# Coded Error Performance of MIMO Systems in Frequency Selective Nakagami Fading Channels

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***Abstract* - In this paper, we have analyzed the convolutional coded error performance of a 2D-RAKE receiver, in combination with transmit diversity on the downlink of a WCDMA system. The analyses assume correlated fading between receive antenna array elements, and an arbitrary number of independent but nonidentical resolvable multipaths combined by the RAKE receiver in the general Nakagami-*m* fading channel framework. The closed form expression of pairwise error probability is given in the simple form of a single finite limit integral, with the integrand being an elementary function of array configuration parameters, spatial correlation and operating environment factors. It is shown that the combination of coding and spatial-path combining lead to dramatic performance improvement in various fading environments.**

## I. INTRODUCTION

Multiple-input-multiple-output (MIMO) systems are a natural extension of developments in antenna array communications. The array gain and spatial diversity obtained by utilizing multiple receive antennas have been thoroughly investigated both theoretically [1]-[4] and experimentally, but the advantages of MIMO communications which exploit the physical channels between multiple transmit and receive antennas, denoted as $(M_T, M_R)$, can provide further performance improvements. It has been shown theoretically [6], [7] that for Rayleigh fading channels the capacity of a multiple antenna system increases linearly with the number of antennas. In addition to the capacity increase, Bjerke, et. al. [8] presented the BER performance of maximal-ratio combining (MRC) or selection receive antenna combining, in combination with transmit diversity, for a (2, 2) WCDMA systems in Rayleigh fading.

Forward error correction (FEC) coding provides an alternative method to improve the performance of communications on wireless fading channels. Performance analysis of coded CDMA systems in Rayleigh fading appeared in [9]-[11]. In particular, Diaz and Agusti [9] presented closed form analytical BER expressions achieved in a coherent BPSK DS-CDMA system for any power delay profile and for either selection combining or MRC in independent Rayleigh fading assuming distinct path SNR's. The concatenation of Reed-Solomon and convolutional codes in asynchronous CDMA with selection antenna diversity is studied in [11], and the trade-off analysis among various system parameters under a fixed bandwidth expansion and a concatenated code constraint requirement is provided. BER performance and capacity assessment of a DS-CDMA system with RAKE reception and convolutional coding under frequency-selective Nakagami fading are given in [12], based on the hard-decision Viterbi decoding. In [13], tight upper bounds on the BER of convolutional codes with soft-decision decoding over independent Nakagami, Rayleigh and Rician fading multipath channels are evaluated. However, the analysis doesn't consider space diversity and the bound for the Nakagami fading case is obtained under the constraint that the ratio of the fading severity parameter and the average path power is the same for all resolvable paths.

We will first extend the analysis in [8] to the general Nakagami fading environment with an arbitrary number of TX and RX antennas. Adopting the Moment Generating Function approach [2], the uncoded BER expression is obtained in the form of a one-fold finite-limit integral, with the integrand being a function of the array configuration, the channel covariance matrix and the operating environment factors. Assuming perfect interleaving and soft-decision maximum-likelihood Viterbi decoding, the exact pairwise error probability is then derived using the alternative expression of the Gaussian Q-function [2]. Since the pairwise error probability is exact, our transfer function bound is tighter than that in [13] for the independent fading case, and more general in the framework of MIMO systems.

The paper is organized as follows. In Section II, we will briefly describe the wideband MIMO channel. Next, the bit error probability of a 2D-RAKE receiver at the mobile terminal with an arbitrary number of TX and RX antenna elements, is presented. The transfer function bound of convolutional coded BER is derived in Section IV. Section V presents several numerical examples to demonstrate the flexibility of evaluating the impact of individual design parameters on the BER performance. In addition, a trade-off analysis between the diversity gain and coding gain is presented. Concluding remarks are given in Section VI.

## II. WIDEBAND MIMO CHANNEL

We consider the downlink of a WCDMA system, with $M_R$ antennas at the mobile station (MS) and $M_T$ antennas at the base station (BS). The general wideband MIMO channel

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can be modeled as a tapped delay line. Assuming there are $L$ multipath components, the equivalent low-pass vector channel is expressed as

$$\mathbf{H}(\tau) = \sum_{l=0}^{L-1} \delta(\tau - \tau_l)\mathbf{h}^{(l)} \quad , \tag{1}$$

where

$$\mathbf{h}^{(l)} = \begin{bmatrix} \mathbf{h}_1^{(l)} & \mathbf{h}_2^{(l)} & \cdots & \mathbf{h}_{M_T}^{(l)} \end{bmatrix} \tag{2}$$

contains $M_T$ SIMO channels $\mathbf{h}_i^{(l)}$, which define the channel through which the signal is transmitted from TX antenna $i$ to all RX antennas at delay $\tau_l$, and each element $h_{ji}^{(l)}$ is the complex channel coefficient from the transmit antenna $i$ at the BS to the receive antenna $j$ at MS. We assume that the paths between each transmit antenna and receive antenna are independent with identical power delay profile.

The potential gain from applying space-time processing is strongly dependent on the spatial correlation coefficient. Considering that only the immediate surroundings of the antenna array impose the correlation between array elements [7], we model the correlation among receiver and transmitter array elements independently from one another. It is assumed that BS antennas are sufficiently separated, so that the transmitted signals are uncorrelated. In the following analysis, we focus on the spatial correlation between RX antennas at MS, defined as

$$\rho_{jk} = < \left|h_{ji}\right|^2, \left|h_{ki}\right|^2 > \ . \tag{3}$$

In numerical examples, we will use the spatial correlation derived from a truncated Gaussian power azimuth spectrum [4], which is a function of antenna spacing, mean angle-of-arrival and the angle spread. For an M-element uniform linear array with omnidirectional elements, we have

$$\mathbf{h}_i^{(l)} = \mathbf{v}(\varphi_l) \cdot \alpha_l e^{j\phi_l(\tau)} \quad , \tag{4}$$

where $\mathbf{v}(\varphi_l)$ represents the array response vector in terms of the azimuthual angle $\varphi_l$. In a Nakagami fading channel, the $\{\alpha_l\}_{l=0}^{L_p-1}$ are independent Nakagami distributed signal envelopes with pdf given in [14] as

$$p(\alpha_l) = \frac{2}{\Gamma(m_l)} \cdot \left(\frac{m_l}{\Omega_l}\right)^{m_l} \cdot \alpha_l^{2m_l-1} \cdot e^{-\frac{m_l}{\Omega_l}\alpha_l^2} \quad , \tag{5}$$

where $\Gamma(.)$ is the Gamma function, $\Omega_l = \overline{\alpha_l^2}$ is the average power on the *lth* path, the phase $\phi_l$ is uniformly distributed over the range $[0,2\pi)$, and $m_l \geq 1/2$ is the fading parameter, with the special case $m_l = 1$ corresponding to the Rayleigh distribution. Following [5], an exponential Multipath Intensity Profile (MIP) is assumed in the analysis, i.e. $\Omega_l = \Omega_0 e^{-l\delta}$, where $\Omega_0$ is the average signal strength corresponding to the first incoming path, and $\delta$ is the rate of average power decay, with $\delta = 0$ corresponding to constant MIP assumption.

## III. Bit Error Performance of 2D-RAKE Receiver in MIMO Systems

For the purpose of illustration, the simple dual transmit diversity space-time block code proposed by Alamouti [15] is adopted in the analysis. Alamouti [15] has shown that the maximum likelihood (ML) estimates of the transmitted data are identical to the ML estimates obtained in a system with a single transmit antenna and dual receive antennas. Therefore, the SINR at the output of 1D-RAKE receiver can be expressed as

$$\gamma = \frac{1}{2}\sum_{l=0}^{L-1}\left(\gamma_{1,l} + \gamma_{2,l}\right), \tag{6}$$

where the factor ½ is due to sharing of the transmitted signal power between two antennas.

Combining both transmit and receive diversity, and assuming perfect channel vector estimation and MRC combining, the instantaneous SINR at the output of 2D-RAKE receiver is given by

$$\gamma = \frac{1}{2}\sum_{l=0}^{L-1}\sum_{j=1}^{M_R}\left(\gamma_{j1,l} + \gamma_{j2,l}\right) \ , \tag{7}$$

where $\gamma_{j1,l}$ and $\gamma_{j2,l}$ are the SINRs of the *lth* path signal from the 1st and 2nd TX antenna to the jth receive antenna, respectively. Since the channels from different TX antennas to the receiver array are assumed to be independent and identically distributed, the characteristic functions of $\gamma_{j1,l}$ and $\gamma_{j2,l}$ have the same form as given in [4], i.e.

$$\Phi_l(t) = \left| I_{M_R} - it\frac{\overline{\gamma}_l}{m_l} \cdot \mathbf{R}_s^{(l)} \right|^{-m_l} \quad , \tag{8}$$

where $I_{M_R}$ is $M_R \times M_R$ dimension identity matix, $\mathbf{R}_s^{(l)}$ denotes the $M_R \times M_R$ dimension spatial correlation matrix, and $\overline{\gamma}_l = E_b/N_e \cdot \Omega_l$ is the average SINR per receive antenna contributed from the *lth* path. Note that $E_b$ is the energy per transmitted bit and $N_e$ is the equivalent power spectral density including AWGN noise and the total interference, which can be approximated as a spatially and temporally white Gaussian noise [5]. Since we assume that all resolvable paths fade independently, the characteristic function of $\gamma$ is simply

$$\Phi_\gamma(t) = \prod_{l=0}^{L-1}\left| I_{M_R} - it\frac{\overline{\gamma}_l}{2m_l} \cdot \mathbf{R}_s^{(l)} \right|^{-2m_l} = \prod_{l=0}^{L-1}\left| I_{M_R} - it\frac{\overline{\gamma}_0 \cdot e^{-l\delta}}{2m_l} \cdot \mathbf{R}_s^{(l)} \right|^{-2m_l} \tag{9}$$

The average bit error probability in the presence of fading is obtained by averaging the conditional error probability $P(e\,|\,\gamma) = Q(\sqrt{2\gamma})$ [16] over the pdf of $\gamma$, where *Q(x)* is the Gaussian Q-function. Using the alternative representation of *Q(x)* given in [2], the average BER can then be written as

$$P_b=\frac{1}{\pi}\int_0^{\infty}\int_0^{\pi/2}\exp\left(-\frac{\gamma}{\sin^2\vartheta}\right)d\vartheta\cdot p(\gamma)d\gamma$$

$$=\frac{1}{\pi}\int_0^{\pi/2}\Phi_\gamma(t)\Big|_{it=-1/\sin^2\vartheta}d\vartheta=\frac{1}{\pi}\int_0^{\pi/2}\prod_{l=0}^{L-1}\left|I_{M_R}+\frac{\overline{\gamma}_0\cdot e^{-l\delta}}{2m_l\cdot\sin^2\vartheta}\cdot\mathbf{R}_s^{(l)}\right|^{-2m_l}d\vartheta\ . \tag{10}$$

It is straightforward to show that the BER of a general $(M_T, M_R)$ MIMO system can be extended from (10) as

$$P_b=\frac{1}{\pi}\int_0^{\pi/2}\prod_{l=0}^{L-1}\left|I_{M_R}+\frac{\overline{\gamma}_0\cdot e^{-l\delta}}{M_T\cdot m_l\cdot\sin^2\vartheta}\cdot\mathbf{R}_s^{(l)}\right|^{-M_T\cdot m_l}d\vartheta\ . \tag{11}$$

It is necessary to point out that the derivation of (10) is based on the use of space-time block codes from orthogonal designs. However, [17] proved that for complex orthogonal designs, only $M_T=2$ Alamouti-code can provide the maximum possible transmission rate and full diversity. For $M_T>2$, the space-time block codes can give full diversity, but lose up to half of the theoretical bandwidth efficiency. Therefore, the generation from (10) to (11) is only valid for real constellation, or with the loss in bandwidth for complex constellation.

Since $\mathbf{R}_s^{(l)}$ is positive definite, it can be easily shown that (11) can be rewritten as

$$P_b=\frac{1}{\pi}\int_0^{\pi/2}\prod_{l=0}^{L-1}\prod_{i=1}^{M_R}\left(1+\frac{\overline{\gamma}_o\cdot e^{-l\delta}}{M_T\cdot m_l\cdot\sin^2\vartheta}\cdot\lambda_{l,i}\right)^{-M_T\cdot m_l}d\vartheta\ , \tag{12}$$

where $\{\lambda_{l,i}, l=0..L-1, i=1..M_R\}$ are the eigenvalues of $\mathbf{R}_s^{(l)}$.

## IV. Combined Coding and Diversity

FEC techniques such as convolutional coding are employed in mobile radio communication systems in order to protect the information against severe fading due to multipath propagation. The convolutional codes will be denoted by $(n,k,C_L)$, where $n$ is the number of encoded output bits per $k$ binary input information bits, and $C_L$ is the code constraint length. Following the usual transfer function bound [16] for maximum likelihood decoding over memoryless channels, the average bit error probability of a rate $r_c=k/n$ linear convolutional code may be bounded as

$$P_e\le\frac{1}{k}\sum_{d=d_{free}}^{\infty}\beta_d P_2(d)\ , \tag{13}$$

where $d_{free}$ is the free distance of the code, the $\{\beta_d\}$ are the coefficients in the expansion of the derivative of $T(D,N)$, the transfer function of the code, evaluated at $N$ = 1 [18]. $P_2(d)$ is the average pairwise error probability of selecting an incorrect path $\underline{x}'$ in the trellis that merges with the all-zero path $\underline{x}$ for the first time, and $\underline{x}'$ differs from $\underline{x}$ in $d$ positions $i_1, i_2, \cdots i_d$.

Since the coded bits in the two paths are identical except in the $d$ positions, the euclidean metric may be written as $\varepsilon(\underline{x},\underline{x}')=\sum_{n=1}^{d}\varepsilon(i_n)$, where $\varepsilon(i_n)$ denotes the energy of $i_n$ th bit. Thus, the conditional pairwise error probability for BPSK modulated signals is simply $Q\left(\sqrt{2\varepsilon(\underline{x},\underline{x}')/N_e}\right)$. For a (1, $M_R$) system in an L-path selective fading channel, we have

$$P_2(d)=E\left[Q\left(\sqrt{\frac{2E_b}{N_e}\sum_{n=1}^{d}\sum_{l=0}^{L-1}\sum_{j=0}^{M_R-1}\left|h_j^{(l)}(i_n)\right|^2}\right)\right]=E\left[Q\left(\sqrt{\frac{2E_b}{N_e}\sum_{n=1}^{d}\hbar_n^2}\right)\right] \tag{14}$$

where $\hbar_n^2=\sum_{l=0}^{L-1}\sum_{j=0}^{M_R-1}\left|h_j^{(l)}(i_n)\right|^2$, $E[\cdot]$ denotes the expectation taken with respect to the channel states $\hbar_n$. Since we assumed that ideal interleaving and deinterleaving make the $\hbar_n$ s i.i.d. random variables, the average over $\{\hbar_n\}$ can be computed as the product of averages. Consequently, the characteristic function of the overall SINR can be represented by the product of the characteristic function of each codesymbol's SINR. Following the similar derivation in Section III, the average pairwise error probability for a $(M_T, M_R)$ system can be written as

$$P_2(d)=\frac{1}{\pi}\int_0^{\pi/2}\prod_{l=0}^{L-1}\left|I_{M_R}+\frac{\overline{\gamma}_0\cdot e^{-l\delta}}{M_T\cdot m_l\cdot\sin^2\vartheta}\cdot\mathbf{R}_s^{(l)}\right|^{-M_T\cdot m_l\cdot d}d\vartheta\ . \tag{15}$$

## V. Numerical Examples

In this section, we present selected numerical results to determine the effects of different code rates and constraint lengths on the coded DS-CDMA system gain in correlated Nakagami-m fading channels, and to illustrate the benefits of combined coding and diversity techniques. It is assumed that the total transmit power is fixed, regardless of the number of transmit antenna elements, and uniform power is allocated to each transmit antenna. In order to emphasize the space diversity gain due to combining only, BER versus SNR per bit will be plotted, under the condition that the total average received SNR is fixed regardless of the value of $M_R$. In addition, constant MIP and identical path fading parameters are employed in the following examples.

Table I lists the parameters of several low rate maximum free distance codes [18]. Since the upper bound computed by using the first six terms of the series expansion of the bounds in (13) very accurately approximates the simulation result as demonstrated in [13], we will only sum the pairwise error probabilities for the six shortest Hamming distances as the coded BER upper bound. For a fair comparison between the coded and uncoded system, the total energy used for transmitting $k$ information bits is assumed to be equal to that used for transmitting $n$ coded bits. That is to say, if $E_b$ is the energy of the information bit.

Table I Maximum free distance convolutional codes

| ***n*** | ***k*** | $C_L$ | $d_{free}$ |
|---|---|---|---|
| 2 | 1 | 7 | 10 |
| 2 | 1 | 4 | 6 |
| 3 | 1 | 4 | 10 |
| 4 | 1 | 4 | 13 |

Fig. 1 shows the uncoded BER and coded BER, using a (2, 1, 7) code, of a (2, 2) system in different Nakagami fading channels with two resolvable paths. The performance improvement achieved by coding is shown to be very impressive. Since the free distance $d_{free} = 10$ results in a coding gain which has the similar effect as introducing $10^{th}$ order diversity, the coded system has an effective order of diversity equal to $\left(M_T \times m \times M_R \times L \times d_{free}\right)$. This also results in more significant coding gain in severe fading channels.

Next in Fig. 2, we consider two RX antennas separated by 1/4 wavelength, with an angular spread of $\sigma = 60^o$, (which gives a spatial correlation of approximately 0.53,) operating in the flat fading or selective fading channel with $m = 0.75$. The comparison between applying the (2,1,7) and the (2,1,4) codes indicates that by increasing the convolutional code constraint length, the performance becomes better due to the larger free distance. The achievable coding gain is smaller in the case of L=4 multipath components compared with signaling over the frequency nonselective channel.

Three codes with the same constraint length $C_L = 4$ and different coding rates applied in (1,1) and (1,2) systems are compared in Fig. 3. Independent antenna branches in a two-path, $m = 0.75$ fading channel are assumed. It is not surprising that for a fixed number of RX antennas, the performance is improved with decreasing code rate. On the other hand, the more powerful the code, the smaller the gain realized by space-path diversity. For example, at BER= $10^{-5}$, dual RX antennas provide 1.4dB and 0.7dB diversity gain when applying $r_c = 1/2,\ 1/4$ codes, respectively. This fact leads to the conclusion that with powerful codes, the main purpose of the 2D-RAKE receiver is to collect the power distributed in various propagation paths, not to improve the performance by introducing additional diversity.

Finally, Fig. 4 illustrates the BER performance of different coding and diversity schemes versus the average received SINR of a 1D-RAKE receiver, i.e. $E_b/N_e \cdot \sum_{l=0}^{L-1}\Omega_l$. Specifically, we consider the single TX and RX antenna system employing codes (2,1,7) and (2,1,4), and a (1, 2) system with or without coding, in $m = 1.25$ fading channel. To illustrate the trade-off between space-path diversity gain and coding gain, we allow the total received power to be proportional to the number of receive antennas. For dual RX diversity, two angular spreads $\sigma = 60^o$ and $25^o$ are assumed, giving spatial correlations equal to 0.53 and 0.82, respectively. It is observed that the performance with antenna diversity alone is better than the coded system with a single antenna up to a certain SNR threshold, and above which the performance is degraded relative to the coded single antenna case. The threshold depends on various factors, such as fading severity, the coding scheme, the spatial correlation and the number of TX/RX antennas. Thus, adding diversity, especially at a low SNR level, can effectively reduce the number of bit errors at the convolutional decoder input, thus the advantages of coding can be fully exploited.

## VI. CONCLUSION

We derived the BER of a general MIMO system using maximal ratio combining and open loop transmit diversity in correlated Nakagami fading channels. In addition, we incorporated the exact pairwise error probability into the transfer function bound, so that the well known performance advantages of convolutional codes in Rayleigh fading are extended to the general MIMO system in frequency selective Nakagami fading. The effective order of diversity (asymptotic slope) is equal to the product of four fundamental design parameters (the number of both the TX and RX antennas, the number of RAKE fingers and the free distance of the convolutional code) and also the fading channel parameter. Our results are sufficiently general and apply to cases where the instantaneous SNR's of the resolvable multipaths come from different Nakagami families, as well as dissimilar average SNR's from multipath components.

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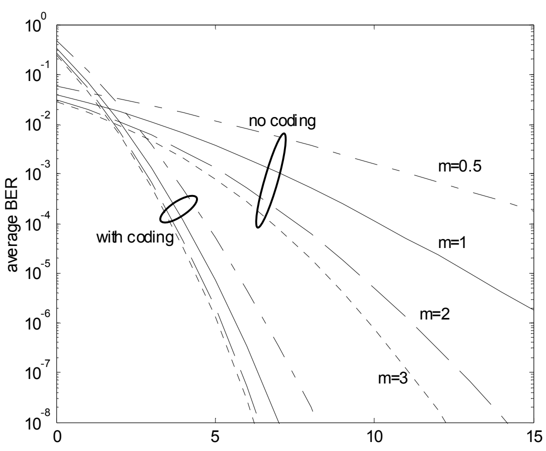


Fig. 1 Uncoded and coded BER of a (2, 2) system in different Nakagami fading channels with two resolvable paths.

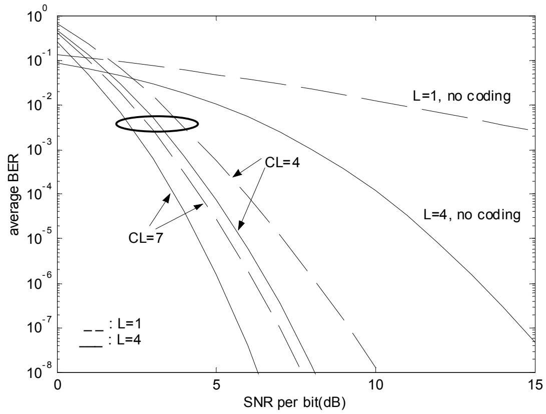


Fig. 2 Performance in flat fading (L=1) or selective fading (L=4) channel with two RX antennas separated by 1/4 wavelength, angular spread $\sigma = 60^o$.

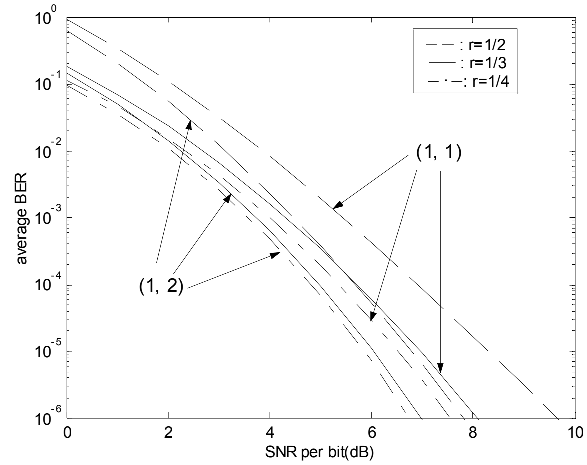


Fig. 3 Performance comparison employing three convolutional codes with the same constraint length and different coding rates: (2,1,4), (3,1,4) and (4,1,4).

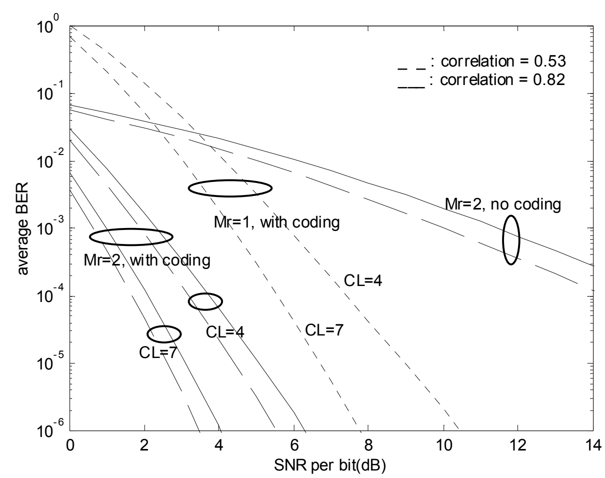


Fig. 4 Performance comparison of different coding and diversity schemes in the $m = 1.25$ fading channel.