Combining Orthogonal Space Time Block Codes with Adaptive Sub-Group Antenna Encoding

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Abstract

An adaptive space time transmit diversity scheme with simple feedback is proposed for the next generation wireless communication systems. By combining orthogonal space time block codes with adaptive sub-group antenna encoding, this new diversity scheme can effectively exploit the diversity potential provided by multiple antenna arrays without introducing interference among the signals transmitted at different antennas. In order to reduce the amount of feedback information as well as the computational complexity, a new quadrant phase constraining method is introduced to formulate the feedback information. With simple operations at both the transmitter and the receiver, the new adaptive diversity scheme out-performs not only open loop space time block encoding techniques, but also some close loop transmit diversity techniques with the same amount of feedback.

Combining Orthogonal Space Time Block Codes with Adaptive Sub-group Antenna Encoding

Jingxian Wu,[1] **Henry Horng,**[2] **Jinyun Zhang,**[2] **Chengshan Xiao,**[3,5] **and Jan C. Olivier**[4]

An adaptive space time transmit diversity scheme with simple feedback is proposed for the next generation wireless communication systems. By combining orthogonal space time block codes with adaptive sub-group antenna encoding, this new diversity scheme can effectively exploit the diversity potential provided by multiple antenna arrays without introducing interference among the signals transmitted at different antennas. In order to reduce the amount of feedback information as well as the computational complexity, a new quadrant phase constraining method is introduced to formulate the feedback information. With simple operations at both the transmitter and the receiver, the new adaptive diversity scheme outperforms not only open loop space time block encoding techniques, but also some close loop transmit diversity techniques with the same amount of feedback.

KEY WORDS: Adaptive antenna; orthogonal space time block codes; transmit diversity

1. INTRODUCTION

Transmit diversity is one of the key techniques adopted by high speed downlink packet access (HSDPA) [1] of the third generation partnership project (3GPP) standard to provide high speed reliable communication. One of the most commonly used transmit diversity techniques is the orthogonal space time block code (STBC) [2]. It is pointed out in [3] that for complex constellations, rate one STBC only exists for systems with exactly two transmit antennas.

Recently, a lot of efforts have been put into extending the STBC encoding scheme to systems with more than 2 transmit antennas without sacrificing the

coding rate [4–8]. In addition, the channel state information (CSI) is utilized by some transmit diversity systems to further improve the system performance. The CSI can be made available to the transmitter through a separate feedback channel, and such encoding schemes are called close loop techniques [9–15]. Most of the existing close loop techniques require considerable amount of information transmitted in the feedback channel, e.g., vectors with complex-valued or real-valued elements, thus a lot of bandwidth of the reverse channel will be consumed by the feedback information.

In this paper, a new adaptive space time encoding scheme is proposed for systems with more than two transmit antennas, and only a very small amount of feedback information (at the order of several bits) are required by this close loop technique. To utilize the orthogonality of the STBC encoder, all of the transmit antennas are divided into two sub-groups, with each sub-group corresponding to one of the two output streams of the STBC encoder. Adaptive space encoding vectors are employed by each of the antenna sub-group to improve the overall diversity gain as well as to keep the orthogonality between the

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signals transmitted by different antenna groups. In order to reduce the amount of feedback as well as the computational complexity, a new quadrant phase constraining method is introduced to formulate the feedback information. The amounts of feedback information can be as few as 1 bit and 2 bits for systems with 3 or 4 transmit antennas, respectively, which is of practical interests for the design of next generation wireless communication systems.

The rest of this paper is organized as follows. Section 2 presents the system structure of the proposed adaptive space-time transmit diversity scheme. In Section 3, a new adaptive quadrant phase constraining method is presented for the computation of the feedback information at the receiver. Simulation results are given in Section 4, and Section 5 concludes the paper.

2. SYSTEM STRUCTURE

We consider a system with M transmit antennas and one receive antenna. The block diagram of the baseband representation of the system is depicted in Figure 1, and only the modules directly related to the transmit diversity scheme are shown in the figure.

At the transmitter, the binary information data are first encoded and modulated, then the modulated symbols are fed into the orthogonal STBC encoder. Without loss of generality, we assume that at two consecutive symbol intervals t_1 and t_2, the input of

the STBC encoder is s_1 and s_2, respectively, where $s_j \in \mathbb{S}$, for $j = 1, 2$, with $\mathbb{S}$ being the modulation symbol set, and the energy of the modulation symbol is $E(|s_j|^2) = E_s$. At the STBC encoder, the input data symbols s_1 and s_2 are demultiplexed into two data streams, and the output of the STBC encoder can be written by

$$\mathbf{d}_1 = [d_{11} \quad d_{12}]^T = [s_1 \quad s_2^*]^T \in \mathbb{C}^{2\times1},$$
$$\mathbf{d}_2 = [d_{21} \quad d_{22}]^T = [s_2 \quad -s_1^*]^T \in \mathbb{C}^{2\times1}, \quad (1)$$

where $\mathbf{d}_k$ corresponds to the kth output stream of the STBC encoder, with d_{kj} being transmitted at the time instant t_j, $(\cdot)^*$ is the complex conjugate operation, and $(\cdot)^T$ denotes matrix transpose.

The M transmit antennas are divided into two antenna sub-groups, with each sub-group corresponding to one of the output streams $\mathbf{d}_1$ and $\mathbf{d}_2$ of the STBC encoder. We assume that the number of antennas contained in the kth group is M_k, for $k = 1$, 2, with $M_1 + M_2 = M$. Adaptive linear space encoders are employed by each of the two antenna sub-groups, and they are used to map the two data streams onto the M transmit antennas with the help of the channel feedback information. If we define the space encoding vector of the kth sub-group as

$$\mathbf{w}_k = [w_{k,1} \quad w_{k,2} \quad \cdots \quad w_{k,M_k}] \in \mathbb{C}^{1\times M_k}, \quad \text{for } k = 1, 2,$$
$$(2)$$

with the power constraint $\mathbf{w}_1\mathbf{w}_1^H + \mathbf{w}_2\mathbf{w}_2^H = 1$, then the encoded signals to be transmitted by the
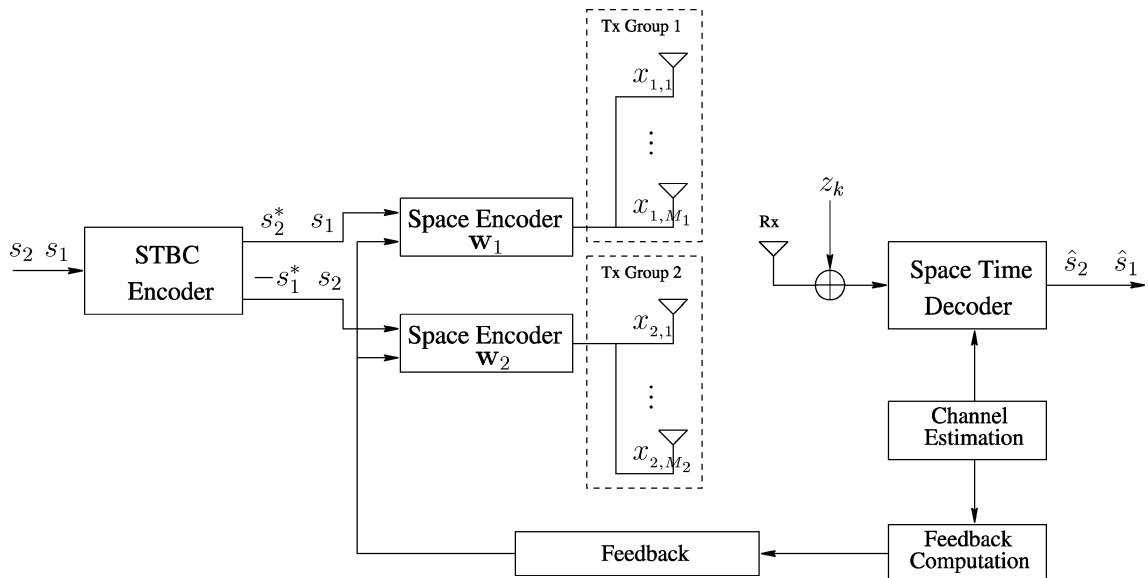

Fig. 1. The block diagram of the system with transmit diversity.

kth antenna sub-group can be written in matrix format

$$\mathbf{X}_k = \mathbf{d}_k \cdot \mathbf{w}_k \in \mathbb{C}^{2 \times M_k}, \quad \text{for } k = 1, 2, \qquad (3)$$

with the symbols on the first row of $\mathbf{X}_k$ transmitted at the symbol period t_1 and symbols on the second row transmitted at t_2.

In the channel, the transmitted signals are corrupted by both the time-varying multipath fading and additive white Gaussian noise (AWGN). The signals received by the receiver are the superposition of the signals from all the transmit antennas plus noise, and they can be represented by

$$\mathbf{r} = \begin{bmatrix} \mathbf{X}_1 & \mathbf{X}_2 \end{bmatrix} \begin{bmatrix} \mathbf{h}_1 \\ \mathbf{h}_2 \end{bmatrix} + \mathbf{z},$$
$$= \mathbf{d}_1 \mathbf{w}_1 \mathbf{h}_1 + \mathbf{d}_2 \mathbf{w}_2 \mathbf{h}_2 + \mathbf{z}, \qquad (4)$$

where $\mathbf{r} = [r_1, r_2]^T$, $\mathbf{z} = [z_1, z_2]^T$ are the receive vector and AWGN noise vector, respectively, with r_k and z_k corresponding to the time instant t_k, $\mathbf{h}_k \in \mathbb{C}^{M_k \times 1}$ is the channel impulse response (CIR) vector defined as

$$\mathbf{h}_k = \begin{bmatrix} h_{k,1} & h_{k,2} & \cdots & h_{k,M_k} \end{bmatrix}^T, \quad \text{for } k = 1, 2, \quad (5)$$

with the element $h_{k,m}$, for $m = 1, 2, \cdots, M_k$, being the CIR between the mth transmit antenna of sub-group k and the receive antenna.

Combining (1) and (4), we can rewrite the input–output relationship of the diversity system as

$$\begin{bmatrix} r_1 \\ r_2^* \end{bmatrix} = \begin{bmatrix} \mathbf{w}_1 \mathbf{h}_1 & \mathbf{w}_2 \mathbf{h}_2 \\ -\mathbf{h}_2^H \mathbf{w}_2^H & \mathbf{h}_1^H \mathbf{w}_1^H \end{bmatrix} \begin{bmatrix} s_1 \\ s_2 \end{bmatrix} + \begin{bmatrix} z_1 \\ z_2^* \end{bmatrix},$$
$$= \mathbf{H} \cdot \mathbf{s} + \tilde{\mathbf{z}}, \qquad (6)$$

where $\mathbf{s} = [s_1 s_2]^T$, $\mathbf{z} = [z_1 z_2^*]^T$ are the signal vector and noise vector, respectively, and the channel matrix $\mathbf{H}$ is defined as

$$\mathbf{H} = \begin{bmatrix} \mathbf{w}_1 \mathbf{h}_1 & \mathbf{w}_2 \mathbf{h}_2 \\ -\mathbf{h}_2^H \mathbf{w}_2^H & \mathbf{h}_1^H \mathbf{w}_1^H \end{bmatrix} \in \mathbb{C}^{2 \times 2}. \qquad (7)$$

The matrix $\mathbf{H}$ is a 2×2 orthogonal matrix, i.e., $\mathbf{H}^H \mathbf{H} = (|\mathbf{h}_1 \mathbf{w}_1|^2 + |\mathbf{h}_2 \mathbf{w}_2|^2) \cdot \mathbf{I}_2$, with $\mathbf{I}_2$ being a 2×2 identity matrix. From (6) and (7), we can compute the decision vector $\mathbf{y} = [y_1, y_2]^T$ as

$$\mathbf{y} = \mathbf{H}^H \mathbf{r},$$
$$= (|\mathbf{w}_1 \mathbf{h}_1|^2 + |\mathbf{w}_2 \mathbf{h}_2|^2) \cdot \mathbf{s} + \mathbf{v}, \qquad (8)$$

where $\mathbf{v} = \mathbf{H}^H \mathbf{z}$ is the noise component with covariance matrix $(|\mathbf{h}_1 \mathbf{w}_1|^2 + |\mathbf{h}_2 \mathbf{w}_2|^2) \cdot \mathbf{I}_2 \cdot N_0$, and $N_0 = E(|z_k|^2)$. With the decision variable given in (8), we can compute the signal to noise ratio (SNR) at the receiver as follows

$$\gamma_M = (|\mathbf{w}_1 \mathbf{h}_1|^2 + |\mathbf{w}_2 \mathbf{h}_2|^2) \cdot \gamma_0, \qquad (9)$$

where $\gamma_0 = \frac{E_s}{N_0}$ is the SNR without diversity. It can be seen from (9) that the SNR γ is a function of the space encoding vectors $\mathbf{w}_1$, $\mathbf{w}_2$ and the CIR vectors $\mathbf{h}_1$, $\mathbf{h}_2$. By choosing appropriate forms of $\mathbf{w}_k$ based on the properties of the fading channels, we can improve the receiver SNR with only a small amount of feedback.

3. ADAPTIVE SPACE ENCODING VECTOR DESIGN: QUADRANT PHASE CONSTRAINING

3.1. General Case

To achieve the maximum SNR at the receiver, the optimum design criterion for the space encoding vectors $\mathbf{w}_1$ and $\mathbf{w}_2$ is

$$(\mathbf{w}_1, \mathbf{w}_2) = \underset{(\mathbf{w}_1, \mathbf{w}_2) \in \mathcal{W}}{\operatorname{argmax}} \left\{ |\mathbf{w}_1 \mathbf{h}_1|^2 + |\mathbf{w}_2 \mathbf{h}_2|^2 \right\}, \qquad (10)$$

where $\mathcal{W}$ is the set of all the possible encoding vector pairs satisfying the power constraint $\mathbf{w}_1 \mathbf{w}_1^H + \mathbf{w}_2 \mathbf{w}_2^H = 1$. The optimum values of $\mathbf{w}_1$ and $\mathbf{w}_2$ can be obtained by exhaustively searching all the elements of the set $\mathcal{W}$. The size of the set $\mathcal{W}$ increases exponentially with the number of transmit antennas, therefore this optimum space encoding vector design method are not appropriate for systems with large number of transmit antennas.

In order to reduce the computational complexity as well as to reduce the amount of feedback information, we introduce a quadrant phase constraining method for the computation of the feedback information and the formulation of the adaptive space encoding vectors. For a general system with M transmit antennas, we let $M_1 = M_2 = \frac{M}{2}$ if M is an even number, and $M_1 = \frac{M+1}{2}$, $M_2 = \frac{M-1}{2}$ if M is an odd number. Define the space encoding vector $\mathbf{w}_k$ as

$$\mathbf{w}_k = \frac{1}{\sqrt{M}} \left[1 \exp\left(-\mathbf{i} \frac{q_{k,2}\pi}{2} \right) \cdots \exp\left(-\mathbf{i} \frac{q_{k,M_k}\pi}{2} \right) \right], \qquad (11)$$

where $\mathbf{i}^2 = -1$ is the imaginary part symbol, $q_{k,m} \in \{0, 1, 2, 3\}$ is the feedback information, and each $q_{k,m}$ contains 2 bits of information. For systems with M transmit antennas, the total number of feedback bits required by the proposed algorithm is $2M-4$. For convenience of representation, we let $q_{1,1} = q_{2,1} = 0$.

With the space encoding vectors defined in (11), we can write the corresponding SNR γ_M given in (9) as

$$\gamma_M = \frac{1}{M} \sum_{k=1}^{2} \left| \sum_{m=1}^{M_k} h_{k,m} \exp\left(-\mathbf{i} \cdot q_{k,m} \frac{\pi}{2}\right) \right|^2 \cdot \gamma_0,$$
$$= (g_c + g_b) \cdot \gamma_0, \tag{12}$$

where the diversity gains g_c and g_b are given by

$$g_c = \frac{1}{M} \sum_{m=1}^{M_1} |h_{1,m}|^2 + \frac{1}{M} \sum_{m=1}^{M_2} |h_{2,m}|^2, \tag{13a}$$

$$g_b = \frac{1}{M} \sum_{m=1}^{M_1} \sum_{n=m+1}^{M_1} \Re\left[h_{1,m} h_{1,n}^* \exp\left(-\mathbf{i}\frac{q_{1,m} - q_{1,n}}{2}\pi\right)\right] +$$
$$\frac{1}{M} \sum_{m=1}^{M_2} \sum_{n=m+1}^{M_2} \Re\left[h_{2,m} h_{2,n}^* \exp\left(-\mathbf{i}\frac{q_{2,m} - q_{2,n}}{2}\pi\right)\right], \tag{13b}$$

with $\Re(\cdot)$ being the real part operator. In the equations above, g_c is the conventional diversity gain, which is fixed for a certain number of transmit antennas; while the feedback diversity gain g_b varies with the value of the space encoding vector $\mathbf{w}_k$, which is in turn determined by the feedback information $q_{k,m}$. The value of $q_{k,m}$ can be chosen based on certain design criteria to improve the value of g_b, thus to improve the receiver SNR γ_M.

It can be seen from (13b) that the feedback diversity gain g_b is expressed in a summation form. For different feedback information $q_{k,m}$, each of the summed elements in (13b) could be either positive or negative, resulting in performance enhancement or degradation, accordingly. If all the summed elements of the feedback diversity gain are non-negative, then the feedback diversity gain is guaranteed to be non-negative. By this means, a performance gain is achieved over conventional open loop techniques.

Taking into account both the system performance and computational complexity, we are choosing a sub-optimum design criterion: *all of the summed elements of the feedback diversity gain g_b are non-negative.*

One of the summed elements of g_b can be expressed as

$$\Re\left[h_{k,m} h_{k,n}^* \exp\left(-\mathbf{i}\frac{q_{k,m} - q_{k,n}}{2}\pi\right)\right]$$
$$= |h_{k,m}||h_{k,n}| \cos(\Delta\theta_{k,mn}), \tag{14}$$

where $\Delta\theta_{k,mn} = \theta_{k,m} - \theta_{k,n} - \frac{q_{k,m} - q_{k,n}}{2}\pi$, with $\theta_{k,m} \in [0, 2\pi)$ being the phase of $h_{k,m}$. The term described in (14) will be non-negative if the following condition is satisfied

$$|\Delta\theta_{k,mn}| \leq \frac{\pi}{2}, \forall m \neq n, \tag{15}$$

which means the differences between all the possible angle pairs within each sub-group should be less than $\pi/2$. One direct way to satisfy this condition is choosing the value of $q_{k,m}$ such that all the rotated phases $\theta_{k,m} - q_{k,m}\frac{\pi}{2}$, for $m = 1, 2, \cdots, M_k$, belonging to one sub-group are in the same $\pi/2$ angle sector in a two dimensional coordinate system.

Without loss of generality, we normalize the phases of all the CIRs of the kth sub-group with the phase of $\theta_{k,1}$, and the normalized phases are represented as

$$\tilde{\theta}_{k,m} = \theta_{k,m} - \theta_{k,1} + 2l\pi, \tag{16}$$

where the integer l is chosen such that $\tilde{\theta}_{k,m}$ is in the range of $[0, 2\pi]$. The normalized phase $\tilde{\theta}_{k,m}$ is rotated clockwise by the angle of $q_{k,m}\frac{\pi}{2}$, such that the rotated angle $\tilde{\theta}_{k,m} - q_{k,m}\frac{\pi}{2}$ is in the quadrant phase sector from $[-\pi/4, \pi/4]$ of the coordinate system as shown in Figure 2. Following the discussions above, we can compute the feedback information $q_{k,m}$ as

$$q_{k,m} = \begin{cases} \lfloor \frac{\tilde{\theta}_{k,m} + \pi/4}{\pi/2} \rfloor, & \tilde{\theta}_{k,m} \in [\frac{\pi}{4}, \frac{7\pi}{4}), \\ 0, & \text{otherwise}, \end{cases} \tag{17}$$

where $\lfloor \cdot \rfloor$ returns the nearest smaller integer. An example is given in Figure 2, where $\tilde{\theta}_{k,m} = 9\pi/8$. From (17) we can get that $q_{k,m} = 2$, and the corresponding rotated angle is $\tilde{\theta}_{k,m} - q_{k,m}\frac{\pi}{2} = \pi/8$, which is in the angle sector of $[-\frac{\pi}{4}, \frac{\pi}{4}]$ of the coordinate system. By performing the same operations to all the normalized phases, the rotated phases of the same sub-group will be in the same quadrant angle sector, and the non-negativity of each summed element of the diversity gain g_b can be guaranteed. This method achieves the feedback diversity gain by constraining all the rotated phases of the CIRs of one sub-group in an angle sector of $\pi/2$, hence we call it quadrant phase constraining method. Since the value of $q_{k,m}$ is computed separately for each of the transmit antennas, the computational complexity of this algorithm increases linearly with the number of transmit antennas.

With $q_{k,m}$ computed from (17), all the summed elements given in (14) are guaranteed to be non-negative, and the feedback diversity gain of (13b) can be written by

$$\tilde{g}_b = \frac{1}{M} \sum_{k=1}^{2} \sum_{m=1}^{M_k} \sum_{n=m+1}^{M_k} |h_{k,m}||h_{k,n}^*|| \cos(\Delta\theta_{k,mn})|, \tag{18}$$
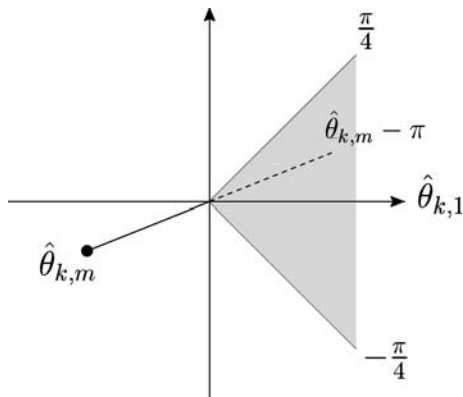
Fig. 2. The two dimensional coordinate system.

and the SNR at the receiver of the proposed algorithm is

$$\gamma_M = (g_c + \tilde{g}_b) \cdot \gamma_0, \tag{19}$$

with g_c and $\tilde{g}_b$ defined in (13a) and (18), respectively.

With the adaptive diversity algorithm presented in this subsection, $2M-4$ bits of feedback information are required to form the space encoding vectors for systems with M transmit antennas. It will be shown next that the amount of feedback information can be further reduced for systems with $M \leq 4$ transmit antennas, which are of practical interests of next generation communication systems.

3.2. Special Cases: Systems with $M \leq 4$ Transmit Antennas

For systems with $M \leq 4$ transmit antennas, each of the two antenna sub-groups will contain at most 2 transmit antennas. For sub-groups with 2 transmit antennas, the sub-optimum design criterion can be satisfied with only 1 bit of feedback information.

For systems with $M = 4$ transmit antennas, the number of antennas in each of the antenna groups is $M_1 = M_2 = 2$. We define the space encoding vector as

$$\mathbf{w}_k = \frac{1}{2}[1\,(-1)^{b_k}], \quad \text{for } k = 1, 2, \tag{20}$$

where $b_k \in \{0, 1\}$ is the feedback information for the kth antenna sub-group, and it is defined as

$$b_k = \begin{cases} 0, & \Re(h_{k,1}h_{k,2}^*) \geq 0, \\ 1, & \text{otherwise} \end{cases} \tag{21}$$

Combining (9), (20) and (21), we have the SNR at the receiver written by

$$\gamma_4 = (g_{4,c} + g_{4,b})\gamma_0, \tag{22}$$

where the conventional diversity gain $g_{4,c}$ and the feedback diversity gain $g_{4,b}$ are defined as

$$g_{4,c} = \frac{1}{4}\left(\sum_{m=1}^{2}|h_{1,m}|^2 + \sum_{m=1}^{2}|h_{2,m}|^2\right), \tag{23a}$$

$$g_{4,b} = \frac{1}{2}\sum_{k=1}^{2}\left|\Re(h_{k,1}h_{k,2}^*)\right|. \tag{23b}$$

We can see that $g_{4,c}$ given in (23a) is the same as the conventional diversity gain of (13a) when we set $M = 4$, and the feedback diversity gain $g_{4,b}$ of (23b) is always non-negative with totally 2 bits of feedback (1 bit for each sub-group) as opposed to the $2 \times 4 - 4 = 4$ bits feedback described for the general case.

Similarly, for systems with $M = 3$ antennas, we have $M_1 = 2$ and $M_2 = 1$. Since there is only 1 antenna in sub-group 2, we have $\mathbf{w}_2 = 1/\sqrt{3}$. For the first sub-group with 2 transmit antennas, we apply the space encoding vector $\mathbf{w}_1$ as defined in (20). With this encoding scheme, the receiver SNR can be computed from (9) as

$$\gamma_3 = (g_{3,c} + g_{3,b}) \cdot \gamma_0, \tag{24}$$

with the conventional diversity gain $g_{3,c}$ and feedback diversity gain $g_{3,b}$ being given by

$$g_{3,c} = \frac{1}{3}\left(\sum_{m=1}^{2}|h_{1,m}|^2 + |h_{2,1}|^2\right), \tag{25a}$$

$$g_{3,b} = \frac{2}{3}\left|\Re(h_{1,1}h_{1,2}^*)\right|. \tag{25b}$$

When there are only two transmit antennas present in the system, we have $w_1 = w_2 = 1/\sqrt{2}$, and this scheme reduced to orthogonal space time block code.

With the method presented in this subsection, we only need 1 bit and 2 bits feedback information for systems with $M = 3$ and $M = 4$ transmit antennas, respectively.

4. SIMULATION RESULTS

Simulations are carried out for 3GPP HSDPA systems to evaluate the performance of the proposed adaptive diversity scheme. The simulation parameters are defined according to the HSDPA technical specifications [1] and are shown in Table I. In the simulation, the fading channels are assumed to be

Table I. Simulation Parameters for 3GPP HSDPA Systems

Carrier Frequency	2 GHz
Spreading Factor (SF)	16
Number of Multicodes (N_c)	10
Frame Length	2 m
Chip Rate (R_c)	3.84 Mbps
CPICH power	10% I_{or}
E_c/I_{or}	70%
I_{or}/I_{oc}	variable
Channel Coding	Turbo, rate $R_t = 3/4$
Fading Model	one path Rayleigh
Correlation Model	i.i.d.
Channel Estimation	perfect
Modulation	64 QAM
Feedback Delay	4 TTI

identically independent distributed, even though the method works well under correlated fading scenario.

The frame error rate (FER) performance for systems with 3 transmit antennas and 4 transmit antennas is shown in Figures 3 and 4, respectively. For comparison purpose, the performance of orthogonal space time block coding with 2 transmit antennas are also depicted in these two figures. Comparing the performance of the proposed algorithms with optimum quantized TxAA scheme [11] with the same amount of feedback, we can see that for FER = 1%, the performance improvements are approximately 4 dB for systems with 3 transmit antennas, and 2 dB

for systems with 4 transmit antennas. Moreover, the performance differences between the two space encoding methods of Section 3.1 and 3.2 are within the range of 0.5 dB.

5. CONCLUSIONS

An adaptive space time transmit diversity scheme is proposed for next generation wireless communication systems. Combining an orthogonal space time block code with adaptive sub-group antenna encoding, this close loop diversity scheme can be used for systems with more than two transmit antennas without introducing interference among the transmitted signals. A new quadrant phase constraining method is proposed for the computation of the feedback information and the formulation of the space encoding vectors. For general systems with M transmit antennas, the amount of feedback information required by the encoders are $2M-4$ bits. For the special cases of systems with 3 and 4 transmit antennas, which are of practical interests for emerging wireless communication systems, the size of feedback information can be as few as 1 bit and 2 bits, respectively. Simulation results show that this new diversity scheme outperforms not only open loop diversity techniques, but also some close loop diver-
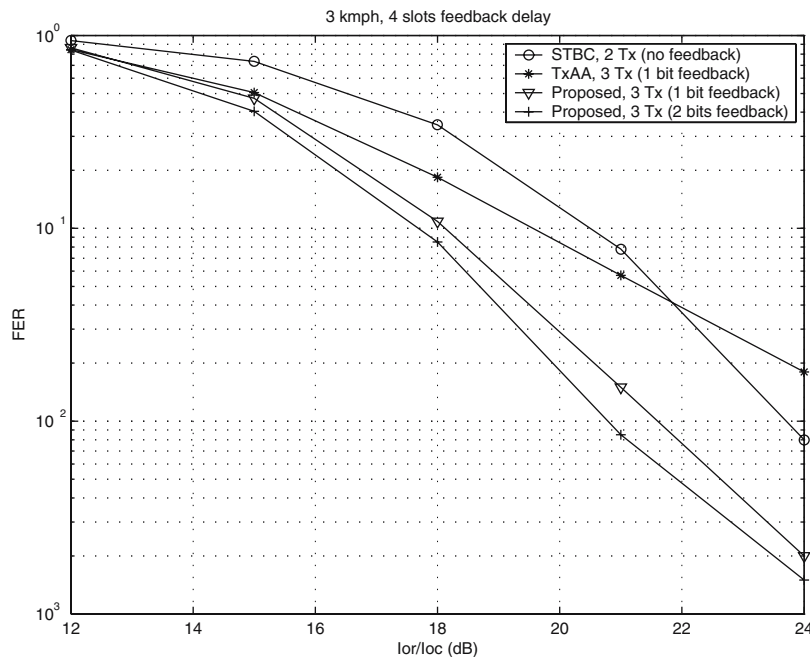

Fig. 3. Comparison of various transmit diversity schemes for systems with 3 transmit antennas.
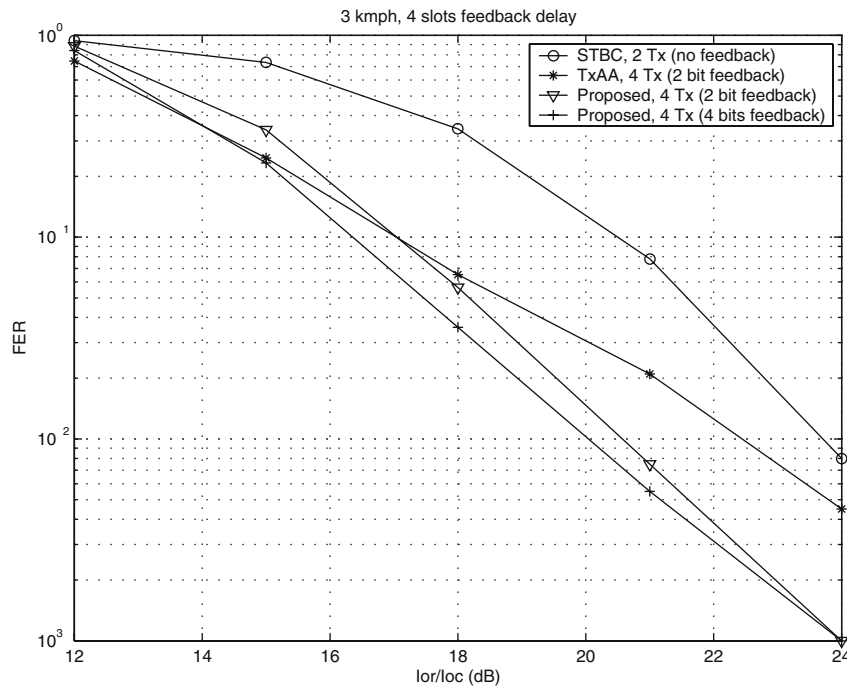
Fig. 4. Comparison of various transmit diversity schemes for systems with 4 transmit antennas.
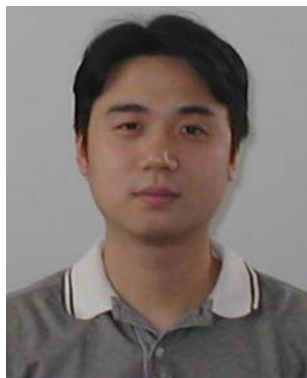
sity techniques with the same amount of feedback information.

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