Decision Feedback Detection with Error Compensation for Hybrid Space-Time Block Codes

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***Abstract*— Spatial division multiplexing (SDM) techniques increase the total throughput by transmitting independent information streams through multiple transmit antennas whereas space time coding (STC) techniques utilize diversity gain. Hybrid space-time block code (STBC) schemes proposed in [1] combine the above two techniques to maximize the link performance. We propose a decision feedback detection method to improve the performance of the hybrid STBC scheme for orthogonal frequency division multiplexing (OFDM). In this scheme, we take the error propagation effect into account to enhance the detection performance. Simulation results show that the proposed method outperforms the conventional hybrid STBC detection algorithm by more than 3dB at 1% frame error rate for frequency selective fading channels.**

***Index Terms*— Space-time block code (STBC), orthogonal frequency division multiplexing (OFDM), decision feedback detection.**

I. Introduction

FOR next generation wireless communication systems, we need to combat several impairments in wideband channel environments. Among them, interference caused by frequency selective fading channels is the most severe one. Orthogonal Frequency Division Multiplexing (OFDM) is one of attractive solutions for improving the transmission quality in the given bandwidth. By employing OFDM modulation, wideband transmission is possible over frequency selective fading channels without applying equalizers [2]. In addition, multi-input/multi-output (MIMO) antenna systems can also improve the system performance by achieving transmission diversity through space time coding (STC) or by increasing total throughput with spatial division multiplexing (SDM).

The hybrid space-time block code (STBC) scheme proposed in [1] is a mixture of the above two techniques. It transmits two Alamouti blocks [3] modulated by the OFDM concurrently after bit interleaved coded modulation (BICM) with four transmit antennas. The BICM combined with the OFDM provides an additional coding gain so that it is now widely adopted by many wireless communication systems such as IEEE 802.11a wireless local area networks (WLANs). For the hybrid STBC receiver, a linear equalization detection scheme is employed to detect the transmitted signal [1].

In this letter, we propose a decision feedback detection scheme which takes the decision errors into account to improve the hybrid STBC detection performance. Bell lab layered space time (BLAST) [4] is a well known detection scheme which cancels out the interference using the previously detected symbols. As the hybrid STBC system transmits the space-time coded block over two symbol periods, the corresponding channel frequency response matrix has 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 is a smaller number of receive antennas than transmit antennas. However, one of major problems in BLAST is error propagation inherent in the successive cancellation operations. In this letter, we introduce a new filter optimization scheme which compensates for the error propagation effect. In the simulation section, it will be shown that the proposed scheme outperforms the conventional hybrid STBC.

II. Hybrid STBC

Consider an OFDM system with multiple antennas for frequency selective fading channels. Let us denote N_t and N_r as the number of transmit and receive antennas, respectively. Before transmitted through multiple antennas, information sequences are assumed to be mapped to signal constellation symbols by the BICM [5]. The BICM introduces diversity by combining convolutional encoding and bit-level interleaving. After the mapping operation, the signal is split into N_t substreams and each substream is transmitted through N_t antennas. With an M-QAM constellation, the coded and bit interleaved 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 hybrid STBC scheme proposed in [1] transmits embedded STBC blocks over two symbol periods with four transmit antennas ($N_t = 4$). As mentioned above, N_r should be greater than or equal to two for the hybrid STBC. Fig. 1 shows the transmitter structure of the hybrid STBC system combined with BICM-OFDM.

Let us denote the modulated symbol $\underline{s}^k$ at the kth subcarrier as $[s_1^k\ s_2^k\ s_3^k\ s_4^k]^T$ with $(.)^T$ representing the transpose operation. Then the STBC code matrix $\underline{c}^k$ at the kth subcarrier can be represented as

$$\underline{c}^k = \begin{bmatrix} s_1^k & s_2^k & s_3^k & s_4^k \\ s_2^{k*} & -s_1^{k*} & s_4^{k*} & -s_3^{k*} \end{bmatrix} .$$

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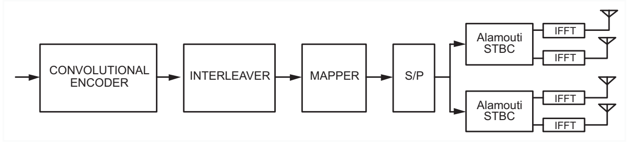

Fig. 1. Transmitter structure of hybrid STBC system.

Here, the first and second row account for the vector transmitted at time t and $t+1$, respectively. After applying the conjugate operation to the received signal at time $t+1$ and combining with the received signal at time t, the received signal vector at the kth subcarrier is modified to

$$\begin{aligned}\underline{y}^k &= \begin{bmatrix} \underline{y}_t^k \\ \underline{y}_{t+1}^{k\,*} \end{bmatrix} = H^k \underline{s}^k + \underline{n}^k \\ &= \begin{bmatrix} h_{1,1}^k & h_{2,1}^k & h_{3,1}^k & h_{4,1}^k \\ h_{1,2}^k & h_{2,2}^k & h_{3,2}^k & h_{4,2}^k \\ h_{2,1}^{k\,*} & -h_{1,1}^{k\,*} & h_{4,1}^{k\,*} & -h_{3,1}^{k\,*} \\ h_{2,2}^{k\,*} & -h_{1,2}^{k\,*} & h_{4,2}^{k\,*} & -h_{3,2}^{k\,*} \end{bmatrix} \underline{s}^k + \underline{n}^k \\ &= \begin{bmatrix} H_1^k & H_2^k \end{bmatrix} \underline{s}^k + \underline{n}^k \end{aligned}$$

where $\underline{y}_t^k$ represents the received signal at the kth subcarrier at time t, $h_{i,j}^k$ indicates the channel frequency response from transmit antenna i to receive antenna j at the kth subcarrier, and $\underline{n}^k$ is defined as $[n_t^{k,1}\ n_t^{k,2}\ n_{t+1}^{k,1\,*}\ n_{t+1}^{k,2\,*}]^T$ with $n_t^{k,j}$ representing the noise component at the kth subcarrier at the receive antenna j at time t. Here, it is assumed that the channel frequency responses are quasi-static. Each noise component has variance σ_n^2. Assume that there exist L taps in a channel profile. Then denoting the sampling period as T, the channel frequency response at the kth subcarrier can be written as $h_{i,j}^k = \sum_{l=1}^{L} \bar{h}_{i,j}(l) e^{-2\pi k \tau_l / FT}$ where F denotes the number of subcarriers in the OFDM system and $\bar{h}_{i,j}(l)$ represents the lth tap of the channel response from the ith transmit antenna to the jth receive antenna with the propagation delay τ_l.

In the conventional hybrid STBC scheme, intersymbol interference (ISI) exists between two STBC blocks, and the detection of $\underline{s}^k$ is performed by a linear equalizer [1]. Thus, the signal detection is carried out by applying the equalization matrix W^k to the received signal $\underline{y}^k$. Since the detection operation is independent of subcarriers, we will omit the subcarrier index k from now on. For the minimum mean square error (MMSE) equalization, the equalizer matrix W is given as [1]

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

where $(.)^H$ denotes the Hermitian transpose and $d_m = \sum_{i=1}^{2} (\|h_{2m-1,i}\|^2 + \|h_{2m,i}\|^2) + \sigma_n^2$, $\delta = |\underline{h}_1^H \underline{h}_3 + \underline{h}_2^T \underline{h}_4^*|^2$, $\gamma = |\underline{h}_1^H \underline{h}_4 - \underline{h}_2^T \underline{h}_3^*|^2$.

III. Decision Feedback Detection with Error Compensation

In this section, we propose a new decision feedback detection scheme for hybrid STBC which takes the error propagation into account. When the decision feedback detection is employed, diversity order can be improved by successive cancellation which reduces the ISI using already detected symbols. However, the interference cancellation performed by incorrectly detected symbols substantially degrades the performance through error propagation. Also, the erroneous cancellation yields incorrect soft values for the channel decoder. Thus, we consider a new formula which takes the error propagation into account to improve the performance.

In the BLAST detection algorithm, the order in which the received symbols are detected is important to achieve the desired performance. The layer to be detected at the current stage can be determined by the position of the minimum diagonal entry of the estimation error covariance matrix [4]. For simplicity of illustration, we assume that $\underline{s}$ is ordered according to the optimal ordering.

Now, we illustrate a new decision feedback detection scheme which takes decision errors into account. We define $\underline{\mathbf{s}}_i = [s_i \dots s_4]^T$, $H_i = [\underline{\tilde{h}}_i \dots \underline{\tilde{h}}_4]$, $\underline{\hat{\mathbf{s}}}_{i-1} = [\hat{s}_1 \dots \hat{s}_{i-1}]^T$ and $\hat{H}_{i-1} = [\underline{\tilde{h}}_1 \dots \underline{\tilde{h}}_{i-1}]$ where $\hat{s}_i$ denotes the detected symbol at the ith stage and $\underline{\tilde{h}}_i$ represents the ith column of the channel matrix H. The detection process is performed by treating undetected signals $s_i \dots s_4$ as interferers and applying linear nulling methods such as MMSE. In such cases, the error propagation is inevitable and this results in performance degradation. In the presence of decision errors, the received symbol vector at the ith stage can be expressed as

$$\underline{y}_i = \underline{y} - \hat{H}_{i-1} \underline{\hat{\mathbf{s}}}_{i-1} = H_i \underline{\mathbf{s}}_i + \hat{H}_{i-1} \underline{\hat{e}}_{i-1} + \underline{n}$$

where $\underline{\hat{e}}_{i-1} = [e_1 \dots e_{i-1}]^T$ is defined with $e_i = s_i - \hat{s}_i$. Note that in the conventional V-BLAST [4], decision errors $\underline{\hat{e}}_{i-1}$ are neglected.

Using the MMSE criterion, the equalizer matrix W_{DFB}^i is formulated to minimize the mean square value of the error $\underline{\mathbf{s}}_i - W_{DFB}^i \cdot \underline{y}_i$. From the orthogonality principle, W_{DFB}^i should satisfy $E[(\underline{\mathbf{s}}_i - W_{DFB}^i \cdot \underline{y}_i) \cdot \underline{y}_i^\dagger] = 0$ where $(.)^\dagger$ denotes the complex-conjugate transpose. Finally, using this result, the equalization matrix W_{DFB}^i at the ith stage can be obtained as

$$W_{DFB}^i = H_i^* (H_i H_i^* + \frac{1}{E_s} \hat{H}_{i-1} Q_{\hat{e}_{i-1}} \hat{H}_{i-1}^* + \sigma_n^2 I)^{-1} \quad (1)$$

where E_s denotes the average signal energy and the decision error covariance matrix $Q_{\underline{\hat{e}}_{i-1}}$ is defined as

$$\begin{aligned} Q_{\underline{\hat{e}}_{i-1}} &= E[\underline{\hat{e}}_{i-1} \underline{\hat{e}}_{i-1}^*] \\ &= \begin{bmatrix} E[|e_1|^2 | \hat{s}_1] & \cdots & E[e_1|\hat{s}_1] E[e_{i-1}^* | \hat{s}_{i-1}] \\ \vdots & \ddots & \vdots \\ E[e_{i-1}|\hat{s}_{i-1}] E[e_1^*|\hat{s}_1] & \cdots & E[|e_{i-1}|^2 | \hat{s}_{i-1}] \end{bmatrix}. \end{aligned}$$

Here, the errors are assumed to be uncorrelated. $E[e_l|\hat{s}_l]$ and $E[|e_l|^2|\hat{s}_l]$ can be obtained as $\sum_{s \in \mathcal{N}_s} (s - \hat{s}) P(s|\hat{s})$ and $\sum_{s \in \mathcal{N}_s} |s - \hat{s}|^2 P(s|\hat{s})$, respectively, where $\mathcal{N}_s$ comprises the neighboring constellation points surrounding the hard decision points $\hat{s}$ and $P(s|\hat{s})$ is the pairwise error probability that a maximum likelihood (ML) detector chooses the erroneous point $\hat{s}$ over the correct point s.

Once W_{DFB}^i is obtained, the ith component of $\underline{s}$ is detected by multiplying the ith row of the weight vector W_{DFB}^i, $\underline{w}_i = [w_{i,1} \dots w_{i,4}]$, with the received signal vector $\underline{y}_i$. Then, the estimated symbol $\hat{s}_i = \underline{w}_i \cdot \underline{y}_i$ can be expressed as

$$\hat{s}_i = \underline{w}_i \underline{\tilde{h}}_i s_i + \sum_{l=1}^{i-1} \underline{w}_i \underline{\tilde{h}}_l e_l + \sum_{l=i+1}^{4} \underline{w}_i \underline{\tilde{h}}_l s_l + \underline{w}_i \underline{n} \ . \qquad (2)$$

The last three terms in (2) can be considered as interference plus noise with total variance σ_i^2. When detecting the signal at the ith stage, there exists interference due to incorrect decisions. Thus, the total interference-plus-noise power considering error propagation is given as [6]

$$\begin{aligned} \sigma_i^2 &= \sum_{l=1}^{i-1} |\underline{w}_i \underline{\tilde{h}}_l|^2 E[|e_l|^2|\hat{s}_l] \qquad (3) \\ &+ \sum_{l=i+1}^{4} |\underline{w}_i \underline{\tilde{h}}_l|^2 E_s + \|\underline{w}_i\|^2 \sigma_n^2 \ . \end{aligned}$$

As an example, for QPSK modulation with the minimum Euclidean distance d between two distinct constellation points, $E[e_l|\hat{s}_l]$ and $E[|e_l|^2|\hat{s}_l]$ can be computed as

$$\begin{aligned} E[e_l|\hat{s}_l] &= (1+j) \cdot d(P_e(1-P_e) + P_e^2) \\ E[|e_l|^2|\hat{s}_l] &= 2d^2 \cdot P_e(1-P_e) + (\sqrt{2}d)^2 \cdot P_e^2 \end{aligned}$$

where $P_e = Q\left(\sqrt{E_s|\underline{w}_{i-1}\underline{\tilde{h}}_{i-1}|^2/\sigma_{i-1}^2}\right)$ and $Q(x)$ is defined as $\frac{1}{\sqrt{2\pi}}\int_x^\infty e^{\frac{-z^2}{2}} dz$. Here P_e^2 term can be neglected.

Now we compute the log likelihood ratio (LLR) for soft channel decoding considering the error propagation. When computing the LLR values, we need to include the interference term due to decision errors as the first term in (3). Let the set $S_d^{i,m}$, d = +1 or -1 be a set of all symbol vectors with a +1 or -1 value of the mth bit of s_i, d_i^m, respectively. Since the interference terms in the MMSE detection can be approximated to be Gaussian, the conditional probability $P(d_i^m = d|\hat{s}_i)$ is obtained as

$$P(d_i^m = d|\hat{s}_i) = \sum_{s \in S_d^{i,m}} \frac{1}{\sqrt{2\pi\sigma_i^2}} \exp(-\frac{1}{\sigma_i^2}|\hat{s}_i - s|^2) P(\hat{s}_i = s).$$

Finally, assuming equally probable symbols, the LLR values for bit d_i^m can be computed as

$$LLR(d_i^m) = \log \frac{\sum_{s\in S_{+1}^{i,m}} \exp(-\frac{1}{\sigma_i^2}|\hat{s}_i - s|^2)}{\sum_{s\in S_{-1}^{i,m}} \exp(-\frac{1}{\sigma_i^2}|\hat{s}_i - s|^2)} \ .$$

Note that the conventional BLAST [4] neglects the decision error e_l in (2), which leads to overestimated LLR values.

IV. Simulation Results

In this section, we present the Monte Carlo simulation results for the proposed system. We adopt the MIMO-OFDM with 64 subcarriers ($F = 64$). The guard interval length is set to 16, and we assume a spatially uncorrelated 5 tap exponentially decaying channel response. Throughout simulations, one frame is assumed to consist of one OFDM symbol for simplicity. We assume that the channel state information is perfectly known to the receiver.

Fig. 2 shows the frame error rate (FER) results[1] for 3bps/Hz case with two receive antennas where QPSK is combined with a rate $\frac{3}{4}$ punctured convolutional code [2] with constraint length 7. A sphere decoding [7] can be another alternative detection scheme for the hybrid STBC. However, we do not consider the sphere decoder here since it does not generate soft values for the channel decoder. For comparison, we also plot the simulation results of the quasi-orthogonal STBC with rotated constellation [8]. Note that the quasi-orthogonal STBC requires a joint two symbol detection. As shown in the figure, the conventional decision feedback detection without error compensations exhibits an error floor due to the error propagation effect. The plot shows that more than a 3dB gain is obtained using the proposed decision feedback scheme at FER=10^{-2}. The results confirm that the proposed method significantly reduces the performance degradation due to the error propagation.

[1]Due to page limitation, one simulation set is presented. However, other simulation results indicate similar performance gains.

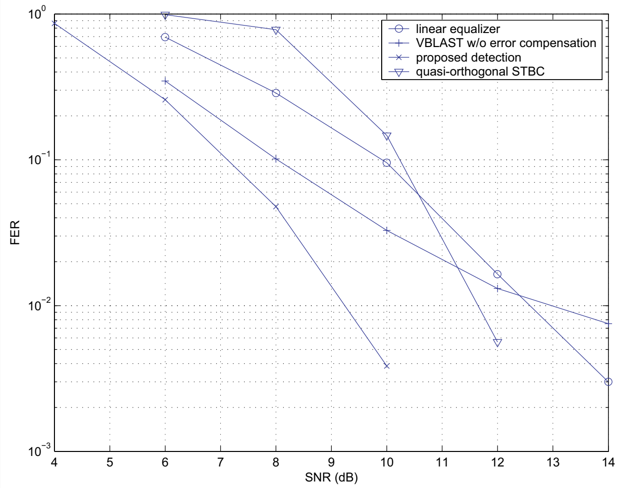

Fig. 2. Simulation results of the 4x2 antenna configuration.

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