An Improved Hybrid STBC Scheme with Precoding and Decision Feedback Detection

Chang-Kyung Sung and Inkyu Lee
Dept. of Communication Engineering
Korea University
Seoul, Korea
cksung@korea.ac.kr, inkyu@korea.ac.kr

***Abstract*— In practical wireless communication systems, the number of the receive antenna is more restricted than that of the transmit antenna. Spatial division multiplexing (SDM) technique normally imposes a restriction that the number of receive antenna should be more than or equal to that of transmit antennas whereas the space time coding (STC) technique can be adopted without such a limitation. The hybrid STBC scheme is a mixture of the above two systems. We propose two approaches to improve the hybrid STBC scheme:a precoding method and a decision feedback detection. Applying a linear precoding matrix with full rank property can improve the performance. Also, the decision feedback detection scheme cancels out the interference using already-detected components. Simulation results show that the proposed methods outperform the conventional hybrid STBC detection algorithm for frequency selective channels.**

I. INTRODUCTION

For next generation wireless communication systems, we need to combat several impairments caused by wideband channel environments. Among them, the interference caused by frequency selective channels is the most severe one.

Combination of Orthogonal Frequency Division Multiplexing (OFDM) and multiple antennas is one of the most attractive solutions for improving the transmission quality in the given bandwidth. By employing the OFDM, wideband transmission is possible over frequency selective fading channels without applying equalizers. OFDM has been proposed for a wide range of radio systems such as wireless local area network (WLAN) system defined by the IEEE 802.11a standard.

In addition, multi-input/multi-output (MIMO) antenna systems can achieve additional diversity which is a classical method to improve transmission over fading channels. It has been shown [1] that MIMO antenna systems provide multiple independent channel, and thus, the channel capacity increases linearly with the number of antennas.

There are generally two approaches to implement the MIMO system. One is to increase the transmission diversity through space time coding (STC) [2] and the other is to raise the total throughput by employing spatial division multiplexing (SDM). STC mitigates fading through spatial diversity whereas SDM increases link capacity. With STC, both the space and time diversity can be obtained by multiple transmit and receive antennas combined with matching modulation and coding. They can be adopted in various antenna configurations.

In practical wireless communication systems, the number of receive antenna is more restricted than that of transmit antenna. SDM technique normally imposes a restriction that the number of receive antenna should be at least equal to that of transmit antenna whereas STC method can be adopted without such a limitation. The hybrid STBC scheme proposed in [3] is a mixture of the above two systems. It transmits two Alamouti blocks [4] concurrently with four transmit antennas and detects the transmitted signal with a linear equalization detection scheme. User information bits are encoded with a convolutional encoder, then interleaved by the bit interleaver and mapped into appropriate signal points using M-PSK or M-QAM mapping. Then, signal streams are split into four substreams and are transmitted using two Alamouti STBC blocks.

In this paper, we propose two approaches to improve the hybrid STBC scheme: a precoding method and a decision feedback detection scheme which is also known as Bell-lab LAyered Space-Time architecture (BLAST) algorithm [5]. The precoding method increases the spatial diversity by mixing two symbols and sending them to different antennas.

Since the hybrid STBC system transmits the space-time coded block over multiple symbol periods, the corresponding channel frequency response matrix can have full rank as long as there are more than two receive antennas. Thus, it is possible to apply the BLAST detection algorithm even if there are smaller number of receive antennas than transmit antenna. The decision feedback detection utilizes decisions from the previous stage and cancels out interferences successively. Since the conventional hybrid STBC detector employs the nulling operation only, it is expected that nulling with cancellation can show a better performance.

In time-varying channels, a conventional detector should employ a solution different from the quasi-static channel case. However, the decision feedback detection scheme can be applied without any modification regardless of channel condition and outperforms the conventional method.

This paper is organized as follows: In section II, we describe the conventional hybrid STBC. In section III, we propose the precoding method and the decision feedback detection method.

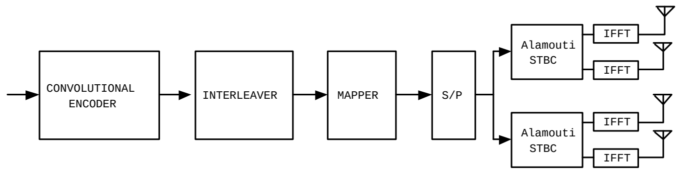

Fig. 1. 4x2 hybrid STBC system model

Finally, we present simulation results in section IV.

II. System Model

Let us denote N_t and N_r as the number of transmit and receive antennas respectively. Denoting T_{STBC} as the block period needed for transmitting embedded STBC blocks, the hybrid STBC can be applied for an $N_t \times N_r$ antenna configuration where $N_r \geq N_t/T_{STBC}$. Before transmitted through multiple antennas, information sequences are mapped to signal constellation by bit interleaved coded modulation (BICM) [6]. The BICM introduces diversity by convolutional encoding and the bit level interleaving. It can achieve the frequency diversity in the OFDM system.

After the mapping operations, the entire signals are splitted into N_t substreams and each substream is transmitted through N_t antennas. With M-QAM constellations, bit interleaved and coded data bits $[d_k^1 \; d_k^2 \ldots d_k^{log_2 M}]$ are mapped into a symbol s_k where k indicates the subcarrier index. The scheme with $T_{STBC} = 2$ and $N_t = 4$ is proposed in [3]. As mentioned above, N_r should be grater than or equal to 2 for the hybrid STBC detector. Figure 1 shows the 4 by 2 Hybrid STBC system architecture. In this system, two Alamouti STBC blocks [4] are transmitted in parallel and a linear equalizer detects transmitted symbols.

Let us denote the modulated symbol $\underline{s}^k$ at the k th subcarrier as $[s_1^k \; s_2^k \; s_3^k \; s_4^k]^T$ with $(.)^T$ representing the transpose. Then the received signal $\underline{y}^k$ at the k th subcarrier can be written as

$$\begin{aligned}
\underline{y}^k &= \begin{bmatrix} \underline{y}_n^k \\ \underline{y}_{n+1}^{k\,*} \end{bmatrix} = H^k \underline{s}^k + \underline{n}^k \\
&= \begin{bmatrix} h_{1,1}^k & h_{2,1}^k & h_{3,1}^k & h_{4,1}^k \\ h_{1,2}^k & h_{2,2}^k & h_{3,2}^k & h_{4,2}^k \\ {h_{2,1}^k}^* & -{h_{1,1}^k}^* & {h_{4,1}^k}^* & -{h_{3,1}^k}^* \\ {h_{2,2}^k}^* & -{h_{1,2}^k}^* & {h_{4,2}^k}^* & -{h_{3,2}^k}^* \end{bmatrix} \underline{s}^k + \underline{n}^k \\
&= \begin{bmatrix} \underline{h}_1^k & \underline{h}_2^k & \underline{h}_3^k & \underline{h}_4^k \\ \underline{h}_2^{k*} & -\underline{h}_1^{k*} & \underline{h}_4^{k*} & -\underline{h}_3^{k*} \end{bmatrix} \underline{s}^k + \underline{n}^k \\
&= \begin{bmatrix} H_1^k & H_2^k \end{bmatrix} \underline{s}^k + \underline{n}^k
\end{aligned} \tag{1}$$

where $\underline{y}_n^k$ represents the received signal vector of the k th subcarrier at time n, $h_{i,j}^k$ indicates the channel frequency response from the transmit antenna i to the receive antenna j at the k th subcarrier and $\underline{n}^k$ is defined as $[n_n^k \; {n_{n+1}^k}^*]^T$. Here, it is assumed that channel frequency responses are quasi-static for simplicity. Each noise component has variance σ_n^2.

For the OFDM modulation, the received signal for each subcarrier can be independently modeled as (1). However, the channel frequency responses are correlated in frequency. Assuming that there exist L taps in a channel profile, the channel frequency response from the transmit antenna i to the received antenna j can be modeled as

$$\hat{h}_{i,j}(t,\tau) = \sum_{n=1}^{L} \bar{h}_{i,j}(n,t)\delta(\tau - \tau_n) \tag{2}$$

where the channel coefficients $\bar{h}_{i,j}(n,t)$ are independent complex Gaussian distribution with zero mean, $\delta(\tau)$ denotes the Dirac delta function and τ_n represents the propagation delay for the n th channel tap. Assuming a quasi-static block fading channel model, we will omit the time index t for simplicity. With the OFDM modulation, the equivalent channel frequency response for each subcarrier is represented as the inverse Fourier transformed version of (2). Then, the channel frequency response for the k th subcarrier can be written as

$$h_{i,j}^k = \sum_{n=1}^{L} \bar{h}_{i,j}(n) e^{-2\pi k \tau_n / FT} \quad .$$

where F denotes the number of subcarriers in the OFDM system.

In the hybrid STBC scheme, intersymbol interference exists between two STBC blocks, and detection of $\underline{s}^k$ is performed by a linear equalizer. So the signal detection is carried out by multiplying the equalization matrix W^k,

$$\hat{\underline{s}}^k = W^k \underline{y}^k \quad .$$

The filter matrix W^k is obtained with respect to the specific equalization method. Since the detection operation is independent with respect to subcarriers, we will omit the subcarrier index k for simplicity. For the minimum mean square error (MMSE) equalization, W can be computed as [3]

$$W_{MMSE} = \frac{1}{d_1 d_2 - (\delta - \gamma)} \begin{bmatrix} d_2 I & -H_1^H H_2 \\ -H_2^H H_1 & d_1 I \end{bmatrix} \begin{bmatrix} H_1^H \\ H_2^H \end{bmatrix} \tag{3}$$

where $(.)^H$ denotes the Hermitian transpose and

$$\begin{aligned}
d_1 &= \sum_{i=1}^{2} (\|h_{1,i}\|^2 + \|h_{2,i}\|^2) + \sigma_n^2 \\
d_2 &= \sum_{i=1}^{2} (\|h_{3,i}\|^2 + \|h_{4,i}\|^2) + \sigma_n^2 \\
\delta &= |\underline{h}_1^H \underline{h}_3 + \underline{h}_2^T \underline{h}_4^*|^2 \\
\gamma &= |\underline{h}_1^H \underline{h}_4 - \underline{h}_2^T \underline{h}_3^*|^2 \quad .
\end{aligned}$$

The zero forcing solution for the hybrid STBC can be easily obtained by discarding σ_n^2 terms in d_1 and d_2. For time varying channels, the channel matrices H_1 and H_2 are not orthogonal any more. In this case, the equalization matrix should be obtained using the companion matrix property [3] which results in somewhat different solutions from the above quasi-static case.

III. Improved Hybrid STBC Scheme

In this section, we propose two methods to improve the performance of the hybrid STBC. A linear precoding matrix is proposed to improve the performance by increasing the rank of the distance matrix A_e, which will be shown later. Also the decision feedback detection scheme is employed to improve the performance by canceling out the interference from already-detected components of the received vector.

A. Precoding

In order to improve the conventional hybrid STBC performance, we adopt a precoding matrix Θ for a 4 by 2 system as

$$\begin{aligned}\Theta &= \frac{1}{\sqrt{2}}\begin{bmatrix}\theta_1 & \theta_2 & 0 & 0\\ \theta_3 & \theta_4 & 0 & 0\\ 0 & 0 & \theta_5 & \theta_6\\ 0 & 0 & \theta_7 & \theta_8\end{bmatrix}\\ &= \frac{1}{\sqrt{2}}\begin{bmatrix}\mathcal{C}_1 & \mathit{0}_2\\ \mathit{0}_2 & \mathcal{C}_2\end{bmatrix}\end{aligned}$$

where $\mathcal{C}_i$ and $\mathit{0}_2$ denote the 2x2 precoding submatrix for each Alamouti STBC block and the zero submatrix respectively. Here, the term $\frac{1}{\sqrt{2}}$ is an energy normalization factor. To achieve additional diversity, Θ should have full rank. We choose $\mathcal{C}_1$ and $\mathcal{C}_2$ as

$$\mathcal{C}_1 = \mathcal{C}_2 = \begin{bmatrix}1 & e^{j\theta}\\ 1 & e^{j(\pi+\theta)}\end{bmatrix} \quad .$$

Of course, one may choose different precoding submatrices with $\mathcal{C}_1 \neq \mathcal{C}_2$. However, symbols precoded by each submatrix are independently space-time coded, thus, we adopt the same submatrix for simplicity.

Then, the received signal $\underline{y}$ is given as

$$\underline{y} = H\Theta\underline{s} + \underline{n} \quad .$$

Let $c = \Theta\underline{s}$ and $\hat{c} = \Theta\underline{\hat{s}}$ be the received signal vectors for correct symbols and erroneous symbols, respectively. Also, define the distance matrix A_e as

$$A_e = (R_h)^{1/2^H}(I_{N_r} \otimes (\hat{c}-c)(\hat{c}-c)^H)(R_h)^{1/2}$$

where R_h is the correlation matrix for the channel matrix, $\otimes$ denotes the Kronecker product. Let us define G_d^i as $rank(A_e)$ and G_c^i as

$$G_c^i = [\prod_{l=0}^{G_d^i-1} \lambda_{e,l}]^{-G_d^i} \quad .$$

Then the diversity gain is defined as $G_d = \min_{c\neq\hat{c}} G_d^i$ and the coding gain with nonzero eigenvalues of A_e, $\lambda_{e,l}$ is obtained as [2],[7]

$$G_c = \min_{c\neq\hat{c}} G_c^i \quad .$$

When the precoding is applied, one can increase the rank of A_e by spreading each symbol to two antennas connecting to the same Alamouti STBC block. The Chernoff bound of the pairwise error probability is given by

$$P(c \to e) \leq [G_c^i \frac{1}{4N_0}]^{-G_d^i} \quad .$$

Therefore the pairwise error probability can be decreased by increasing the rank of A_e.

Now, we consider the soft value computation. The soft value of the d_i^m is obtained by the joint detection of two coupled precoded symbols. Denote S as a set of the signal alphabet and the subset $S_d^{i,m}$, $d = +1$ or $d = -1$, of S as a set of all symbols with a +1 or -1 value of the bit d_i^m, respectively. Then the log-likelihood ratio (LLR) for the bit d_i^m is obtained as [8]

$$LLR(d_{i=2l}^m) = \log \frac{\sum_{\tilde{s}_{even}^{+1}} \sum_{k=0}^{1} \exp(-\frac{1}{\sigma^2}\|s_{2l+k} - \tilde{s}\mathcal{C}^k)\|^2}{\sum_{\tilde{s}_{even}^{-1}} \sum_{k=0}^{1} \exp(-\frac{1}{\sigma^2}\|s_{2l+k} - \tilde{s}\mathcal{C}^k)\|^2}$$

$$LLR(d_{i=2l+1}^m) = \log \frac{\sum_{\tilde{s}_{odd}^{+1}} \sum_{k=0}^{1} \exp(-\frac{1}{\sigma^2}\|s_{2l+k} - \tilde{s}\mathcal{C}^k)\|^2}{\sum_{\tilde{s}_{odd}^{-1}} \sum_{k=0}^{1} \exp(-\frac{1}{\sigma^2}\|s_{2l+k} - \tilde{s}\mathcal{C}^k)\|^2}$$

where l is given as $l = 0, 1$ and $\mathcal{C}^k$ denotes the k th row of the 2 by 2 precoding submatrix. Here, $\tilde{s}_{even}^d$ is defined as $[s_1\ s_2]$ where $s_1 \in S_d^{l,m}, s_2 \in S$, and also $\tilde{s}_{odd}^d$ is similarly defined as $[s_1\ s_2]$ where $s_1 \in S, s_2 \in S_d^{l,m}$.

B. Decision Feedback Detection

When the decision feedback detection is employed, the successive cancellation reduces the intersymbol interference by using already detected symbols. Since the channel matrix in (1) has full rank, it is possible to detect the transmitted signal using the BLAST algorithm. We apply the decision feedback detection with the minimum mean square error (MMSE) criterion. The filter matrix for the decision feedback detection at stage i can be computed by

$$W_{DFB}^i = (H_i^* H_i + \sigma_n^2 I)^{-1} H_i^* \quad .$$

The initial channel matrix H_0 is set to H in (1) and H_i at stage i is obtained by discarding columns corresponding to the previously detected layers [5].

In the BLAST detection algorithm, the order in which the received symbols are detected is important to achieve the desired performance. The covariance matrix of the estimation error $s - \hat{s}$ is used to determine the detection order. The layer to be detected at the current stage i can be determined by the position of the minimum diagonal entry of the covariance matrix. For simplicity of illustration, we assume that $\underline{s}$ is ordered according to the optimal ordering. The covariance matrix is given as

$$\begin{aligned}E[(s-\hat{s})(s-\hat{s})^*] &= E[(s - W_{DFB}^i \cdot y_i)(s - W_{DFB}^i \cdot y_i)^*]\\ &= (H_i^* H_i + \sigma_n^2 I)^{-1}\end{aligned}$$

where y_i is the corresponding received signal column vector at stage i which is obtained by

$$y_i = y_{i-1} - (H)_{i-1} Q(\hat{s}_{i-1}) \tag{4}$$

where $(H)_i$ and $Q(.)$ denote the $i-1$ th column of H and the quantization function, respectively. Here, $\hat{s}_{i-1}$ represents the decision made in the previous stage $i-1$.

Then, the i th component of $\underline{s}$ at stage i is detected by multiplying the i th row of the weight vector W^i_{DFB}, $\underline{w}_i = [w_{i,1} \ldots w_{i,4}]$ to the received signal vector $\underline{y}_i$. Thus, the estimated symbol can be expressed as

$$\begin{aligned}
\hat{s}_i &= \underline{w}_i \cdot y_i \\
&= [w_{i,1} \ldots w_{i,4}] \begin{bmatrix} y^i_1 \\ y^i_2 \\ \vdots \\ y^i_4 \end{bmatrix} = \sum_{j=1}^{4} w_{i,j} y^i_j \\
&= [w_{i,1} \ldots w_{i,4}] \begin{bmatrix} \sum_{l=1}^4 \tilde{h}_{1,l} s_l \\ \sum_{l=1}^4 \tilde{h}_{2,l} s_l \\ \vdots \\ \sum_{l=1}^4 \tilde{h}_{4,l} s_l \end{bmatrix} + \sum_{l=1}^{4} w_{i,l} n_l \\
&= \sum_{j=1}^{4} w_{i,j} \sum_{l=1}^{4} \tilde{h}_{j,l} s_l + \sum_{l=1}^{4} w_{i,l} n_l \\
&= \sum_{j=1}^{4} w_{i,j} \tilde{h}_{j,i} s_i + \sum_{j=1}^{4} w_{i,j} \sum_{\substack{l=1 \\ l \neq i}}^{4} \tilde{h}_{j,l} s_l + \sum_{l=1}^{4} w_{i,l} n_l
\end{aligned} \tag{5}$$

where $\tilde{h}_{i,j}$ is the (i,j) element in the channel matrix H and n^i_l denotes the l th noise component with variance $\sigma^2_{n,i}$. The second and third term in (5) can still be considered as additive noise terms with variance σ^2_i.

Let us consider the computation of LLR values. The LLR for bit d^m_i is computed as [8], [9]

$$LLR(d^m_i) = \log \frac{\Sigma_{s \in S^{i,m}_{+1}} \exp(-\frac{1}{\sigma^2} \|s_i - s\|^2)}{\Sigma_{s \in S^{i,m}_{-1}} \exp(-\frac{1}{\sigma^2} \|s_i - s\|^2)}$$

where

$$\sigma^2_i = \| \sum_{k=1}^{4} \sum_{i<l \leq N_t} w_{ik} h_{kl} \|^2 E_s + \sum_{l=1}^{4} \|w_{il}\|^2 \sigma^2_{n,i} \quad .$$

where E_s denotes the average symbol energy. After detecting $\hat{s}_i$, the signal vector y_i is updated for the next detection as in (4).

Noting that there exists residual variance $E[\|s_i - Q(\hat{s}_i)\|^2]$ due to incorrect decisions. The total interference-plus-noise power considering error propagation can be computed as

$$\sigma^2_i + (H)_i E[\|s_i - Q(\hat{s}_i)\|^2 | \underline{y}] \ .$$

The variance of the residual interference can be computed as [10]

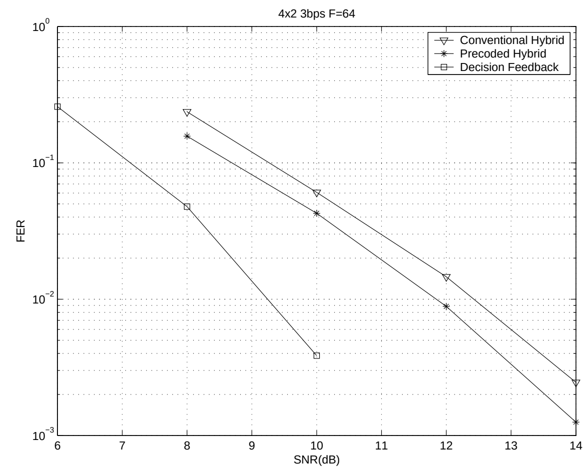

Fig. 2. 3bps/Hz 4x2 antenna configuration in the quasi-static channel

$$E[\|s_i - Q(\hat{s}_i)\|^2 | \underline{y}] = \sum_{s \in \mathcal{N}_{x_k}} \|s - x_k\|^2 P(s|x_k)$$

where $\mathcal{N}_{x_k}$ comprises the neighboring constellation points surrounding the hard decision points $x_k = Q(\hat{s}_i)$

IV. Simulation Results

In this section, we present the Monte Carlo simulation results to compare the proposed system with the conventional hybrid STBC. We adopt the MIMO-OFDM with 64 subcarriers ($F = 64$). The guard interval length is set to 16, and we assume the 5 tap exponentially decaying channel. This channel accounts for approximately the root mean square (RMS) spread of 100ns and each ray is assumed to be independently Rayleigh fading. Throughout the simulations, one frame is assumed to consist of one OFDM symbol for simplicity. Thus the size of the random interleaver is given by $F \cdot N_t \cdot \log_2 M$. The spectral efficiency of the system is defined as [8] $R_T = R_c \cdot N_t \cdot \log_2 M / T_{STBC}$ bps/Hz where R_c is the convolutional code rate.

Figure 2 shows the simulation results for 3bps/Hz case. To achieve 3bps/Hz, $R = \frac{3}{4}$ punctured convolutional code [11] with QPSK is used. We assume that the channel state information is perfectly known to the receiver. In Figure 2, we plot the frame error rate (FER) of the proposed system and the conventional method at quasi-static channels as a function of signal-to-noise ratio (SNR). The plot shows that about 0.5dB gain is achieved by employing the precoding method, and more than 2dB gain is obtained using the proposed decision feedback scheme at FER=10^{-2}.

Figure 3 exhibits the simulation results for 2bps/Hz. For 2 bps/Hz, the convolutional encoder with $R = \frac{1}{2}$ is used with the same constellation.

The plot also shows that almost 3dB gain is achieved at 1% FER. This indicates that the performance gain of the proposed scheme increases at the lower modulation level.

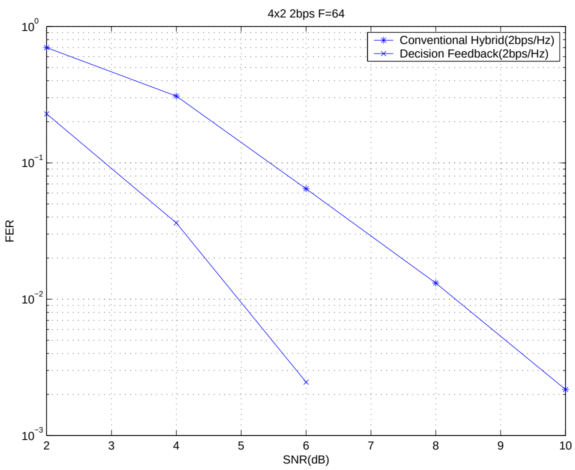

Fig. 3. 2bps/Hz 4x2 antenna configuration in the quasi-static channel

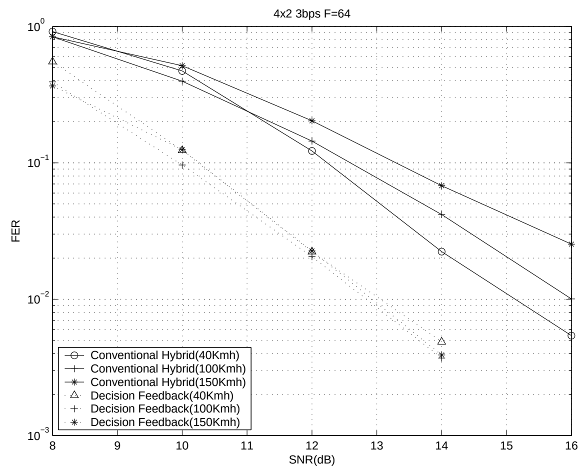

Fig. 4. 3bps/Hz 4x2 antenna configuration in the time varying fading channel

Now we consider time varying channels. At fast fading channels, the channel frequency responses during the STBC block period T_{STBC} no longer satisfies the quasi-static condition. In that case, the solution for the conventional system given by (3) must be modified. So the filter matrix should be obtained by applying the companion matrix property [3]. However, the decision feedback detection proposed in this paper maintains the same solution as in the quasi-static case.

We compare two systems in slowly fading channels by using the Jake's model [12]. For our simulation, we choose the carrier frequency f_c as 5.8 GHz and the sampling rate f_s as 12.5 KHz. We change the mobile speed from 40km/h to 150km/h. In the conventional STBC, it requires quasi-static channel assumption for proper combining at the receiver. However, the decision feedback detection does not impose such a quasi-static channel assumption, so it is expected that our proposed system can show a better performance than the conventional method in time varying channels. We adopt the companion matrix solution proposed in [3] for the conventional hybrid STBC. Figure 4 shows that the proposed system is quite effective even at high mobility with little performance degradation, whereas the performance of the conventional system is considerably degraded at higher speed.

V. Conclusion

We have introduced two methods for improving the hybrid STBC. The precoding scheme improves the hybrid STBC by introducing additional diversity. The decision feedback detection scheme achieves the performance gain by canceling out the interference using the previously detected symbols. This scheme can be applied to any configuration of STBC embedded SDM systems. In addition, the decision feedback detection can adopt more sophisticated BLAST detection algorithms. The proposed decision feedback scheme does not require the quasi-static channel condition. Therefore, this exhibits a better performance in time varying channels. Also, the performance of the proposed scheme is quite insensitive to the mobile speed. The simulation results shows that about 3dB gain is achieved by applying the proposed scheme.

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