MITSUBISHI ELECTRIC RESEARCH LABORATORIES
http://www.merl.com

Adaptive Transmit Diversity with Quadrant Phase Constraining Feedback

Jingxian Wu, Henry Horng, Jinyun Zhang, Andreas Molisch, Chengshan Xiao

TR2004-109   December 2004

Abstract

An adaptive transmit scheme with quadrant phase constraining feedback is proposed in this paper. With simple linear operations at both transmitter and receiver, the proposed algorithm can achieve better system performances with only 2M - 2 bits of feedback information for systems with M transmit antennas. Theoretical performance bounds of the proposed transmit diversity scheme are derived. Simulation examples and theoretical analyses show that the proposed transmit diversity scheme outperforms not only the conventional open-loop transmit diversity techniques, but also some closed-loop transmit diversity techniques with more information transmitted in the feedback channel.

PIMRC 2004

Adaptive Transmit Diversity with Quadrant Phase Constraining Feedback

Jingxian Wu[†], Henry Horng[‡], Jinyun Zhang[‡], Andreas Molisch[‡], and Chengshan Xiao[†]
[†]Department of ECE, University of Missouri, Columbia, MO 65211, USA
[‡]Mitsubishi Electric Research Laboratories, Cambridge, MA 02139, USA

***Abstract*— An adaptive transmit diversity scheme with quadrant phase constraining feedback is proposed in this paper. With simple linear operations at both transmitter and receiver, the proposed algorithm can achieve better system performances with only $2M-2$ bits of feedback information for systems with M transmit antennas. Theoretical performance bounds of the proposed transmit diversity scheme are derived. Simulation examples and theoretical analyses show that the proposed transmit diversity scheme outperforms not only the conventional open-loop transmit diversity techniques, but also some closed-loop transmit diversity techniques with more information transmitted in the feedback channel.**

I. Introduction

The next generation wireless communication systems are required to provide high quality voice services as well as broadband data services with rates far beyond the limitations of current wireless systems. To achieve this goal, one of the key technologies employed by emerging wireless systems is transmit diversity.

The most commonly used transmit diversity technique is orthogonal space time block code (STBC) [1], which can achieve full diversity order without any knowledge of the fading channel, and it is classified as open-loop transmit diversity technique. On the other hand, partial or perfect information of the fading channel, if available at the transmitter through separate feedback channel (closed-loop), can be utilized to further improve the performance of the diversity system [3]-[7]. Most of the closed-loop diversity techniques require the feedback information to be complex-valued matrices or vectors with elements either directly being the channel impulse response (CIR), or some statistics (*e.g.*, mean or covariance) of the CIR. It's apparent that considerable bandwidth in the reverse link will be consumed by the feedback information.

To overcome this problem, optimum quantization [6] are applied in transmit antenna array (TxAA) to reduce the amount of feedback information, and it will show in this paper that the computational complexity of optimum quantization increases *exponentially* with the number of antennas and quantization bits. Another closed-loop transmit diversity with less feedback information is introduced in [7] as adaptive space time transmit diversity (ASTTD), and it is extended from the open-loop STBC, therefore it can only be used for systems with two transmit antennas for full rate systems [2].

In this paper, a simple adaptive transmit diversity scheme with quadrant phase constraining feedback is proposed for systems with arbitrary number of transmit antennas, and the objective of this diversity technique is to achieve better system performance with less feedback information and less computations. It is observed in this paper that for systems with M transmit antennas, $2M-2$ bits of feedback information will guarantee the performance improvement over corresponding open loop transmit diversity systems. A new quadrant phase constraining method is introduced for the computation of the feedback information, and the computational complexity is only *linearly* proportional to the number of transmit antennas, as opposed to the *exponentially* increasing complexity of TxAA with optimum quantization [6]. Based on the statistical properties of the received signals, theoretical error probability bounds are derived for the proposed transmit diversity technique.

Theoretical analyses and simulation results show that our binary adaptive transmit diversity technique outperforms not only the conventional open-loop diversity techniques, but also some of the closed-loop techniques with more information transmitted in the feedback channel.

II. System Model

We consider a system with M transmit antennas and one receive antenna, and the block diagram of the baseband representation of the diversity system is shown in Fig. 1.

At the time instant k, the modulated symbol $s_k \in \mathcal{S}$ with unit energy is linearly pre-encoded in the space domain by the space encoding vector $\mathbf{p}_k = [p_1(k), p_2(k), \cdots, p_M(k)] \in \mathbb{C}^{1\times M}$ at the transmitter, where $\mathcal{S}$ is the modulation symbol set. The encoded transmit vector is $\mathbf{x}_k = [x_1(k), x_2(k), \cdots, x_M(k)] = \mathbf{p}_k \cdot s_k$, with $x_m(k)$ being transmitted at the mth transmit antenna. In the adaptive transmit diversity described in this paper, the space encoding vector $\mathbf{p}_k$ is determined by the feedback information sent back from the receiver.

At the receiver, the received signals are the sum of the propagation signals from all the transmit antennas plus additive noise, and the samples at the receiver can be expressed by

$$r(k) = \sqrt{\frac{E_s}{M}} \cdot \mathbf{x}_k \mathbf{h}_k + z_k = \sqrt{\frac{E_s}{M}} \cdot (\mathbf{p}_k \mathbf{h}_k) \cdot s_k + z_k, \qquad (1)$$

where E_s is the total transmit energy of all the transmit antennas, z_k is the additive white Gaussian noise (AWGN) with variance $N_0/2$ per dimension, $\mathbf{h}_k = [h_1(k), h_2(k), \cdots, h_M(k)]^T \in \mathbb{C}^{M\times 1}$ is the time-varying channel impulse response (CIR) of the fading channels, with $h_m(k)$ being the CIR of the fading channel between the mth transmit antenna and the receive antenna, and $(\cdot)^T$ denotes matrix transpose.

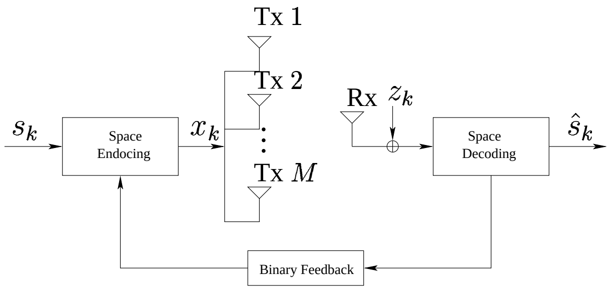

Fig. 1. The block diagram of the system with transmit diversity.

With the system model defined in (1), it is well known that the optimum precoding vector $\hat{\mathbf{p}}_k$ that maximizes the output SNR is [5],

$$\hat{\mathbf{p}}_k = \frac{\mathbf{h}_k^H}{\mathbf{h}_k \mathbf{h}_k^H}, \tag{2}$$

where $(\cdot)^H$ denotes matrix Hermitian. Forming this optimum vector requires perfect knowledge of the channel CIR vector $\mathbf{h}_k$, which contains $2M$ real-valued scalars. The amount of feedback can be reduced with optimum quantization [6], which chooses the quantized space encoding vector $\hat{\mathbf{p}}_k$ as follows

$$c = \underset{\hat{\mathbf{p}}_k \in \mathcal{P}}{\text{argmin}}\; \hat{\mathbf{p}}_k \mathbf{h}_k \mathbf{h}_k^H \hat{\mathbf{p}}_k^H, \tag{3}$$

where $\mathcal{P}$ is the set of all the possible quantized space encoding vectors, and it contains $2^{b(M-1)}$ possible vectors for systems with b bits quantization. In order to find the optimum quantization vector, the receiver must exhaustively compute the values of $\hat{\mathbf{p}}_k \mathbf{h}_k \mathbf{h}_k^H \hat{\mathbf{p}}_k^H$ for all the $2^{b(M-1)}$ possible vectors to choose the optimum one, and each computation involves approximately M^2 *complex* multiplications. Therefore the total amount of computational complexity incurred by the feedback information alone is in the order of $O(2^{b(M-1)} \times 2 \times M^2)$, which is quite considerable when the number of antennas is larger than 2.

To balance the system performance, the size of feedback information, and the system computational complexity, a suboptimum adaptive transmit diversity scheme with quadrant phase constraining feedback is discussed in the next section.

III. Adaptive Transmit Diversity with Quadrant Phase Constraining Feedback

For systems with M transmit antennas, we define the space encoding vector as

$$\mathbf{p}_k = \left[\begin{array}{cccc} 1 & \exp[\frac{i \cdot q_2(k)\pi}{2}] & \cdots & \exp[\frac{i \cdot q_M(k)\pi}{2}] \end{array}\right], \tag{4}$$

where $i^2 = -1$, and $q_m(k) \in \{0, 1, 2, 3\}$, for $m = 2, 3, \cdots, M$, are the feedback information from the receiver. For consistence of representation, we let $q_1(k) = 0$, $\forall k$. By such definitions, each $q_m(k)$ contains two bits of information, and there are totally $2(M-1)$ bits feedback information used to form the space encoding vector $\mathbf{p}_k$.

Combining (1) and (4), we can write the received sample $r(k)$ as

$$r(k) = \sqrt{\frac{E_s}{M}} \left\{ \sum_{m=1}^{M} \exp\left[\frac{i \cdot q_m(k)\pi}{2}\right] h_m(k) \right\} \cdot s_k + z_k. \tag{5}$$

At the decoder, the decision variable $y(k)$ is obtained by multiplying the received sample $r(k)$ with $(\mathbf{p}_k \mathbf{h}_k)^H$, and it can be written by

$$y(k) = \sqrt{\frac{E_s}{M}} \left| \sum_{m=1}^{M} \exp\left[\frac{i \cdot q_m(k)\pi}{2}\right] h_m(k) \right|^2 \cdot s_k + v_k, \tag{6}$$

where $v_k = (\mathbf{p}_k \mathbf{h}_k)^H z_k$ is the noise component with variance $|\mathbf{p}_k \mathbf{h}_k|^2 N_0$. According to (6), the instantaneous SNR of the decision variable $y(k)$ is

$$\gamma = \gamma_0 \cdot (g_c + g_b), \tag{7}$$

where $\gamma_0 = \frac{E_s}{N_0}$ is the SNR without diversity, and g_c and g_b are defined as

$$g_c = \frac{1}{M} \sum_{m=1}^{M} |h_m(k)|^2, \tag{8a}$$

$$g_b = \frac{2}{M} \sum_{m=1}^{M} \sum_{n=m+1}^{M} \Re \left\{ h_m(k) h_n^*(k) \exp\left[i \cdot \pi \frac{q_m(k) - q_n(k)}{2} \right] \right\}, \tag{8b}$$

where $\Re(\cdot)$ is the real part operator. In (8), g_c and g_b represent the SNR improvement contributed by the transmit diversity, and they are called diversity gains. As defined in (8a), g_c is the conventional diversity gain, which is fixed for a certain value of M, while the feedback diversity gain g_b is the extra gain contributed by te feedback information. By choosing appropriate value of the feedback information $q_m(k)$, we can improve the values of the feedback diversity gain g_b, thus the overall system performance.

We choose $q_m(k)$ such that all the summed elements of g_b are non-negative. According to (8b), one of the summed elements of g_b can be written by

$$\Re \left\{ h_m(k) h_n^*(k) \exp\left[i \cdot \pi \frac{q_m(k) - q_n(k)}{2} \right] \right\} = |h_m(k)||h_n(k)| \cos(\Delta\theta_{mn}), \tag{9}$$

where $\Delta\theta_{mn} = \theta_m - \theta_n + \frac{q_m(k) - q_n(k)}{2}\pi$ with $\theta_m \in [0, 2\pi)$ being the phase of $h_m(k)$. The terms described in (9) will be non-negative if the following condition is satisfied

$$|\Delta\theta_{mn}| \leq \pi/2, \qquad \forall m \neq n. \tag{10}$$

One direct way to satisfy the condition of (10) is adjusting $q_m(k)$ so that all the rotated phases $\theta_m + \frac{q_m(k)}{2}\pi$, for $m = 1, 2, \cdots, M$,

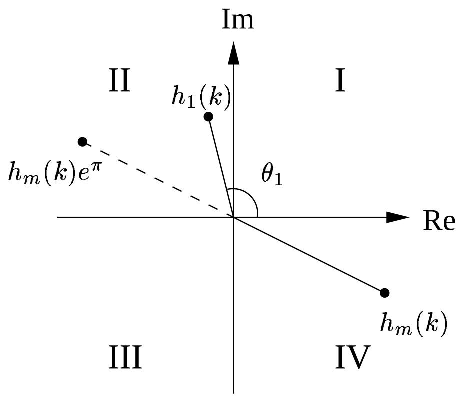

Fig. 2. The four quadrants of the coordinate system.

are in the same coordinate quadrant, therefore we call this method quadrant phase constraining. Without loss of generality, we keep the phase θ_1 of the first sub-channel $h_1(k)$ unchanged, and rotates the phases θ_m of all the rest sub-channels by $\frac{q_m(k)}{2}\pi$ so that the rotated phases are in the same quadrant as θ_1. Since there are four quadrants, two bits information of each $q_m(k)$ are enough to achieve this goal.

If we label the four quadrants of the Cartesian coordinate system as shown in Fig. 2, then the quadrant number of any angle $\phi \in [0, 2\pi)$ can be computed as $\lceil \frac{2\phi}{\pi} \rceil$, where $\lceil \cdot \rceil$ denotes rounding to the nearest bigger integer. With the analyses above, the feedback information $q_m(k)$ for $m = 2, 3, \cdots, M$ can be computed at the receiver based on the phases of the estimated CIRs

$$q_m(k) = \left(\left\lceil \frac{2\theta_1}{\pi} \right\rceil - \left\lceil \frac{2\theta_m}{\pi} \right\rceil \right)_4, \tag{11}$$

where $(a)_N$ denotes the residue of $\frac{a}{N}$ with $(a)_N \in [0, N-1]$. An example is shown in Fig. 2 with θ_1 in quadrant II and θ_m in quadrant IV. With (11), we can get that $q_m(k) = 2$, which corresponds to rotate θ_m by π radians count-clockwise, and the rotated phase $\theta_m + \frac{q_m(k)}{2}\pi$ is in quadrant II.

The feedback information defined in (11) guarantees that all the elements described in (9) are non-negative $\forall m \neq n$, and the diversity gain g_b contributed by the feedback information can be written by

$$g_b = \frac{2}{M} \sum_{m=1}^{M} \sum_{n=m+1}^{M} |h_m(k)||h_n(k)|| \cos(\Delta\theta_{mn})|. \tag{12}$$

Combining (6), (8) and (12), we will have the output SNR γ at the detector

$$\gamma = \gamma_0 \cdot \left[\frac{1}{M} \sum_{m=1}^{M} |h_m(k)|^2 + \frac{2}{M} \sum_{m=1}^{M} \sum_{n=m+1}^{M} |h_m(k)||h_n(k)|| \cos(\Delta\theta_{mn})| \right]. \tag{13}$$

When $M = 2$, we can see that the conventional diversity gain g_c is exactly the same as the diversity gain of the orthogonal STBC encoder. The quadrant phase constraining feedback proposed in this paper will guarantee that $g_b \geq 0$, thus the proposed adaptive transmit diversity will consistently outperform the conventional orthogonal STBC. Moreover, full rate orthogonal STBC can only be implemented for systems with 2 transmit antennas, while the proposed diversity scheme can be implemented for systems with arbitrary number of transmit antennas. This is extremely useful for high speed downlink data transmission of next generation wireless communication systems, where higher diversity order are required to guarantee high data throughput in the downlink with multiple transmit antennas and one receive antenna.

Moreover, the computational complexity of the proposed algorithm is much lower than TxAA with optimum quantization. The feedback information of the proposed diversity scheme is computed for each antenna separately, therefore the computational complexity of our algorithm is *linearly* proportional to the number of transmit antennas. Moreover, the computation of each of the feedback coefficient $p_m(k)$ involves approximately 2 *real* multiplications. Therefore the computational complexity of the proposed algorithm is in the order of $O((M-1) \times 2)$. According to the analyses above, we define the ratio η between our diversity scheme and TxAA with optimum quantization as

$$\eta = \frac{(M-1)}{2^{b(M-1)} \times M^2} \times 100\%. \tag{14}$$

For systems with M = 4 Tx antennas and 2 bits representation of each element of the space encoding vector, we have $\eta = 0.3\%$, which means the proposed transmit diversity scheme needs only 0.3% of computation of TxAA with optimum quantization to achieve similar performance. Even larger computational saving can be observed for systems with more antennas.

IV. Performance Analysis

In this section, theoretical performances bounds of the proposed transmit diversity scheme with binary phase shift keying (BPSK) modulation in Rayleigh fading channels are analyzed.

The error probability of a system over fading channels can be computed by

$$P(E) = \int_0^{+\infty} P(E|\gamma) d\gamma, \tag{15}$$

where $P(E|\gamma)$ is the conditional error probability (CEP), and γ is the instantaneous output SNR. For systems with BPSK modulation, the CEP can be written by [8, eqn. (23)]

$$P(E|\gamma) = \frac{1}{\pi} \int_0^{\frac{\pi}{2}} \exp\left[-\frac{\gamma}{\sin^2\theta} \right] d\theta. \tag{16}$$

With the expressions of γ given in (13), it is difficult to get the closed form expressions of the the probability density function (pdf) of γ. Therefore, we resort to the upper and lower bound of the error probability instead of directly evaluating the exact error probability.

To obtain the upper bound of the error probability, we simply set the feedback diversity gain $g_b = 0$, and the corresponding SNR γ^U becomes

$$\gamma^U = \frac{\gamma_0}{M} \sum_{m=1}^{M} |h_{k_m}|^2. \tag{17}$$

The SNR γ^U scaled by the factor M has similar form as the output SNR of maximal ratio combiner (MRC). For Rayleigh fading channel, γ^U is χ^2 distributed with $2M$ degrees of freedom, and the upper bound error probability $P^U(E)$ is [8, eqn. (23)]

$$P^U(E) = \frac{1}{\pi} \int_0^{\frac{\pi}{2}} \prod_{m=1}^{M} \left(1 + \frac{\bar{\gamma}_m}{\sin^2\theta}\right)^{-1} d\theta, \tag{18}$$

where $\bar{\gamma}_m$ is the average SNR of the mth sub-channel given by

$$\bar{\gamma}_m = \frac{\gamma_0}{M} \cdot E(|h_m(k)|^2) = \frac{\gamma_0}{M} \cdot \sigma_m^2, \tag{19}$$

with σ_m^2 being the variance of the CIR $h_m(k)$.

Next we are going to find the lower bound of the system error probability $P^L(E)$ with the help of the output SNR. The feedback diversity gain of the proposed transmit diversity scheme is given in (12), and it can be easily seen that

$$g_b \leq \frac{2}{M} \cdot \sum_{m=1}^{M} \sum_{n=m+1}^{M} |h_m(k)||h_n(k)|. \tag{20}$$

Based on (20), we define the SNR corresponding to the lower bound error probability as

$$\begin{aligned} \gamma^L &= \frac{\gamma_0}{M} \cdot \left[\sum_{m=1}^{M} |h_m(k)|^2 + 2 \cdot \sum_{m=1}^{M} \sum_{n=m+1}^{M} |h_m(k)||h_n(k)| \right], \\ &= \frac{\gamma_0}{M} \cdot \left[\sum_{m=1}^{M} |h_m(k)| \right]^2. \end{aligned} \tag{21}$$

It is interesting to note that the SNR γ^L has exactly the same form as the output SNR of coherent equal gain combiner (EGC) [9, eqn. (1)] with M receive antennas and one transmit antenna. Therefore, the lower bound error probability $P^L(E)$ can be expressed in the same form as the error probability expression for EGC receivers [9, eqn. (20)]

$$P^L(E) = \frac{1}{2} - \frac{1}{\pi} \int_0^{\pi/2} \frac{\exp(-\tan\theta)}{\sin 2\theta} \Im\left\{\phi(2\sqrt{\tan\theta})\right\} d\theta, \tag{22}$$

where $\Im(\cdot)$ is the imaginary part operator, the function $\phi(x)$ is defined as

$$\phi(x) = \prod_{m=1}^{M} \left[{}_1F_1\left(1, \frac{1}{2}; -\frac{\bar{\gamma}_m x^2}{4}\right) + i \cdot x \sqrt{\frac{\pi \bar{\gamma}_m}{4}} \exp\left(-\frac{\bar{\gamma}_m x^2}{4}\right) \right], \tag{23}$$

${}_1F_1(a, b; c)$ is the confluent hypergeometric function of the first kind, and $\bar{\gamma}_m$ is the average SNR of the mth sub-channel defined in (19).

With the upper bound and lower bound error probability given in (18) and (22), the error probability $P(E)$ of the proposed algorithms in uncoded communication systems can be expressed by

$$P^L(E) \leq P(E) \leq P^U(E). \tag{24}$$

V. Simulation Examples

Simulation examples for both uncoded systems and 3GPP high speed downlink data packet access (HSDPA) systems [10] are given in this section to evaluate the performance of the proposed diversity scheme.

In Fig. 3 are shown the uncoded BER performances of various transmit diversity schemes for BPSK systems with 2 transmit antennas and 1 receive antenna. It is assumed that there is no feedback delay in this example. It can be seen from Fig. 3 that the proposed algorithm outperforms the standard orthogonal STBC up to 2 dB. Moreover, we can see that the theoretical lower bound presented in (22) is only 0.2 dB away from the simulation results, which means the lower bound is very tight. It should be noted that $P^L(E)$ is also the lower bound for TxAA with optimum quantization. Therefore our algorithm has nearly the same performance as optimum quantization, but with much less computation.

In Fig. 3 we also show the BER performances of the closed-loop ASTTD [7] with 2 real-valued scalars as feedback. With less feedback, our algorithm is better than ASTTD for about 0.8 dB. This example demonstrates that our adaptive transmit diversity scheme outperforms not only conventional open-loop transmit diversity techniques, but also closed-loop algorithm which require more information in the feedback channel.

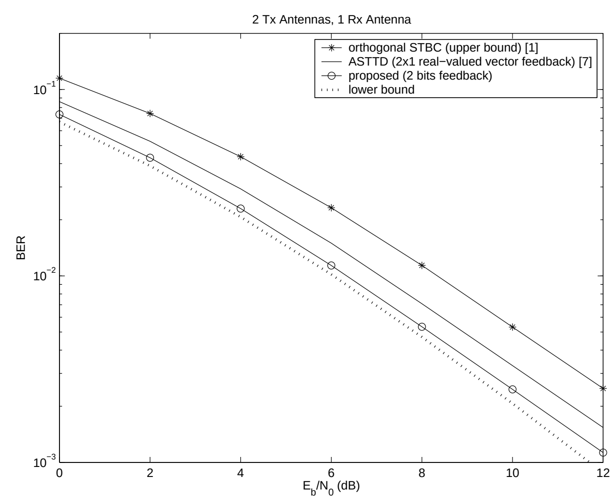

Fig. 3. Comparison of various transmit diversity schemes for systems with 2 transmit antennas

STTD and ASTTD can be implemented for systems with exactly 2 transmit antennas. However, the new adaptive transmit

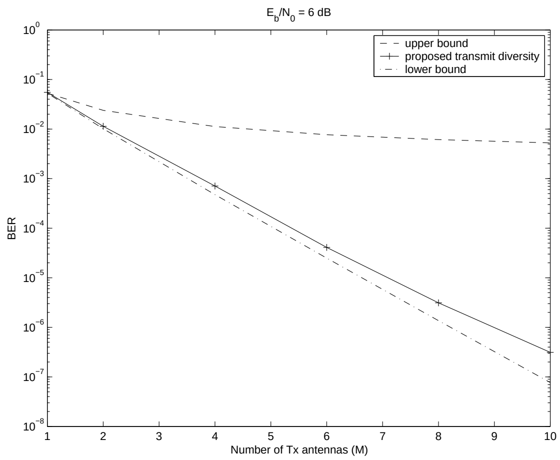

Fig. 4. The effects of transmit antenna number on the performance of the proposed algorithm.

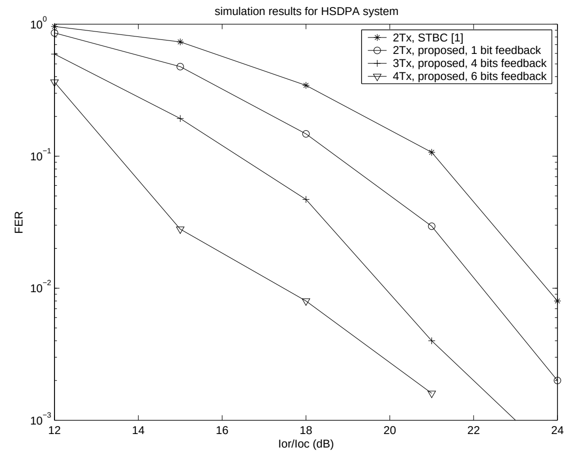

Fig. 5. Performances of transmit diversity in turbo coded HSDPA system.

diversity algorithm can be used for systems with arbitrary number of transmit antennas. Fig. 4 shows the BER performances of the transmit diversity systems with different number of transmit antennas. From the figure, we can see that the performance of the proposed algorithm improves almost linearly with the increasing number of transmit antennas. It is worth pointing out that the theoretical upper bound corresponds to the performance of orthogonal STBC. Even real orthogonal design exists for $M > 2$ (*e.g.*, $M = 4$, $M = 8$, [2]), it can be seen from Fig. 4 that no apparent extra performance gain can be achieved with orthogonal STBC when $M > 4$. However, with the algorithm proposed in this paper, the system performance improves almost linearly with the number of transmit antennas.

The next example is used to illustrate the performance of the algorithms in practical High Speed Downlink Packet Access (HSDPA) WCDMA systems [10]. The simulation parameters are defined according to the HSDPA technical specifications [10] and are shown in Table 1. The frame error rate (FER) of different transmit diversity algorithms in HSDPA systems are shown in Fig. 5, and the results are consistent with those obtained from uncoded TDMA systems. This example justifies the implementation of the newly proposed algorithms in practical communication systems.

Table 1. Simulation Parameters for 3GPP HSDPA Systems

Carrier Frequency	2 GHz
Spreading Factor (SF)	16
Number of Multicodes (N_c)	10
Frame Length	2ms
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
Data Rate	10.8Mbps

VI. CONCLUSIONS

In this paper, a novel adaptive transmit diversity scheme with quadrant phase constraining feedback is proposed for wireless communication systems. The algorithm proposed in this paper can achieve better system performance with less feedback information and less computational complexity. Simulation results show that the performance of the proposed algorithms is up to 2 dB better than the performance of open-loop STBC. When compared with other closed-loop techniques, the performance is better than or nearly the same as the performance of some closed-loop algorithms, even much less feedback information and computations are required by our approach.

REFERENCES

[1] S.M. Alamouti, "A simple transmit diversity technique for wireless communications," *IEEE J. Select. Area Commun.*, vol.16, pp.1451-1458, Oct. 1998.
[2] V. Tarokh, H. Jafarkhani, and A.R. Calderbank, "Space-time block codes from orthogonal designs," *IEEE Trans. Info. Theory*, vol.45, pp.1456-1467, Jul. 1999.
[3] G. Jongren, M. Skoglund, and B. Ottersten, "Combining beamforming and orthogonal space-time block coding," *IEEE Trans. Info. Theory*, vol.48, pp.611-627, Mar. 2002.
[4] S. Zhou, G.B. Giannakis, "Optimal transmitter eigen-beamforming and space-time block coding based on channel mean feedback," *IEEE Trans. Signal Processing*, vol.50, pp.2599-2613, Oct. 2002.
[5] K. Rohani, M. Harrison, and K. Kuchi, "A comparison of base station transmit diversity methods for third generation cellular standards," in *Proc. IEEE Veh. Techno. Conf. VTC'99 Spring*, pp.351-355, May, 1999.
[6] R.W. Heath Jr., and A. Paulraj, "A simple scheme for transmit diversity using partial channel feedback," in *Proc. IEEE 32nd Asilomar Conf. Signals, Systems and Computers.*, pp.1073-1078, 1998.
[7] Huawei:TSGR1-02-0711 "STTD with adaptive transmitted power allocation," May, 2002. (available at http://www.3gpp.org)
[8] M.K. Simon, and M.-S. Alouini, "A unified approach to the performance analysis of digital communication over generalized fading channels," *Proc. of IEEE*, vol.86, pp.1860-1877, Sep. 1998.
[9] A. Annamalai, C. Tellambura, and V.K. Bhargava, "Equal-gain diversity receiver performance in wireless channels," *IEEE Trans. Commun.*, vol.48, pp.1732-1745, Oct. 2000.
[10] 3GPP TR25.848 v4.0.0, "3GPP technical report: Physical layer aspects of ultra high speed downlink packet access,", Mar. 2001.