신호 공간 다이버시티 기법을 이용한 OFDM 기반의 부호화된 시공간 전송기법

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Coded Layered Space-Time Transmission with Signal Space Diversity in OFDM Systems

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요 약

V-BLAST (Vertical Bell Labs Layered Space-Time) 시스템은 다중 안테나를 사용하는 시스템에서 각각의 레이어를 간섭 무효화 및 제거기법을 통하여 추정하기 때문에 간단한 복잡도에 비해 높은 전송률을 가능하게 한다. 본 논문에서는 신호 공간 다이버시티 기법을 적용시킨 V-BLAST 시스템을 제안하고자 한다. 신호 공간 다이버시티 기법의 큰 장점은 신호좌표를 회전시키고 인페이즈 (inphase) 와 쿼드러처 (quadrature) 성분을 섞어줌으로서 추가적인 대역폭이나 전송 파워의 상승 없이도 다이버시티 이득을 얻을 수 있다. 모의실험을 통해서, 본 논문에서 제안하는 시스템의 성능이 이상적인 시스템 성능에 비해 0.5dB로 근접함을 보여주고자 한다.

Key Words: Multi-Input Multi-output (MIMO) Orthogonal Frequency Division Multiplexing (OFDM), Vertical Bell Labs Layered Space-Time (V-BLAST), Bit-Interleaved Coded Modulation (BICM) and Signal Space Diversity (SSD)

ABSTRACT

In multiple antenna systems, vertical Bell Labs Layered Space-Time (V-BLAST) systems enable very high throughput by nulling and cancelling at each layer detection. In this paper, we propose a V-BLAST system which combines with signal space diversity technique. The benefit of the signal space diversity is that we can obtain an additional gain without extra bandwidth and power expansion by applying inphase/quadrature interleaving and the constellation rotation. Through simulation results, it is shown that the performance of the proposed system is less than 0.5dB away from the ideal upper bound.

Ⅰ. Introduction

It is widely acknowledged that reliable wireless transmission of video, data and voice at high rates will be an important part of the next generation communication systems. One way to allow high data rate transmission over rich-scattering channels is to use multiple transmit and receive antennas. As shown in [1], the link-level capacity in multi-input multi-output (MIMO) systems increases linearly with the number of antennas.

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Moreover, another benefit of MIMO systems is to achieve spatial diversity gain. Theoretically, the maximum diversity order in M transmit and N receive antenna systems is equal to MN[2], and the diversity order achievable in MIMO systems depends on the detection method employed on the receiver side. The maximum likelihood (ML) detector yields the diversity order of N[3], whereas zero-forcing linear equalizer (ZF-LE) has diversity order N-M+1[4]. Although the ML detector maintains its optimality, its complexity grows exponentially with $|\chi|^M$, where $|\chi|$is the constellation size. To reduce the complexity, the ML detector can be implemented using other methods such as the sphere decoding algorithm[5] with little performance loss[6]. The sphere decoding, however, still imposes a complexity issue in coded systems. Thus, many suboptimal reduced complexity detection algorithms have been studied and developed.

Vertical Bell Labs Layered Space-Time (V-BLAST) architecture[7] is a very attractive solution in multiple antenna systems. The V-BLAST uses a combination of linear and/or nonlinear detection techniques: first nulling out the interference from undetected signals, then cancelling out the interference using detected signals. Although the V-BLAST system does not fully exploit the diversity, it exhibits the best trade-off between performance and complexity.

As increasing demand for higher data rate leads to wideband communications, wireless channels become frequency selective. To mitigate the effect of frequency selective channels, several techniques have been studied. Orthogonal frequency division multiplexing (OFDM) has been employed as one of effective techniques to handle inter-symbol interference (ISI). The OFDM technique divides the frequency selective fading channel into multiple flat fading subchannels. Such a property of OFDM enables to deploy the receiver designed for flat fadings into frequency selective environments. As a result, the OFDM system combined with the V-BLAST is expected to be a promising solution for next generation wireless communication systems.

However, the performance of the V-BLAST system suffers from the error propagation, which is inherent in nulling and cancelling process, and this is a main source of performance degradation.

Thus, various methods to improve the performance of the V-BLAST have been developed. One method to deal with the error propagation issue of the V-BLAST system is based on the iterative detection and decoding (IDD) process between the detector and decoder[8][9]. To improve the accuracy of the estimation through the iterative processing, we can reduce the effect of the error propagation. However, this method requires high complexity as the modulation level, the number of antennas and/or the number of iteration increase. In contrast, other methods introduced in [10] and [11], which are based on cancelling using decoded decision (CDD) where the interference is cancelled using hard decisions from the decoder of the previously decoded layers. This method exhibits a lower complexity compared to the IDD method.

One way to obtain an extra diversity gain is to employ the signal space diversity (SSD) method[12]. The SSD achieves additional diversity gains through the inphase/quadrature shuffling and the constellation rotation. A main advantage of the SSD is that it is possible to increase the diversity order without extra bandwidth and power expansion. In this paper, we extend the SSD method to V-BLAST systems combined with CDD in OFDM. We show that the overall performance of the V-BLAST system improves with the SSD. The simulations demonstrate that the performance of the proposed system is close to the ideal upper bound from the simulation.

The rest of this paper is organized as follows. In Section II, we present an overview for the conventional V-BLAST system. In Section III, we introduce the proposed V-BLAST system with SSD. In Section III, we analyze the proposed V-BLAST system with the SSD in terms of diversity and the detection ordering. In Section IV, simulation results are presented. Finally, the conclusion is given in Section V.

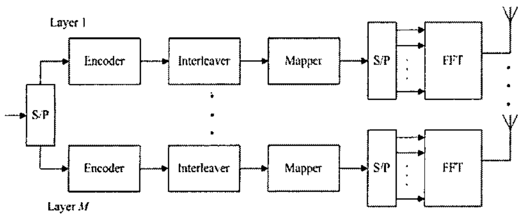

Fig. 1. Transmitter for coded layered space-time OFDM system

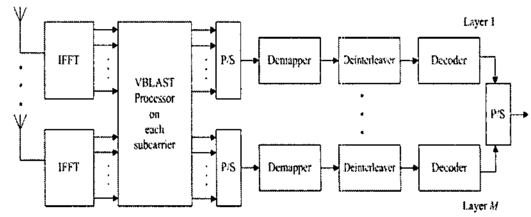

Fig. 2. Receiver for coded layered space-time OFDM system

Ⅱ. System Overview

In this section, we describe coded layered space-time OFDM systems implemented in the V-BLAST architecture with M transmit and N receive antennas.

2.1 System Model

Figures 1 and 2 depict the transmitter and the receiver structure for the coded layered space-time OFDM system, respectively, assuming N_c subcarriers. At the transmitter, each data stream, called layer, is independently encoded, bit interleaved and symbol mapped. The individual symbol stream is then transmitted through each antenna. At the receiver, the V-BLAST detection algorithm is deployed on every subcarriers.

Let $\mathbf{x}_k$ and $\mathbf{y}_k$ denote the transmitted and received signal vectors in the kth subcarrier, respectively. Then we can write the signal model as

$$\mathbf{y}_k = \mathbf{H}_k \mathbf{x}_k + \mathbf{w}_k \qquad (1)$$

where

$$\mathbf{H}_k = \begin{bmatrix} h_{1,1,k} & \cdots & h_{1,M,k} \\ \vdots & \ddots & \vdots \\ h_{1,1,k} & \cdots & h_{1,1,k} \end{bmatrix}, \qquad \mathbf{w}_k = \begin{bmatrix} w_{1,k} \\ \vdots \\ w_{N,k} \end{bmatrix}. \qquad (2)$$

Here, $h_{j,i,k}$ represents the channel coefficient between the ith transmit antenna and the jth receive antenna at the kth subcarrier. The time domain channel impulse response for the ith transmitter and the jth receiver link is given as

$$\mathbf{a}_{j,i} = \left[a_{j,i}(0) \quad a_{j,i}(1) \quad \cdots \quad a_{j,i}(L-1) \right]$$

where L is the number of channel taps, and $a_{i,j}(l)$ represents the lth tap of the channel. Then, the channel coefficient $h_{j,i,k}$ is computed by the N_c-point FFT as

$$h_{j,i,k} = \sum_{l=0}^{N_c-1} a_{j,i}(l) W_{N_c}^{lk}, \qquad k = 1, \cdots, N_c$$

where $W_{N_c}^{lk} = \exp\left(-\frac{j2\pi l(k-1)}{N_c} \right)$.

In (2), the noise components in $\mathbf{w}_k$ are assumed to be independent and identically-distributed (i.i.d) complex Gaussian with $E\{\mathbf{w}_k \mathbf{w}_k^H\} = \sigma_w^2 \mathbf{I}_N$, where $E\{\cdot\}$ and $(\cdot)^H$ indicate the expectation and the conjugate transpose, respectively, and $\mathbf{I}_N$ denotes an identity matrix with size N.

2.2 V-BLAST Detection

The processing of the V-BLAST detection is described as follows[7].

1) Set m=1.

2) For the mth layer, compute the minimum mean squared error (MMSE) equalization matrix $\mathbf{G}_k$:

$$\mathbf{G}_k = \left(\mathbf{H}_k^H \mathbf{H}_k + \alpha \mathbf{I}_{M+1-m} \right)^{-1} \mathbf{H}_k^H$$

where $\alpha = \sigma_w^2 / \sigma_x^2$, and σ_x^2 is the average energy with $E\{\mathbf{x}_k^H \mathbf{x}_k\} = \sigma_x^2 \mathbf{I}_M$.

3) Estimate the detected symbol :

$$\hat{x}_{m,k} = \left(\mathbf{G}_k \right)_m \mathbf{y}_k$$

where $(\mathbf{G}_k)_m$ represents the mth row vector of $\mathbf{G}_k$, and $\hat{x}_{m,k}$ denotes the mth element of $\mathbf{x}_k$.

4) Cancel the interference using the estimated symbol $\hat{x}_{m,k}$:

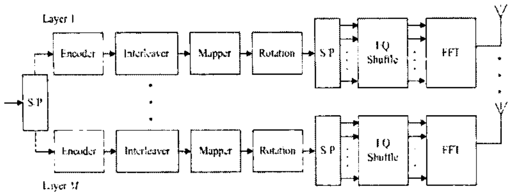

Fig. 3. Transmitter for coded layered space-time OFDM system with SSD

$$\mathbf{y}_k \leftarrow \mathbf{y}_k - \mathbf{h}_{m,k}\hat{x}_{m,k}$$
$$\mathbf{H}_k \leftarrow \left[\mathbf{h}_{m+1,k} \quad \cdots \quad \mathbf{h}_{M,k}\right]$$
$$m \leftarrow m+1$$

where $\mathbf{h}_{j,k}$ represents the jth column vector of $\mathbf{H}_k$.

5) Go back to step 2) until $m=M$.

If we employ the CDD, some detection procedures are changed. At step 3), the whole soft bit values for the mth layer are calculated and are passed to the decoder. The decoder output is then transformed to the symbol stream. At step 4), the interference cancellation takes place using the symbol stream obtained from the decoder. See [10] and [11] for detailed description of the CDD scheme.

Ⅲ. System Model with SSD

As introduced in [12], a key point of the SSD technique is to make the inphase and quadrature components of the received symbol experience independent fadings. To make this happen, component interleavers are employed to break the correlation between the inphase and quadrature components. The complexity increase of the SSD scheme is very small. Except for the component interleaving, the transmitter and the receiver structure remain the same as the conventional system. Also, the size of the component interleaving equals the number of subcarrier, which is small compared to the bit-level interleaving.

Figure 3 shows the SSD scheme applied to the coded layered space-time OFDM system considered in this paper. We define the $M \times N_c$ complex matrix as the transmit symbol matrix as

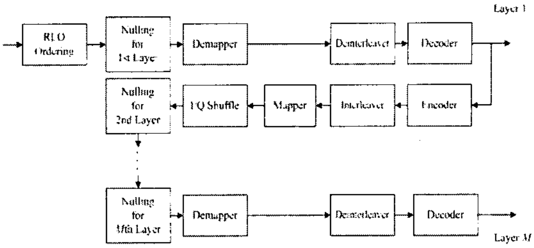

Fig. 4. Receiver for coded layered space-time OFDM system with SSD

$$\mathbf{S} = \begin{bmatrix} s_{1,1} & \cdots & s_{1,k} & \cdots & s_{1,N_c} \\ \vdots & \ddots & \ddots & \ddots & \vdots \\ s_{M,1} & \cdots & s_{M,k} & \cdots & s_{M,N_c} \end{bmatrix}$$

where $s_{i,j}$ indicates the transmitted symbol through the ith transmit antenna at the jth subcarrier. For conventional systems without SSD, we have $\mathbf{s}_k=\mathbf{x}_k$ for $k=1,\cdots,N_c$, where $\mathbf{s}_k$ denotes the kth column vector of $\mathbf{S}$.

The transmitted symbol matrix can be broken into two real matrices $\mathbf{S}^I$ and $\mathbf{S}^Q$ where $\mathbf{S}^I$ and $\mathbf{S}^Q$ consist of the inphase and quadrature components of $\mathbf{S}$, respectively, as

$$\mathbf{S}^I = \begin{bmatrix} s_{1,1}^I & \cdots & s_{1,k}^I & \cdots & s_{1,N_c}^I \\ \vdots & \ddots & \ddots & \ddots & \vdots \\ s_{M,1}^I & \cdots & s_{M,k}^I & \cdots & s_{M,N_c}^I \end{bmatrix},$$
$$\mathbf{S}^Q = \begin{bmatrix} s_{1,1}^Q & \cdots & s_{1,k}^Q & \cdots & s_{1,N_c}^Q \\ \vdots & \ddots & \ddots & \ddots & \vdots \\ s_{M,1}^Q & \cdots & s_{M,k}^Q & \cdots & s_{M,N_c}^Q \end{bmatrix}.$$

Then, we can rewrite the transmitted symbol matrix as

$$\mathbf{S} = \mathbf{S}^I + j\mathbf{S}^Q. \qquad (5)$$

In order to achieve full diversity, the constellation rotation is employed in the mapper block shown in the SSD structure in Figure 3[12]. By employing the component interleavers, the columns of $\mathbf{S}^I$ and $\mathbf{S}^Q$ are exchanged. Notice that the interleaving patterns of the inphase and quadrature components need to be different from each other, whereas the same interleaving patterns are allowed among layers. After the component interleaving, $\mathbf{S}^I$ and $\mathbf{S}^Q$ are transformed to $\mathbf{X}^I$ and $\mathbf{X}^Q$, respectively. As indicated in Figure 3, the symbol matrix passed to IFFT is expressed as

$$\mathbf{X} = \mathbf{X}^I + j\mathbf{X}^Q.$$

In the same manner as in (4) and (5), we define the $N \times N_c$ received signal matrix as

$$\mathbf{Y} = \begin{bmatrix} \mathbf{y}_1 & \cdots & \mathbf{y}_k & \cdots & \mathbf{y}_{N_c} \end{bmatrix} = \begin{bmatrix} y_{1,1} & \cdots & s_{1,k} & \cdots & s_{1,N_c} \\ \vdots & \ddots & \ddots & \ddots & \vdots \\ y_{M,1} & \cdots & y_{M,k} & \cdots & y_{M,N_c} \end{bmatrix}.$$

Now we will describe the receiver structure which combines the SSD and the V-BLAST with CDD illustrated in Figure 4. Before the detection process starts, we should determine the detection ordering. As mentioned in the previous section, each layer needs to be decoded before the cancellation in the V-BLAST with CDD. For this reason, the detection ordering should be decided first. In this paper, we employ the representative layer ordering (RLO) introduced in [11]. The RLO is determined based on maximizing the capacity of each layer. Denoting C_m as the Shannon capacity for the mth layer, C_m can be obtained by

$$C_m = -\frac{1}{N_c}\sum_{k=1}^{N_c} \log_2 \left\{ \left[\left((\rho / M) \mathbf{H}_k^H \mathbf{H}_k + \mathbf{I}_M \right)^{-1} \right]_{mm} \right\}$$

where ρ is the average received power to noise ratio per receive antenna, and $[\cdot]_{ij}$ means the (i,j) element of a matrix.

Comparing the capacity of each layer, the layer with the largest C_m should be detected first. Let $O(m)$ be the layer index with the mth detection order. Then, the new received signal model can be written as

$$\mathbf{y}_k = \bar{\mathbf{H}}_k \bar{\mathbf{x}}_k + \mathbf{w}_k$$

where

$$\bar{\mathbf{H}}_k = \begin{bmatrix} \mathbf{h}_{O(1),k} & \mathbf{h}_{O(2),k} & \cdots & \mathbf{h}_{O(M),k} \end{bmatrix}$$
$$\bar{\mathbf{x}}_k = \begin{bmatrix} x_{O(1),k} & x_{O(2),k} & \cdots & x_{O(M),k} \end{bmatrix}^T.$$

After the ordering, the detection process begins according to the layer order. To simplify presentations, we assume $O(m)=m$ for all m, which means $\bar{\mathbf{H}}_k = \mathbf{H}_k$ and $\bar{\mathbf{x}}_k = \mathbf{x}_k$. At the mth stage, we calculate the equalization matrix $\mathbf{G}_k$ for the whole subcarriers using $\mathbf{H}_k$ and (3). Then, the $\mathbf{G}_k$ is applied to the received signal for each subcarriers. Denoting r_k as the output of the equalization filter at the kth subchannel, we can write r_k as

$$r_k = (\mathbf{G}_k)_m \mathbf{y}_k = \underbrace{(\mathbf{G}_k)_m \mathbf{h}_m}_{\beta_k} x_m + \underbrace{\sum_{j=m+1}^{N} (\mathbf{G}_k)_m \mathbf{h}_j x_j + (\mathbf{G}_k)_m \mathbf{w}_k}_{\mu_k}.$$

As the MMSE equalization is a biased processing[13], we obtain the unbiased estimation z_k by scaling r_k as

$$z_k = r_k / \beta_k = x_k + v_k \quad \text{for} \quad k = 1, \cdots, N_c. \quad (6)$$

Then, these unbiased estimations are component-wise deinterleaved and are entered into the demapper.

Next, the Log-Likelihood Ratio (LLR) values are calculated in the demapper block. As the residual ISI of the MMSE filter μ_k is well approximated by a complex Gaussian distribution[14], we can assume that the probability of v_k is also complex Gaussian. Moreover, due to the component deinterleaving, the inphase and the quadrature parts of v_k are statistically independent with different variances. Let u be the complex Gaussian random variable with independent inphase and quadrature components. Then, we can write the probability density function (pdf) of u as

$$f(u) = f(u^I) f(u^Q) = K \exp\left(-\frac{|u^I|^2}{2\sigma_I^2}\right) \exp\left(-\frac{|u^I|^2}{2\sigma_Q^2}\right)$$

where σ_I^2 and σ_Q^2 represent the variances of the inphase and quadrature parts, respectively, and $K = 1/2\pi\sqrt{\sigma_I^2\sigma_Q^2}$.

Under this assumption, the conditional probability $p(z_k, z_{k'} | c_{m,p})$ in (6) can be written as

$$p(z_k, z_{k'} | c_{m,p}) = K \exp\left(-\frac{|z_k^I - c_{m,p}^I|^2}{2\sigma_I^2}\right) \exp\left(-\frac{|z_k^Q - c_{m,p}^Q|^2}{2\sigma_I^2}\right)$$

where k and k' indicate the indices of subcarriers corresponding to index p of the inphase and quadrature components after deinterleaving, respectively, and $\sigma_I^2 = E\{|v_k|^2\}/2$ and $\sigma_Q^2 = E\{|v_{k'}|^2\}/2$.

Denoting $b_{m,p}^i$ as the ith bit of $s_{m,p}$, the LLR value for $b_{m,p}^i$ is computed using $p(z_k, z_{k'} | c_{m,p})$ as

$$LLR(b_{m,p}^i) = \log \frac{\sum_{c_{m,p} \in \chi_0^i} p(z_k, z_{k'} | c_{m,p})}{\sum_{c_{m,p} \in \chi_1^i} p(z_k, z_{k'} | c_{m,p})},$$
$$i = 1, 2, \cdots, \log_2 |\chi| \quad \text{and} \quad p = 1, \cdots, N_c$$

where χ_d^i is a set of all symbols whose ith bit is d (=0,1) in the rotated constellation χ.

After the LLR values are passed to the decoder, the decoder output is re-encoded and transformed to the symbol stream $\tilde{\mathbf{s}}(m)$ denoted by

$$\tilde{\mathbf{s}}(m) = \begin{bmatrix} \tilde{s}_{m,1} & \cdots & \tilde{s}_{m,N_c} \end{bmatrix}$$

Applying the inphase and quadrature interleaving, we obtain the symbol vector $\tilde{\mathbf{x}}(m)$ to be applied in the interference cancellation as

$$\tilde{\mathbf{x}}(m) = \tilde{\mathbf{x}}^I(m) + j\tilde{\mathbf{x}}^Q(m) = \begin{bmatrix} \tilde{x}_{m,1} & \cdots & \tilde{x}_{m,k} & \cdots & \tilde{x}_{m,N_c} \end{bmatrix}$$

where $\tilde{\mathbf{x}}^I(m)$ and $\tilde{\mathbf{x}}^Q(m)$ are column-wise shuffled vectors of $\tilde{\mathbf{s}}^I(m)$ and $\tilde{\mathbf{s}}^Q(m)$, respectively.

Finally, $\tilde{\mathbf{x}}(m)$ is used to cancel the interference and the received signal vector and the channel matrix are updated as

$$\mathbf{y}_k \leftarrow \mathbf{y}_k - \mathbf{h}_{m,k}\tilde{x}_{m,k}$$
$$\mathbf{H}_k \leftarrow \begin{bmatrix} \mathbf{h}_{m+1,k} & \cdots & \mathbf{h}_{M,k} \end{bmatrix}.$$

For the next layer detection, we set $m \leftarrow m+1$, and repeat the same procedure until $m=M$.

Ⅳ. Simulation Results

In this section, we will show the numerical results for the proposed SSD scheme in V-BLAST systems. 64-point IFFT/FFT are used as OFDM modulator/demodulator, and the guard period is set to 16 symbols. We consider an exponentially decaying 5-tap channel, which is referred to as 'channel5' with 4QAM constellation. We assume that the channel state information (CSI) is known to the receiver. Also, the channel response is time-invariant during one OFDM symbol interval, assuming to be statistically independent between OFDM symbol intervals. The optimum rotation angle for 4QAM in uncoded SSD system is derived in [15]. Although various angles have been reported based on different analysis, we choose 0.322 radian for the rotation angle.

A convolutional encoder with rate 3/4 and constraint length 7 is employed, and the same encoder is used for each layer. We consider two systems with M=N=2 and M=N=4 for the simulation and their spectral efficiencies are 3bps/Hz and 6bps/Hz, respectively.

To compare the proposed scheme and other methods, we employ the following legends:

• Conventional : The conventional V-BLAST

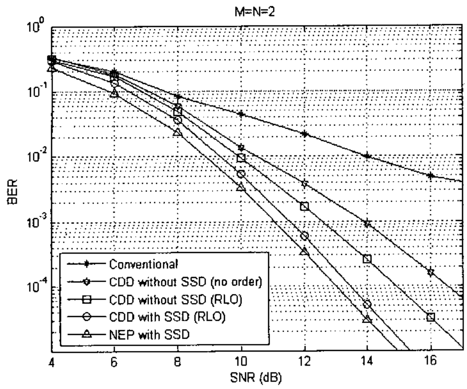

Fig. 5. BER for 2x2 system with different detection and ordering schemes

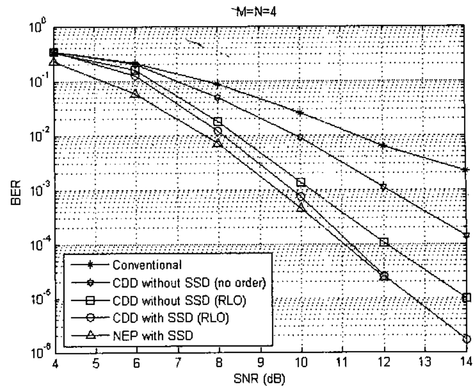

Fig. 6. BER for 4x4 system with different detection and ordering schemes

detection.

- NEP : No error propagation case where interferences are perfectly cancelled out, which serves as an ideal upper bound.
- SSD : The signal space diversity scheme proposed in this paper.

In Figure 5, we show the performance results of different coded layered space-time structures for $M=N=2$. Note that the V-BLAST processing using the CDD exhibits a significant improvement over the conventional V-BLAST. Also we note the RLO is quite effective for the V-BLAST with CDD. When the RLO is employed, The proposed CDD with SSD shows an additional 2dB gain at BER 10^{-4} over the CDD without SSD. Moreover, the performance of the proposed scheme is very close to the ideal case. This confirms that our scheme can achieve an additional diversity gain by utilizing the SSD.

Figure 6 depicts the performance results for $M=N=4$. Similar to the $M=N=2$ case, our proposed scheme shows an additional gain over the system without SSD. Although the performance gain for $M=N=4$ is smaller than that for $M=N=2$, we confirm that the performance of the proposed system becomes much closer to the ideal upper bound. In our simulation, we decide the order of detection layer before CDD process as shown in Figure 4. Actually, the layer ordering should be done for every step before detection. In this case, however, the performance is not improved significantly compared to that for our case considered in this paper[11].

When lower rate encoders are employed, the performance gain of the proposed scheme may be reduced, since the achievable frequency diversity from the SSD becomes smaller. With the similar reason, the performance gain of the proposed SSD decreases with larger number of antennas. The increased spatial diversity sacrifices the achievable gain in the SSD.

V. Conclusion

In this paper, we investigate the coded layered space-time system in frequency selective fading channels. In order to fully exploit the frequency diversity, the signal space diversity technique has been employed for the proposed systems. To eliminate the error propagation effect, cancellation using decoded decision is performed by utilizing the decoder output. Simulation results demonstrate that about 2dB gain at BER 10^{-4} is achieved over the conventional system.

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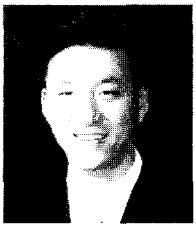

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