings in computational complexity. Also, unlike conventional V-BLAST systems which require the number of receive antennas to be at least the same as that of transmit antennas, it will be shown that the proposed scheme works well even when there are more transmit antennas than receive antennas with an aid from the IDD process.

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 present a channel model and derive enhanced detector design criteria. In addition, we describe how to calculate the soft bit information when error propagation is taken into account. 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 other schemes. Finally, this paper is terminated with conclusions in Section VI.

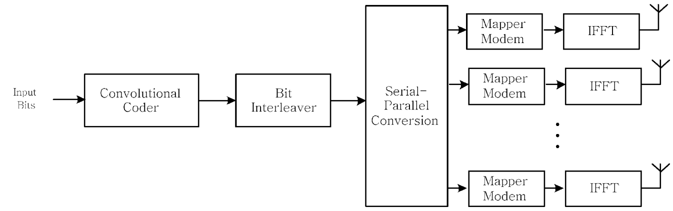

Fig. 1. Coded layered space–time OFDM system transmitter model.

II. System Descriptions

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. At the transmitter, the information is first encoded using a linear forward error correction (FEC) code. We will assume that a convolutional code is used for FEC codes. The encoded data stream is then interleaved, symbol-mapped, and transmitted via individual transmit antennas.

For a channel model, we make the following assumptions. Considering the time-domain channel impulse response between the ith transmit and jth 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 nth time slot, $\delta(\cdot)$ represents 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. The assumption of independent paths holds if antennas at each end of the communication links are separated by more than half a wavelength [3]. 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 kth subcarrier can be written as

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

where

$$\mathbf{H}_k = \left[\mathbf{h}_k^1\ \mathbf{h}_k^2, \ldots, \mathbf{h}_k^{N_t}\right] = \begin{bmatrix} h_k^{11} & \cdots & h_k^{1N_t} \\ \vdots & \ddots & \vdots \\ h_k^{N_r 1} & \cdots & h_k^{N_r N_t} \end{bmatrix}$$

$$\mathbf{n}_k = \begin{bmatrix} n_k^1 \\ \vdots \\ n_k^{N_r} \end{bmatrix}.$$

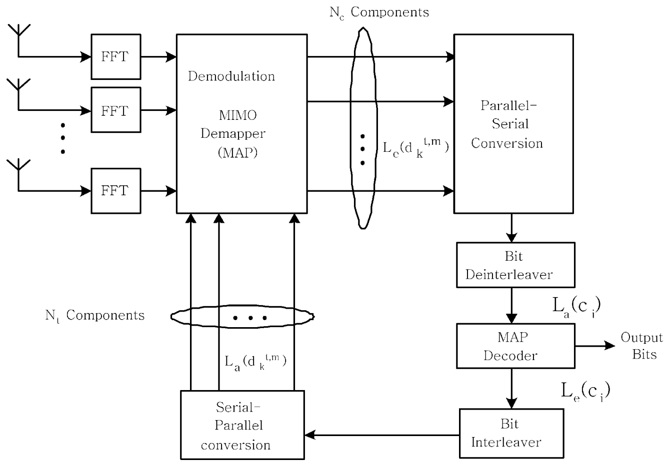

Fig. 2. Receiver structure of turbo-MIMO in OFDM systems.

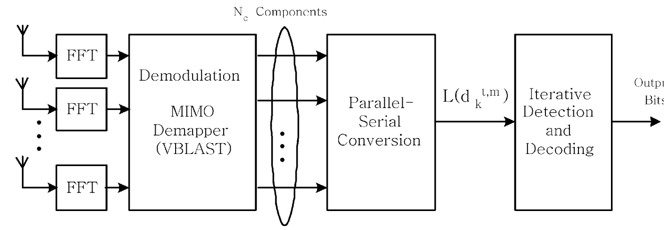

Fig. 3. Receiver structure of the proposed IDD scheme.

Here, $\mathbf{h}_k^n$ denotes the nth column of the $N_r \times N_t$ channel matrix $\mathbf{H}_k$ and the additive noise terms in $\mathbf{n}_k$ are independent and identically distributed complex Gaussian with variance σ_n^2.

Under the constraints that the average total power of $\mathbf{x}_k$ is assumed to be P and the instantaneous channel state is not known at the transmitter, the transmitted signal power needs to be distributed equally over the N_t transmit antennas with the same variance σ_s^2 in order to achieve the maximum capacity [21]. Thus, 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 $E[\cdot]$ and $(\cdot)^\dagger$ indicate expectation and the complex-conjugate transpose, respectively, and $\mathbf{I}_{N_t}$ denotes an identity matrix of size N_t.

In the following, we briefly describe a near-optimum receiver for turbo-MIMO systems to compare the performance of our proposed IDD scheme. Fig. 2 shows the receiver structure of the turbo-MIMO system[1] [22], which exploits the idea of space–time bit-interleaved coded modulation (ST-BICM) [23], [24]. The ST-BICM schemes achieve diversity gains on fading channels with higher order modulation constellation combined with conventional binary convolutional codes through the use of iterative processing between the MIMO demapper and the MAP decoder.

For a wideband systems, the OFDM demodulator is applied for each transmitter stream over N_c parallel subchannels. The optimal MIMO demapper relying on the *a priori* information on the transmitted data from the MAP decoder produces the MAP estimates of the demapped bits for each subcarrier in the N_t transmitter streams. We refer to this estimation method as MIMO demapper in Fig. 2 [22].

Let $d_k^{i,m}$ be the mth bit ($m = 1, 2, \ldots, \log_2 M$) of the constellation symbol at the ith transmit antenna ($i = 1, 2, \ldots, N_t$) at the kth subcarrier, where M is the constellation size. We denote $L(d_k^{i,m})$ as the log-likelihood ratio (LLR) value for the bit $d_k^{i,m}$. Also, $L_e(d_k^{i,m})$ is defined as an extrinsic LLR value obtained by subtracting the input *a priori* LLR value $L_a(d_k^{i,m})$ from the output $L(d_k^{i,m})$. Then, the LLR values are given by

$$L\left(d_k^{i,m}\right) = \log \frac{P\left(d_k^{i,m} = +1\right)}{P\left(d_k^{i,m} = -1\right)}. \tag{2}$$

Let $\mathbb{S}_d^{i,m}, d = +1$ (for bit 0) or -1 (for bit 1), be a set of all symbol vectors with $d_k^{i,m} = d$. The number of elements in such a set is $2^{N_t \log_2 M - 1}$. The LLR in (2) conditioned on the channel state information is represented as

$$\log \frac{P\left(d_k^{i,m} = +1 | \mathbf{y}_k, \mathbf{H}_k\right)}{P\left(d_k^{i,m} = -1 | \mathbf{y}_k, \mathbf{H}_k\right)} = \log \frac{\sum_{\mathbf{x}_k \in \mathbb{S}_{+1}^{i,m}} p(\mathbf{x}_k, \mathbf{y}_k, \mathbf{H}_k)}{\sum_{\mathbf{x}_k \in \mathbb{S}_{-1}^{i,m}} p(\mathbf{x}_k, \mathbf{y}_k, \mathbf{H}_k)}. \tag{3}$$

The joint probability density in (3) is then related to [22]

$$p(\mathbf{x}_k, \mathbf{y}_k, \mathbf{H}_k) \propto \exp\left(-\frac{1}{\sigma_n^2}\|\mathbf{y}_k - \mathbf{H}_k\mathbf{x}_k\|^2 + \frac{1}{2}\mathbf{d}_k^T \mathbf{L}_k^a\right)$$

where $\mathbf{d}_k$ and $\mathbf{L}_k^a$ are column vectors of length $N_t \log_2 M$ comprised of $d_k^{i,m}$ and the *a priori* LLR values $L_a(d_k^{i,m})$ from the MAP decoder, respectively, and $(\cdot)^T$ denotes the transpose. The elements of the *a priori* likelihood ratio vector $\mathbf{L}_k^a$ are set to zero for the first pass.

The MAP decoder shown in Fig. 2 employs the BCJR algorithm [25] which generates *a posteriori* probability. Once the BCJR algorithm obtains the soft information from the MIMO demapper, the MAP decoder, in turn, produces its own soft output. This soft decision from the decoder is then properly interleaved and fed back to the MIMO demapper, which completes one iteration cycle.

The main problem with this approach is that when computing LLR values in the MIMO demapper in (3), the search candidate number M^{N_t} of elements in $\mathbb{S}_d^{i,m}$ 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 Error Compensation

Fig. 3 shows the receiver structure for the proposed IDD scheme. The soft outputs are generated by a V-BLAST detector and sent to the IDD block for subsequent iterations. The detailed description of the IDD block in Fig. 3 will be presented in Section IV. In this section, we derive an enhanced V-BLAST which takes decision errors into account.

[1] We refer to the receiver in Fig. 2 as "turbo-MIMO" instead of ST-BICM, where N_t is typically larger than N_r.

A. Filtering in V-BLAST

We start with a new signal model of the layered space–time OFDM system which includes the error propagation effect. Let us define the transmitted symbols as a signal vector $\mathbf{x}_k = [x_k^1\ x_k^2\ \cdots x_k^{N_t}]^T$, where x_k^n represents the symbol transmitted from the nth antenna at the kth subchannel. We denote $\hat{x}_k^n$ as the detected symbol for layer n. For simplicity, we assume that the ordering of the decisions $\{\hat{x}_k^1\ \hat{x}_k^2 \cdots \hat{x}_k^{i-1}\}$ have been made according to the optimal detection order as in [4]. Also, we define $\mathbf{x}_k^i \triangleq [x_k^i\ x_k^{i+1} \cdots x_k^{N_t}]^T$, $\mathbf{H}_k^{i:j} \triangleq [\mathbf{h}_k^i\ \mathbf{h}_k^{i+1} \cdots \mathbf{h}_k^j]$, and $\hat{\mathbf{x}}_k^{i-1} \triangleq [\hat{x}_k^1\ \hat{x}_k^2 \cdots \hat{x}_k^{i-1}]^T$. In the conventional V-BLAST algorithm, the predetected 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 $\mathbf{y}_k^i$

$$\begin{aligned}\mathbf{y}_k^i &= \mathbf{y} - \mathbf{H}_k^{1:i-1}\hat{\mathbf{x}}_k^{i-1}\\ &= \mathbf{H}_k^{i:N_t}\mathbf{x}_k^i + \mathbf{n}_k. \end{aligned} \tag{4}$$

Here, we assume that all the previous decisions are correct ($\hat{x}_k^n = x_k^n$ for $n = 1, 2, \ldots, i-1$).

At step i, in order to detect x_k^t, the remaining undetected symbols $[x_k^i, \ldots, x_k^{t-1}, x_k^{t+1}, \ldots, x_k^{N_t}]$ are treated as interferers. In such cases, error propagation is inevitable and this results in performance degradation. Equation (4) is accurate only when the predetected symbol vector $\hat{\mathbf{x}}_k^{i-1}$ is correct. However, in the presence of decision errors, (4) becomes

$$\begin{aligned}\mathbf{y}_k^i &= \mathbf{H}_k^{i:N_t}\mathbf{x}_k^i + \mathbf{H}_k^{1:i-1}\hat{\mathbf{e}}_k^{i-1} + \mathbf{n}_k\\ &= \mathbf{H}_k \begin{bmatrix}\hat{\mathbf{e}}_k^{i-1}\\ \mathbf{x}_k^i\end{bmatrix} + \mathbf{n}_k\end{aligned} \tag{5}$$

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

In what follows, we will derive the nulling matrix based on the MMSE criterion which accounts for the decision errors based on (5). The equalizer matrix $\mathbf{G}$ is formulated to minimize the mean square value of the error defined as

$$\mathbf{e} = \mathbf{x}_k^i - \mathbf{G}\mathbf{y}_k^i. \tag{6}$$

Then, $\mathbf{G}$ can be obtained by invoking the orthogonality principle [26]. Thus, it follows:

$$E[\mathbf{e}\mathbf{y}_k^{i\dagger}] = E[(\mathbf{x}_k^i - \mathbf{G}\mathbf{y}_k^i)\mathbf{y}_k^{i\dagger}] = 0$$

which gives rise to

$$\mathbf{G} = \mathbf{Q}_{\mathbf{x}_k^i\mathbf{y}_k^i}\mathbf{Q}_{\mathbf{y}_k^i}^{-1} \tag{7}$$

where we define the covariance matrix as $\mathbf{Q}_{\mathbf{AB}} = E[\mathbf{AB}^\dagger]$ and $\mathbf{Q}_{\mathbf{A}} = E[\mathbf{AA}^\dagger]$.

Finally, denoting $\alpha = \sigma_n^2/\sigma_s^2$, $\mathbf{G}$ is obtained from (5) and (7) as

$$\begin{aligned}\mathbf{G} &= \sigma_s^2 \mathbf{H}_k^{i:N_t\dagger}\left(\mathbf{H}_k \begin{bmatrix}\mathbf{Q}_{\hat{\mathbf{e}}_k^{i-1}} & \mathbf{0}\\ \mathbf{0} & \sigma_s^2\mathbf{I}_{N_t-i+1}\end{bmatrix}\mathbf{H}_k^\dagger + \sigma_n^2\mathbf{I}_{N_r}\right)^{-1}\\ &= \mathbf{H}_k^{i:N_t\dagger}\left(\mathbf{H}_k^{i:N_t}\mathbf{H}_k^{i:N_t\dagger}\right.\\ &\qquad \left. + \frac{1}{\sigma_s^2}\mathbf{H}_k^{1:i-1}\mathbf{Q}_{\hat{\mathbf{e}}_k^{i-1}}\mathbf{H}_k^{1:i-1\dagger} + \alpha\mathbf{I}_{N_r}\right)^{-1}\end{aligned} \tag{8}$$

where we have used the fact that $\mathbf{Q}_{\mathbf{x}_k^i} = \sigma_s^2\mathbf{I}_{N_t-i+1}$ and $\mathbf{Q}_{\mathbf{n}_k} = \sigma_n^2\mathbf{I}_{N_r}$. Here, the decision error covariance matrix $\mathbf{Q}_{\hat{\mathbf{e}}_k^{i-1}}$ of size $i-1$ is defined as

$$\mathbf{Q}_{\hat{\mathbf{e}}_k^{i-1}} = \begin{bmatrix} E[|e_k^1|^2|\hat{x}_k^1] & \cdots & E[e_k^1 e_k^{i-1*}|\hat{x}_k^1, \hat{x}_k^{i-1}]\\ \vdots & \ddots & \vdots\\ E[e_k^{i-1}e_k^{1*}|\hat{x}_k^{i-1}, \hat{x}_k^1] & \cdots & E[|e_k^{i-1}|^2|\hat{x}_k^{i-1}]\end{bmatrix} \tag{9}$$

where * denotes the complex conjugate. Here, the conditional expectation $E[e_k^m e_k^{n*}|\hat{x}_k^m, \hat{x}_k^n]$ is used to indicate that the errors e_k^m and e_k^n are made from incorrect decisions $\hat{x}_k^m \neq x_k^m$ and $\hat{x}_k^n \neq x_k^n$, respectively. In Section III-B, we will show how to obtain the diagonal elements $E[|e_k^m|^2|\hat{x}_k^m]$. More details on the computation of the covariance matrix $\mathbf{Q}_{\hat{\mathbf{e}}_k^{i-1}}$ is described in [27].[2] Note that the off-diagonal entries $E[e_k^m e_k^{n*}|\hat{x}_k^m, \hat{x}_k^n]$ with $m \neq n$ can be neglected due to symmetry in constellations by approximating the decision error as a Gaussian process. Simulations show that when off-diagonal elements of $\mathbf{Q}_{\hat{\mathbf{e}}_k^{i-1}}$ are neglected, the performance loss due to this approximation is less than 0.5 dB compared with the case without any simplification. Therefore, from now on, we ignore the off-diagonal elements of $\mathbf{Q}_{\hat{\mathbf{e}}_k^{i-1}}$ to reduce the complexity. With this simplification, $\mathbf{Q}_{\hat{\mathbf{e}}_k^{i-1}}$ then reduces to

$$\mathbf{Q}_{\hat{\mathbf{e}}_k^{i-1}} = \mathrm{diag}[E[|e_k^1|^2|\hat{x}_k^1], \ldots, E[|e_k^{i-1}|^2|\hat{x}_k^{i-1}] \tag{10}$$

where $\mathrm{diag}(\cdot)$ indicates a diagonal matrix.

It is straightforward to show that the proposed matrix $\mathbf{G}$ reduces to the conventional MMSE matrix when previously detected decisions are assumed to be perfect so that no error propagation occurs: That is, $\mathbf{Q}_{\hat{\mathbf{e}}_k^{i-1}} = \mathbf{0}$ in (8), so we have

$$\begin{aligned}\mathbf{G} &= \mathbf{H}_k^{i:N_t\dagger}\left(\mathbf{H}_k^{i:N_t}\mathbf{H}_k^{i:N_t\dagger} + \alpha\mathbf{I}_{N_r}\right)^{-1}\\ &= \left(\mathbf{H}_k^{i:N_t\dagger}\mathbf{H}_k^{i:N_t} + \alpha\mathbf{I}_{N_t-i+1}\right)^{-1}\mathbf{H}_k^{i:N_t\dagger}.\end{aligned} \tag{11}$$

In the simulation section, we will show that the proposed MMSE equalization matrix improves the performance substantially.

B. Soft-Output Demapper

It is well known that a significant improvement in performance is obtained by applying soft decoding [28]. With some assumptions on the output of the MMSE equalization, we will derive the optimal soft bit metric which takes the detection errors into account.

After matrix manipulations, the covariance matrix $\mathbf{Q}_{\mathbf{e}}$ of the estimation error $\mathbf{e} = \mathbf{x}_k^i - \mathbf{G}\mathbf{y}_k^i$ can be computed as

$$\mathbf{Q}_{\mathbf{e}} = E[\mathbf{e}\mathbf{e}^\dagger] = \sigma_s^2(\mathbf{I}_{N_t-i+1} - \mathbf{G}\mathbf{H}_k^{i:N_t}). \tag{12}$$

Diagonal elements of $\mathbf{Q}_{\mathbf{e}}$ then represent the mean-squared error (MSE) value for each to-be-detected symbol. Thus, the subsequent detection order is determined by the location of the smallest diagonal element of $\mathbf{Q}_{\mathbf{e}}$, which is equivalent to the position of the largest diagonal entry of $\mathbf{G}\mathbf{H}_k^{i:N_t}$ in (12).

[2]The structure studied in [27] is the horizontal coding, whereas this paper focuses on the vertical coding structure.

Denote t as the position where the MSE is minimum in $\mathbf{Q_e}$. In other words, $\hat{x}_k^t$ is chosen as a decision at step i ($i \leq t \leq N_t$). Let us define $\mathbf{g}_t$ as the tth row of $\mathbf{G}$ which corresponds to the equalizer for $\hat{x}_k^t$. Applying this equalizer vector into (5) yields

$$\begin{aligned}\tilde{x}_k^t &= \mathbf{g}_t\mathbf{H}_k^{i:N_t}\mathbf{x}_k^i + \mathbf{g}_t\mathbf{H}_k^{1:i-1}\hat{\mathbf{e}}_k^{i-1} + \mathbf{g}_t\mathbf{n}_k \\ &= \mathbf{g}_t\mathbf{h}_k^t x_k^t + \sum_{\substack{j=i \\ j\neq t}}^{N} \mathbf{g}_t\mathbf{h}_k^j x_k^j + \mathbf{g}_t\mathbf{H}_k^{1:i-1}\hat{\mathbf{e}}_k^{i-1} + \mathbf{g}_t\mathbf{n}_k \\ &= \beta x_k^t + w \end{aligned} \tag{13}$$

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.

It is shown in [29] that the error probability for the MMSE detector can be well approximated by assuming that the distribution of the residual interference-plus-noise is Gaussian. Therefore, we assume that the terms in w make complex Gaussian distribution. Since those terms in w are assumed to be independent with each other, the variance of w is computed as

$$\sigma_w^2 = \sum_{\substack{j=i \\ j\neq t}}^{N} |\mathbf{g}_t\mathbf{h}_k^j|^2\sigma_s^2 + \sum_{j=1}^{i-1} |\mathbf{g}_t\mathbf{h}_k^j|^2 E[|e_k^j|^2|\hat{x}_k^j] + \sigma_n^2\|\mathbf{g}_t\|^2. \tag{14}$$

The above expression for the variance is given only to show the effect of the error propagation on the interference-plus-noise terms. From (6), (12), and (13), we can compute σ_w^2 simply by [8]

$$\sigma_w^2 = \sigma_s^2(\beta - \beta^2). \tag{15}$$

Note that the second term in (14) corresponds to increased variance due to decision errors until step $i-1$ and that this term has been neglected in the conventional V-BLAST. We will show in Section V that including this error propagation term has a noticeable effect on the soft decoding performance.

Now, we illustrate how to compute $E[|e_k^t|^2|\hat{x}_k^t]$ in (10) using β in (13). For the next step $i+1$, we need to calculate the conditional expected value $E[|e_k^t|^2|\hat{x}_k^t]$ which can be obtained as

$$E[|e_k^t|^2|\hat{x}_k^t] = \sum_{s\in\mathcal{N}_{\hat{x}_k^t}} |s - \hat{x}_k^t|^2\mathbf{P}(x_k^t = s|\hat{x}_k^t) \tag{16}$$

where $\mathcal{N}_{\hat{x}_k^t}$ comprises the neighboring constellation points surrounding the hard decision point $\hat{x}_k^t$. The conditional probability $\mathbf{P}(x_k^t = s|\hat{x}_t)$ can be computed from the Gaussian approximation, and its computation is shown in [27]. For example, in the case of 4-QAM, $\mathbf{P}(x_k^t = s|\hat{x}_t)$ can be obtained by [30]

$$\mathbf{P}(x_k^t = s|\hat{x}_t) = \begin{cases} 1-2P_e+P_e^2, & \text{if } s=\hat{x}_t \\ P_e-P_e^2, & \text{if } s \text{ is one of two nearest neighbors of } \hat{x}_t \\ P_e^2, & \text{else} \end{cases}$$

where P_e accounts for the probability of error in one dimension for (13) and is given as

$$P_e \triangleq Q\left(\sqrt{\frac{\beta^2\sigma_s^2}{\sigma_w^2}}\right). \tag{17}$$

Here, $Q(x) = \int_x^\infty (1/2\pi)\exp(-u^2/2)du$. An accurate approximation of the Q-function is presented in [31]. Furthermore, plugging (15) into (17) yields a simple expression for 4-QAM as

$$P_e = Q\left(\sqrt{\frac{\beta}{1-\beta}}\right).$$

In this example, we can neglect P_e^2 terms to simplify the computation with little performance loss. After computing (16), this term is added to the covariance matrix $\mathbf{Q}_{\hat{\mathbf{e}}_k^i}$ in (10) for the next step $i+1$.

Next, we briefly describe the LLR computation for the soft bit information [28]. As w in (13) is assumed to be Gaussian, the conditional probability density function (pdf) of $\tilde{x}_k^t$ is given by

$$p(\tilde{x}_k^t|x_k^t = s) = \frac{1}{\pi\sigma_w^2}\exp\left(-\frac{|\tilde{x}_k^t - \beta s|^2}{\sigma_w^2}\right).$$

Let $\mathcal{S}$ be a set of constellation symbols and denote s as an element of the set $\mathcal{S}$. We represent the ith bit of x_k^t by $d_k^{t,i}$ and define two mutually exclusive subsets of $\mathcal{S}$, $\mathcal{S}_0^i$, and $\mathcal{S}_1^i$, which comprise the symbols with "0" and "1" in the ith bit position ($i = 1, 2 \ldots \log_2 M$), respectively. Then, the *a posteriori* LLR of $b_k^{t,i}$ [28] can be defined as

$$\begin{aligned} L(b_k^{t,i}) &\triangleq \log\frac{P[d_k^{t,i} = 0|\tilde{x}_k^t]}{P[d_k^{t,i} = 1|\tilde{x}_k^t]} \\ &= \log\frac{\sum_{s\in\mathcal{S}_0^i} P[x_k^t = s|\tilde{x}_k^t]}{\sum_{s\in\mathcal{S}_1^i} P[x_k^t = s|\tilde{x}_k^t]}. \end{aligned} \tag{18}$$

Assuming that the transmitted symbols have equal probability of occurrence, (18) can be represented as

$$L(b_k^{t,i}) = \log\frac{\sum_{s\in\mathcal{S}_0^i}\exp\left(-\frac{|\tilde{x}_k^t-\beta s|^2}{\sigma_w^2}\right)}{\sum_{s\in\mathcal{S}_1^i}\exp\left(-\frac{|\tilde{x}_k^t-\beta s|^2}{\sigma_w^2}\right)}.$$

Note that a further approximation of the above equation can be made by applying the max-log rule [28].

As addressed in [27], the added computational complexity of the enhanced V-BLAST is small compared with the conventional V-BLAST.

IV. Iterative Detection and Decoding

We now describe an IDD scheme combined with V-BLAST for MIMO-OFDM systems. In the preceding section, we have tried to minimize the effect of error propagation 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 either BCJR algorithm or Viterbi algorithm.

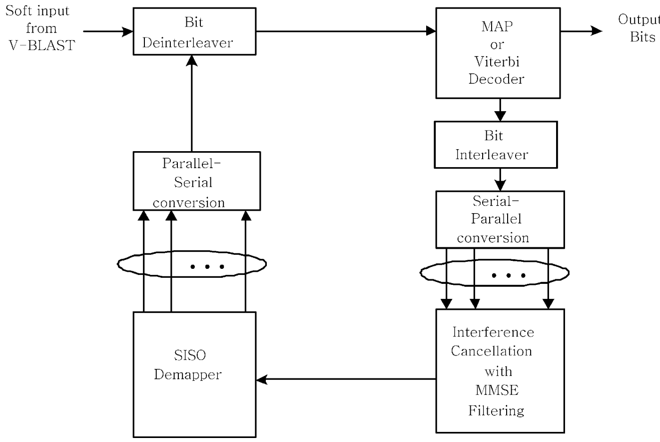

Fig. 4. Proposed IDD block.

Comparing 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 simplifies the computation of the MIMO demapper in the turbo-MIMO, as this block dominates the complexity of the turbo-MIMO concept. Different from other suboptimal detection and decoding 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 order to detect x_k^t, the hard decisions for all the other symbols $x_k^1, \ldots, x_k^{t-1}, x_k^{t+1}, \ldots, x_k^{N_t}$ are used to cancel the interference from $\mathbf{y}_k$ in (1). Let $\hat{x}_k^i$ be the hard decision made in the decoder. Then, defining $\hat{\underline{\mathbf{x}}}_k^t = \left[\hat{x}_k^1, \ldots, \hat{x}_k^{t-1}, 0, \hat{x}_k^{t+1}, \ldots, \hat{x}_k^{N_t}\right]$, the received signal $\mathbf{y}_k$ is modified by canceling the interference $\hat{\underline{\mathbf{x}}}_k^t$ as

$$\begin{aligned}\underline{\mathbf{y}}_k^t &= \mathbf{y}_k - \mathbf{H}_k \hat{\underline{\mathbf{x}}}_k^t \quad \text{for } k = 1, 2, \ldots, N_c \\ &= \mathbf{H}_k \underline{\mathbf{x}}_k^t + \mathbf{n}_k \end{aligned} \tag{19}$$

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

Then, we apply an MMSE filter $\mathbf{w}_t$ 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 variance of the estimation error defined as

$$e = x_k^t - \mathbf{w}_t \underline{\mathbf{y}}_k^t$$

where $\mathbf{w}_t$ is a row vector of length N_r.

Similarly, from (7), the MMSE weight $\mathbf{w}_t$ can be obtained as

$$\mathbf{w}_t = \sigma_s^2 \mathbf{h}_k^{t\dagger} \left(\mathbf{H}_k \mathbf{Q}_{\underline{\mathbf{x}}_k^t} \mathbf{H}_k^\dagger + \sigma_n^2 \mathbf{I}_{N_r}\right)^{-1} \tag{20}$$

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

To simplify the computation, we neglect the off-diagonal terms in the covariance matrix $\mathbf{Q}_{\underline{\mathbf{x}}_k^t}$ since the elements in $\underline{\mathbf{x}}_k^t$ are assumed to be independent in the presence of an interleaver between the MIMO demapper and the decoder. Thus, the computation for the covariance matrix reduces to the equation shown at the bottom of the page.

The computation of $\mathbf{Q}_{\underline{\mathbf{x}}_k^t}$ is different depending 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 $\mathbf{w}_t = \mathbf{h}_k^{t\dagger}/(|\mathbf{h}_k^t|^2 + \sigma_n^2/\sigma_s^2)$. It will be shown in Section V that the significant complexity reduction can be obtained without causing much performance degradation.

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

$$\begin{aligned} E[|e_k^i|^2] &\triangleq E[|x_k^i - \hat{x}_k^i|^2 \mid L(d_k^{i,1}), \ldots, L(d_k^{i,\log_2 M})] \quad \text{for } i \neq t \\ &= \sum_{s \in \mathcal{S}} |s - \hat{x}_k^i|^2 \mathbf{P}(x_k^i = s \mid L(d_k^{i,1}), \ldots, L(d_k^{i,\log_2 M})). \end{aligned} \tag{21}$$

Unlike (16) where the conditional probability of each symbol is approximated by assuming that the output interference-plus-noise is Gaussian, the symbol probability in (21) can be directly obtained using the LLRs of the coded bits at the output of the decoder as shown in Fig. 4. Due to the interleaver, we assume all the bits are independent. Then, it is straightforward to show that

$$\begin{aligned} &\mathbf{P}(x_k^i = s \mid L(d_k^{i,1}), \ldots, L(d_k^{i,\log_2 M})) \\ &\qquad = \prod_{m=1}^{\log_2 M} \mathbf{P}(d_k^{i,m} = d_s^m | L(d_k^{i,m})) \end{aligned}$$

where d_s^m denotes the mth bit of s. From the LLR definition (2), $\mathbf{P}(d_k^{i,m} = d_s^m | L(d_k^{i,m}))$ is given by

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

After computing $E[|e_k^i|^2]$ from (21), $\mathbf{w}_t$ can be obtained in (20) with $\mathbf{Q}_{\underline{\mathbf{x}}_k^t}$. Applying the MMSE filter $\mathbf{w}_t$ to $\underline{\mathbf{y}}_k^t$ in (19) yields the biased estimate of x_k^t as

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

As a result, the LLR value for x_k^t can be computed in a much simpler SISO demapper instead of the MIMO demapper in (3)

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

for the turbo-MIMO. Now, it is straightforward to show that 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 $\sigma_v^2 = \sigma_s^2(\alpha - \alpha^2)$. Here, we simply assume that symbols s are equally probable.

In this SISO demapper, the number of candidates to compute the above LLR values is only M for any antenna configurations, while the MIMO demapper in (3) for the turbo-MIMO requires M^{N_t}. Thus, the proposed IDD reduces the computational complexity substantially. In the following section, the performance of the proposed IDD scheme is illustrated and compared with other existing schemes through simulation results in MIMO-OFDM systems.

V. Simulation Results

In this section, we present the simulation results for the proposed IDD systems illustrated in the previous sections, comparing with the conventional schemes to demonstrate the efficacy of our proposed methods. Gray mapping are assumed throughout simulations. The spectral efficiency of the system is $\eta = R_c \cdot N_t \cdot \log_2 M$ b/s/Hz, where R_c is the rate of the convolutional code used. For V-BLAST, ordered successive interference cancellation (OSIC) [4] is assumed.

A. Simulation Results for the Enhanced V-BLAST

First, we compare the performance of the enhanced V-BLAST with the conventional V-BLAST. Here, we consider flat fadings. Especially, we focus on the block fading channel model where fading coefficients are static over a block of transmitted symbols, and independent over blocks. The block length is set to 200. No iterative decoding is assumed for the evaluation. A binary convolutional code with polynomials (133, 171) in octal notation of rate 1/2 is used for the simulations.

Figs. 5 and 6 show the simulation results for MIMO systems with $N_t = N_r = 4$ in terms of frame error rate (FER) with respect to signal-to-noise ratio (SNR) in decibels (dB). The x axis represents the received SNR per antenna in decibels, and the y axis indicates the FER. For 4-QAM constellation, we can see that the enhanced V-BLAST provides about 6 dB gain at 1% FER over the conventional V-BLAST. Note that those improvements are achieved by considering the decision errors in the equalization process and the soft bit metric generation. As observed in Fig. 6, the gain of the enhanced V-BLAST over the conventional V-BLAST increases to 8 dB at 1% FER for the case of 16-QAM. These results confirm that the decision error compensation is crucial for the coded layered space–time architectures. More simulation results are found in [27].

B. Simulation Results for the Proposed IDD Schemes

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 five-tap power delay profile having exponentially decayed fading characteristics, where each ray is independently Rayleigh fading. 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) and (23,35,27) are used for the code of rate 1/2 and 1/3, respectively, because of the complexity in the BCJR decoder. The code rate and the number of decoding iteration are set to 1/2 and 4, respectively, if not specified otherwise. Here, one iteration refers to the case with no feedback from the decoder.

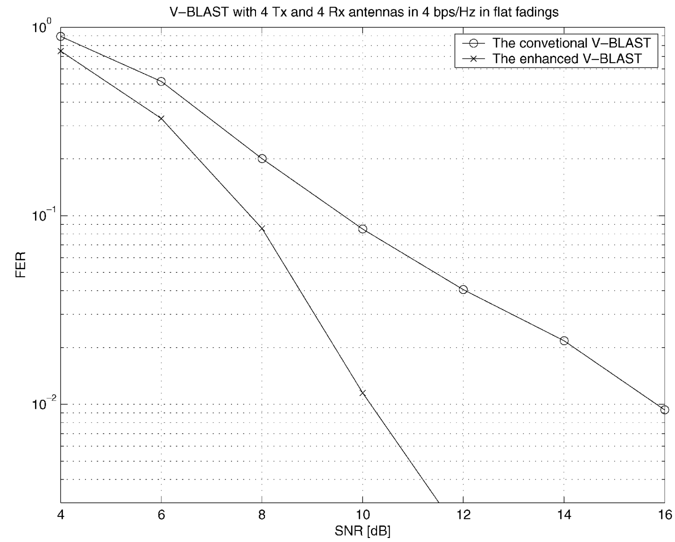

Fig. 5. Frame error probability of coded layer space–time architectures with 4-QAM.

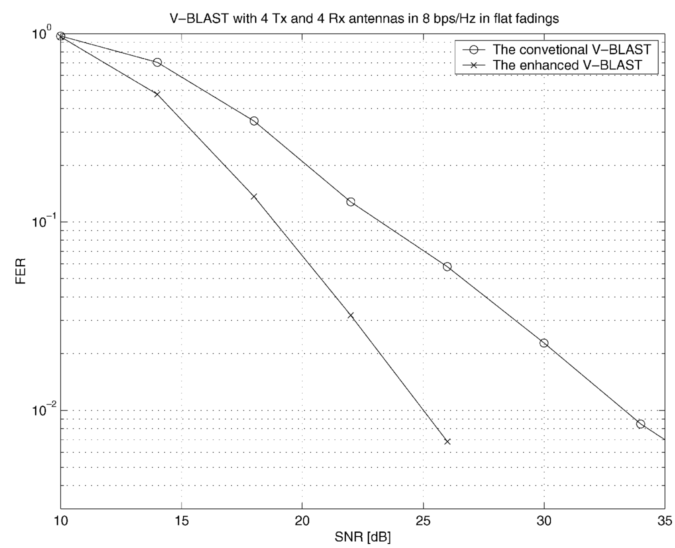

Fig. 6. Frame error probability of coded layer space–time architectures with 16-QAM.

In order to demonstrate the performance of the proposed scheme, 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.

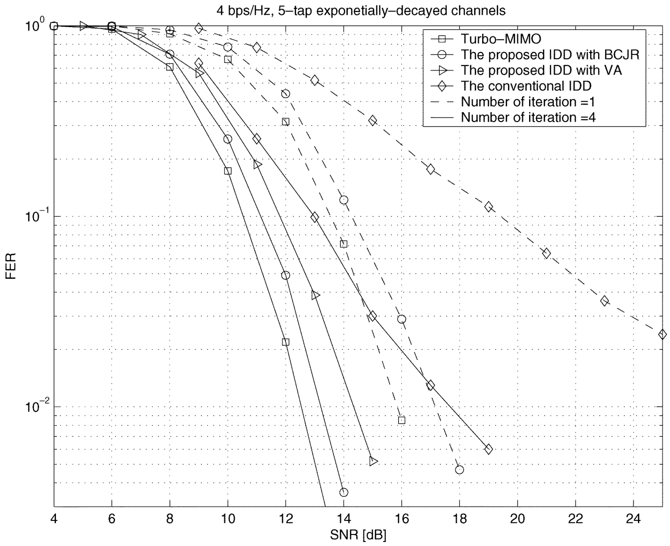

Fig. 7. Frame error probability of different iterative decoding schemes with 16-QAM and $N_t = N_r = 2$.

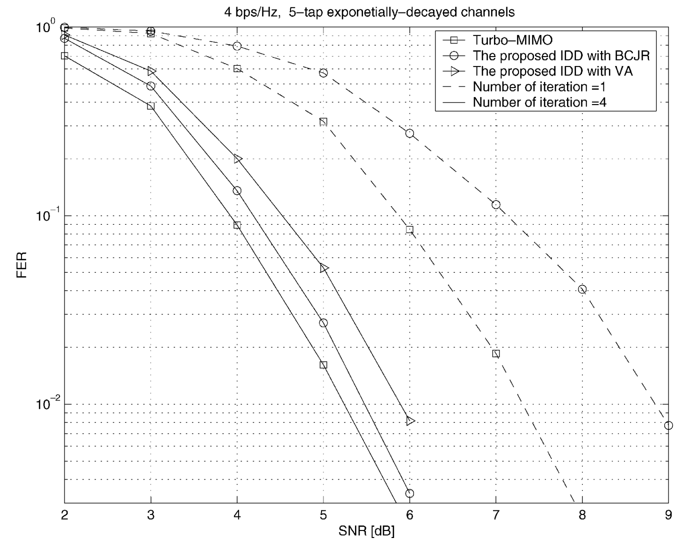

Fig. 9. Frame error probability of different iterative decoding schemes with 4-QAM and $N_t = N_r = 4$.

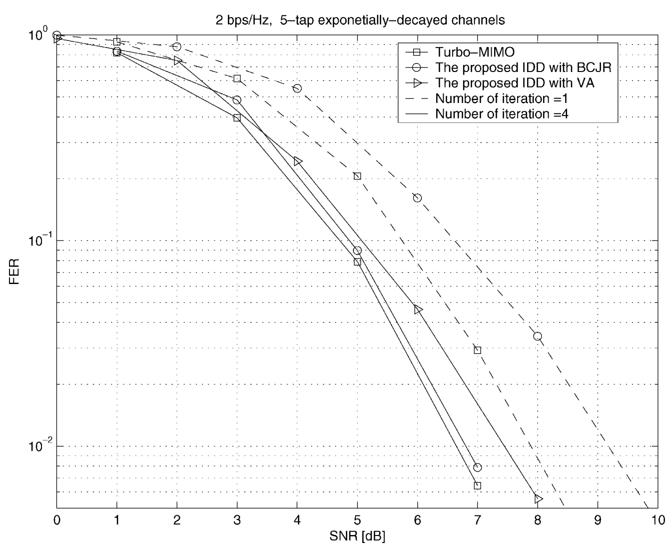

Fig. 8. Frame error probability of different iterative decoding schemes with 4-QAM and $N_t = N_r = 2$.

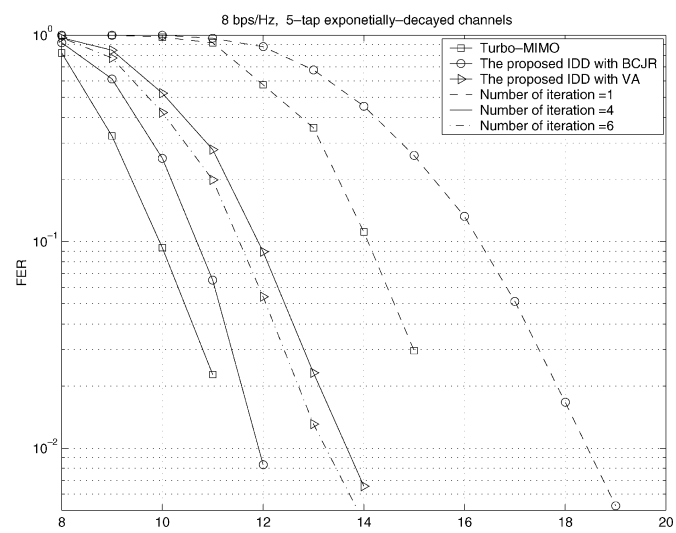

Fig. 10. Frame error probability of different iterative decoding schemes with 16-QAM and $N_t = N_r = 4$.

- *The conventional IDD:* Employing the conventional V-BLAST with Viterbi decoder in the IDD block.

Note that at the first iteration, the performance of the proposed IDD with VA is identical to that of the proposed IDD with BCJR since no feedback from the decoder is available.

Fig. 7 shows the FER of different iterative schemes for 16-QAM 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 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. 7 indicates that the proposed IDD performs within 1–2 dB of the turbo-MIMO while achieving significant complexity reduction. Similarly, Fig. 8 provides simulation results for 4-QAM cases. We can find that compared with the turbo-MIMO, the performance loss of the proposed IDD is only less than 1 dB as the modulation level decreases.

Fig. 9 depicts the performance of MIMO-OFDM systems with the increased number of transmit and receive antennas. In the four transmit antenna case, the performance gap between the turbo-MIMO and the proposed iterative scheme is still less than 1 dB, similarly with $N_t = 2$. It should be noted that the total number of candidates to search in the proposed SISO demapper remains only four for both Figs. 8 and 9, while the number of candidates for the optimum demapper in the turbo-MIMO increases from 16 to 256. In this case, the performance of the proposed IDD with Viterbi decoder is only a few tenth of a dB away from that of the turbo-MIMO, while the complexity is significantly lower.

In Fig. 10, we consider the proposed IDD scheme for a high spectral efficiency. This plot presents the performance comparison in the 4×4 MIMO-OFDM system with 16-QAM which leads to the spectral efficiency of 8 b/s/Hz. This extreme example is impracticable for the turbo-MIMO because the MIMO demapper is simply infeasible. In this case, the optimum MIMO

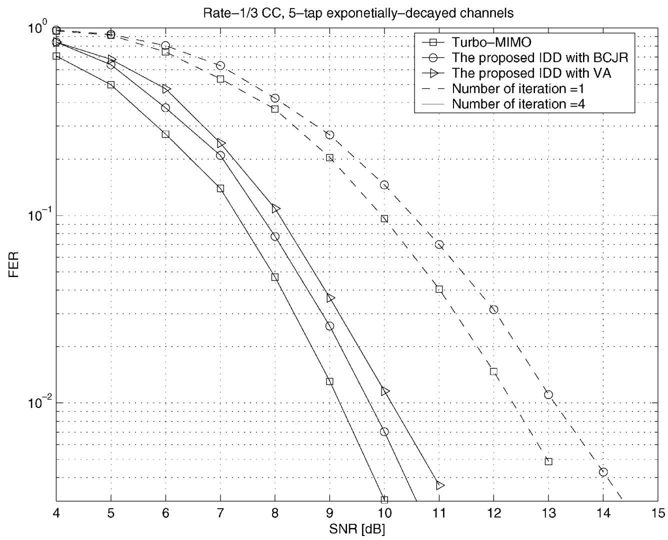

Fig. 11. Frame error probability of different iterative decoding schemes with 16-QAM and $N_t = N_r = 2$.

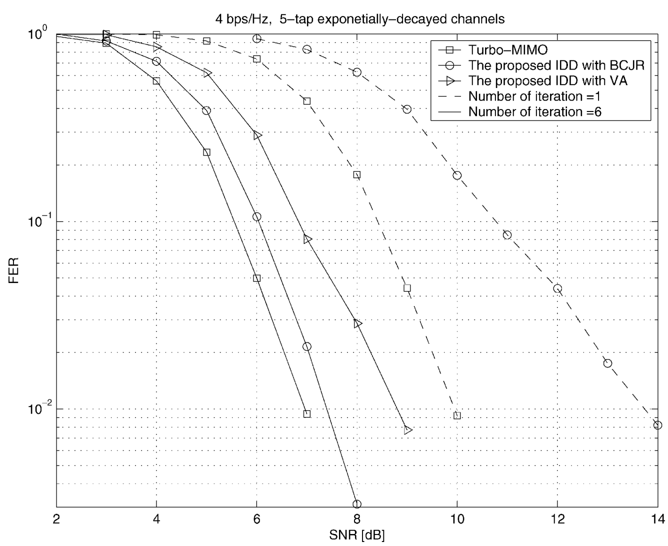

Fig. 12. Frame error probability of different iterative decoding schemes with $N_t = 4, N_r = 3$, and 4-QAM.

demapper needs to search $16^4 = 65\,536$ 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.

Also, the simulation results for lower rate codes are presented in Fig. 11. This simulation assumes the 2×2 system with 16-QAM and rate-1/3 coding. It is important to note that, compared with higher rate codes in Fig. 7, the performance gain of the turbo-MIMO over the proposed IDD scheme is reduced to less than 1 dB for all SNR ranges. Therefore, when lower rate codes are employed, the proposed IDD schemes become much more attractive than the turbo-MIMO.

Finally, we consider a MIMO system with fewer receive antennas than transmit antennas. Fig. 12 shows the simulation results for the 4×3 system with 4 b/s/Hz. In order to effectively implement the V-BLAST algorithm, it is known that the number of receive antennas should be greater than or equal to the number of transmit antennas. Therefore, the performance of the proposed IDD algorithm at the first iteration is poor. However, the performance loss is largely compensated with an aid of the IDD scheme. The figure reveals that a substantial gain in the FER performance is obtained by the proposed IDD scheme.

From the simulation results presented in this section, it is clear that for all simulation configurations, the proposed IDD scheme is quite effective in approaching the performance of the near-optimum turbo-MIMO with a significantly reduced complexity.

VI. Conclusion

In this paper, we have proposed pragmatic schemes for the layered space–time architectures in MIMO-OFDM systems. We have addressed the issue of mitigating the error propagation problem for the vertical coding structure. By introducing a comprehensive signal modeling which includes error propagation, we have derived an improved signal detector based on the MMSE criterion and described the corresponding soft bit LLR value computation method. In addition, we have demonstrated the efficacy of a simple iterative MIMO detector based on the interference cancellation and MMSE filtering. Employing the enhanced V-BLAST as a front-end demodulator, the proposed IDD scheme enables us to achieve further performance gain. Simulation results indicate that the performance of the proposed iterative scheme is just less than 1 dB away from the near-optimum turbo-MIMO for all the simulation configurations with remarkably reduced complexity. The simulation results 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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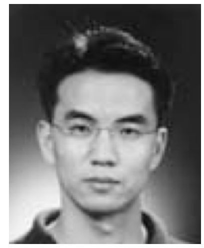

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