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PERFORMANCE OF SPACE TIME CODES WITH TRANSMIT ANTENNA SELECTION OVER FREQUENCY SELECTIVE FADING CHANNELS IN THE PRESENCE OF CHANNEL ESTIMATION ERRORS

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ABSTRACT

This paper presents a performance analysis of space-time coded systems with transmit antenna selection over frequency selective fading channels when the erroneously estimated fading coefficients are available only at the receiver. An upper bound on the pairwise error probability is derived and numerical examples are presented. The analysis does not assume any specific coding or channel estimation algorithm. With imperfect channel state information (CSI) used both at the antenna selection and the space-time decoding processes, the achievable diversity order does not decrease compared to perfect CSI scenario.

1. INTRODUCTION

Antenna selection [1] has become a popular technique which requires only a few RF chains switched to selected antennas. This can be highly effective in reducing the cost and the complexity of space time coded (STC) [2, 3, 4] systems especially over frequency selective fading channels in high rate communications.

In the literature, there has been considerable research on antenna selection generally about fast selection and error performance. Although most works consider the selection only at the receiver [5, 6], STC systems with transmit antenna selection [7, 8] and joint transmit/receive antenna selection [9] have also been studied recently. In general, it has been shown that full space diversity can still be achieved with antenna selection with the assumption of perfect channel state information (CSI) available at the receiver. Furthermore, antenna selection with imperfect CSI is addressed in [10] where specific space-time block coding (STBC) systems employing single antenna selection at the receiver are studied. In high rate wireless transmission systems, frequency selective channel model is more common than flat fading. However, there are only a few papers about the performance of antenna selection considering frequency-selectivity [11].

In this paper, we present the performance of general STCs with transmit antenna selection based on the largest received powers over frequency selective fading channels. We assume that an imperfect channel estimation algorithm provides erroneous fading coefficients which are used at the selection and the space-time decoding processes. By deriving an upper bound on the pairwise error probability of STCs and performing computer simulations we show that the diversity order of STC systems with transmit antenna selection is not degraded compared to perfect CSI assumption.

The rest of the paper is organized as follows: Section 2 describes the system model. The pairwise error probability bound for transmit selection when the receiver has imperfect channel estimates is derived in Section 3. Numerical examples are provided in Section 4 followed by the conclusions in Section 5.

2. SYSTEM DESCRIPTION

In this section, we describe the system model for general STC systems in the presence of channel estimation errors. Figure 1 shows a STC system with transmit antenna selection. The channel is modeled as a quasi-static MIMO Rayleigh frequency selective fading channel where the different sub-channels fade independently. The channel estimation uses the demodulated signals from the N receive antennas to estimate the fading coefficients which are used in space time decoding and in the antenna selection based on the largest received powers. We assume that there are M transmit antennas available, however, only L_T of them are selected and used in each frame. The indices of the L_T transmit antennas are fedback periodically which only requires at most M bits, thus, it does not slow down the transmission rate significantly. At the transmitter, the information sequence is encoded by a space-time encoder, then, multiplexed into L_T data streams which are modulated and transmitted through the selected antennas simultaneously.

For a general multiple antenna system with L_T transmit and N receive antennas, and D intersymbol interference (ISI) taps, the received signal at antenna n at time k can be written as

$$y_n(k) = \sqrt{\frac{\rho}{L_T D}} \sum_{d=0}^{D-1} \sum_{m=1}^{L_T} h^d_{m,n} s_m(k-d) + w_n(k), \qquad (1)$$

where $h^d_{m,n}$ is the fading coefficient between transmit antenna m and receive antenna n, corresponding to the ISI tap d. $s_m(k)$ is the transmitted symbol from antenna m and $w_n(k)$ is the noise term at antenna n at time k, $k = 1, \cdots, K$, where K is the frame length. Both fading channel coefficients, and noise terms are modeled as zero mean complex Gaussian random variables with variance $1/2$ per dimension. The fading coefficients are spatially independent, but they are assumed to be constant over an entire frame (i.e., quasi-static fading) and we assume uniform multipath delay profile. Signal constellation at each transmit antenna is normalized so that the

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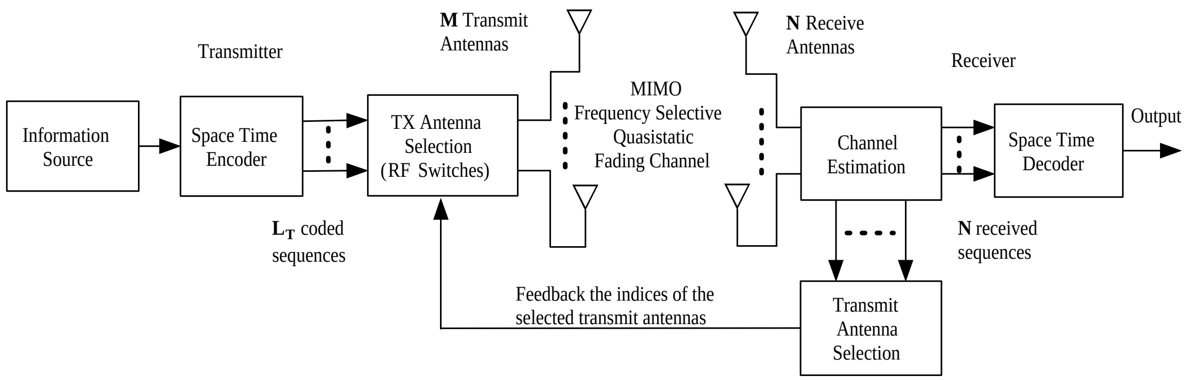

Figure 1: Block diagram of space-time coded multiple antenna system with transmit antenna selection.

average power of the transmitted signals is unity, and ρ is interpreted as the average signal to noise ratio (SNR) at each receive antenna. We assume that the receiver obtains the CSI via some training symbols, however, the transmitter does not have access to this, and thus it evenly splits its power across L_T transmit antennas. The received signals can be stacked in a matrix form as

$$\mathbf{Y} = \sqrt{\frac{\rho}{L_T D}}\mathbf{HS} + \mathbf{W}, \tag{2}$$

where the $N \times (K+D-1)$ received signal matrix is

$$\mathbf{Y} = \begin{pmatrix} y_1(1) & \dots & y_1(K+D-1) \\ \vdots & \ddots & \vdots \\ y_N(1) & \dots & y_N(K+D-1) \end{pmatrix}, \tag{3}$$

the $N \times L_T D$ channel coefficient matrix is

$$\mathbf{H} = \begin{pmatrix} h_{1,1}^0 & \dots & h_{1,1}^{D-1} & \dots & h_{L_T,1}^0 & \dots & h_{L_T,1}^{D-1} \\ \vdots & \vdots & \vdots & \ddots & \vdots & \vdots & \vdots \\ h_{1,N}^0 & \dots & h_{1,N}^{D-1} & \dots & h_{L_T,N}^0 & \dots & h_{L_T,N}^{D-1} \end{pmatrix}, \tag{4}$$

the $L_T D \times (K+D-1)$ codeword matrix is

$$\mathbf{S} = \begin{pmatrix} s_1(1) & \dots & s_1(K) & 0 & \dots & 0 \\ 0 & s_1(1) & \dots & s_1(K) & \dots & 0 \\ \vdots & \vdots & \vdots & \ddots & \vdots & \vdots \\ 0 & \dots & 0 & s_1(1) & \dots & s_1(K) \\ \vdots & \vdots & \vdots & \vdots & \vdots & \vdots \\ \vdots & \vdots & \vdots & \vdots & \vdots & \vdots \\ s_{L_T}(1) & \dots & s_{L_T}(K) & 0 & \dots & 0 \\ 0 & s_{L_T}(1) & \dots & s_{L_T}(K) & \dots & 0 \\ \vdots & \vdots & \vdots & \ddots & \vdots & \vdots \\ 0 & \dots & 0 & s_{L_T}(1) & \dots & s_{L_T}(K) \end{pmatrix}, \tag{5}$$

and the $N \times (K+D-1)$ noise matrix is

$$\mathbf{W} = \begin{pmatrix} w_1(1) & \dots & w_1(K+D-1) \\ \vdots & \ddots & \vdots \\ w_N(1) & \dots & w_N(K+D-1) \end{pmatrix}. \tag{6}$$

For any given $\mathbf{H}$, the PEP of erroneously receiving $\widehat{\mathbf{S}}$, when $\mathbf{S}$ is transmitted, is given by

$$P(\mathbf{S} \to \widehat{\mathbf{S}}|\mathbf{H}) = \frac{1}{2} erfc\left(\sqrt{\frac{\rho}{4DL_T}}\|\mathbf{HB}\|\right), \tag{7}$$

which can be upper bounded as

$$P(\mathbf{S} \to \widehat{\mathbf{S}}|\mathbf{H}) \le \exp\left(-\frac{\rho}{4DL_T}\|\mathbf{HB}\|^2\right), \tag{8}$$

where $\mathbf{B} = \mathbf{S} - \widehat{\mathbf{S}}$ is the codeword difference matrix. $\|.\|^2$ represents the Frobenius norm (i.e., the sum of magnitude squares of all entries).

3. TRANSMIT ANTENNA SELECTION IN THE PRESENCE OF CHANNEL ESTIMATION ERRORS

In this section, we investigate the performance, especially diversity order, of STC with transmit antenna selection over quasi-static frequency selective fading channels. The receiver obtains imperfect channel estimates which are used in antenna selection based on received powers.

In practical receivers, estimated channel coefficients can be written as follows [12],

$$\hat{h}_{m,n}^d = h_{m,n}^d + \varepsilon_{m,n}^d, \tag{9}$$

where $\varepsilon_{m,n}^d$ is a complex Gaussian random variable representing the channel estimation error independent of $h_{m,n}^d$, having zero mean and variance σ_e^2. $\hat{h}_{m,n}^d$ is a complex Gaussian random variable with zero mean, variance σ^2 per dimension and dependent on $h_{m,n}^d$ with the following correlation coefficient,

$$\mu = \frac{1}{\sqrt{1+\sigma_e^2}}, \tag{10}$$

where, σ_e^2 can be estimated from the SNR, the number of pilots, and the method of estimation. In the presence of channel estimation errors, as in [12], when $\mathbf{S}$ is transmitted, the conditional mean of the received signal can be written as

$$E\{y_n(k)|\hat{h}_{m,n}^d, s_m(k)\} = \frac{\mu}{\sqrt{2}\sigma}\sqrt{\frac{\rho}{L_T D}}\sum_{m=1}^{L_T}\sum_{d=0}^{D-1}\hat{h}_{m,n}^d s_m(k-d),$$

and the conditional variance is as follows

$$Var\{y_n(k)|\hat{h}^d_{m,n}, s_m(k)\} = 1 + (1-|\mu|^2)\frac{\rho}{L_T D}\sum_{d=0}^{D-1}\sum_{m=1}^{L_T}|s_m(k-d)|^2.$$

We note that the Euclidean distance term can be written as

$$d^2(\mathbf{S},\widehat{\mathbf{S}}) = \sum_{n=1}^{N}\sum_{k=1}^{K}\left|\sum_{d=0}^{D-1}\sum_{m=1}^{L_T}\frac{\hat{h}^d_{m,n}}{\sqrt{2}\sigma}s_m(k-d)\right|^2 = \frac{1}{2\sigma^2}\|\hat{\mathbf{H}}\mathbf{B}\|^2,$$

where the $N \times L_T D$ matrix $\hat{\mathbf{H}}$ contains the estimated channel coefficients, $\hat{h}^d_{m,n}$. Then, the PEP bound conditioned on $\hat{\mathbf{H}}$ can be obtained as,

$$P(\mathbf{S}\to\widehat{\mathbf{S}}|\hat{\mathbf{H}}) \le \exp\left(-\hat{\rho}\frac{1}{2\sigma^2}\|\hat{\mathbf{H}}\mathbf{B}\|^2\right). \qquad (11)$$

where the modified SNR term is defined as follows

$$\hat{\rho} = \frac{\mu^2\frac{\rho}{L_T D}}{4+4L_T D(1-|\mu|^2)\frac{\rho}{L_T D}}.$$

The unconditional PEP upper bound can be obtained by averaging the above conditional PEP using the statistics of the selected channel coefficients.

The derivations for the selection of arbitrary number of antennas are quite lengthy due to frequency-selectivity. Therefore, for the clarity of the analysis, we now study the selection of single transmit antenna ($L_T = 1$) which provides enough insight for the general case. First, we start with the statistics of $\hat{\mathbf{H}}$ matrix of size $N \times D$ having the largest norm selected from the complete channel matrix $\tilde{\mathbf{H}}$ of size $N \times MD$ containing all estimated fading coefficients between NM antenna pairs and for all D multipaths. Similar to [6], by using Gaussian and Chi-square statistics, the joint probability density function (pdf) $f(\hat{\mathbf{H}})$ of $\hat{\mathbf{H}}$ can be written as

$$f(\hat{\mathbf{H}}) = M\left(1-e^{-\frac{\|\hat{\mathbf{H}}\|^2}{2\sigma^2}}\sum_{n=0}^{ND-1}\frac{\left(\frac{\|\hat{\mathbf{H}}\|^2}{2\sigma^2}\right)^n}{n!}\right)^{M-1}\frac{1}{(\pi 2\sigma^2)^{ND}}e^{-\frac{\|\hat{\mathbf{H}}\|^2}{2\sigma^2}}.$$

Now, we take the average of the PEP bound in the expression (11) over all possible selected channel coefficients,

$$P(\mathbf{S}\to\widehat{\mathbf{S}}) \le \int_{C^{ND}}\exp\left(-\frac{\hat{\rho}}{2\sigma^2}\|\hat{\mathbf{H}}\mathbf{B}\|^2\right)f(\hat{\mathbf{H}})d\hat{\mathbf{H}},$$

where C^{ND} is the ND dimensional complex space. Then, using the following simplification [6]

$$1-e^{-x}\sum_{n=0}^{N-1}\frac{x^n}{n!} \le \frac{x^N}{N!} \text{ for } x>0, \qquad (12)$$

the upper bound on the PEP becomes,

$$\begin{aligned} P(\mathbf{S}\to\widehat{\mathbf{S}}) \quad \le \quad & \frac{M}{(\pi 2\sigma^2)^{ND}}\left(\frac{1}{ND!}\right)^{M-1}\int_{C^{ND}}e^{\left(-\frac{\hat{\rho}}{2\sigma^2}\|\hat{\mathbf{H}}\mathbf{B}\|^2\right)} \\ & \left(\frac{\|\hat{\mathbf{H}}\|^2}{2\sigma^2}\right)^{ND(M-1)}e^{-\frac{\|\hat{\mathbf{H}}\|^2}{2\sigma^2}}d\hat{\mathbf{H}}. \end{aligned}$$

where we can utilize eigenvalue decomposition of $\mathbf{B}\mathbf{B}^* = \mathbf{U}\mathbf{\Lambda}\mathbf{U}^*$ ($\mathbf{U}$ a unitary matrix). Since we use only a single transmit antenna, the codeword matrix $\mathbf{S}$ is $D\times K$, and therefore, $\mathbf{B}\mathbf{B}^*$ is $D\times D$ where $\mathbf{\Lambda}$ is a diagonal matrix with eigenvalues $\lambda_1,\ldots,\lambda_D$. Then, we note that

$$\|\hat{\mathbf{H}}\mathbf{B}\|^2 = trace\left((\hat{\mathbf{H}}\mathbf{U})\mathbf{\Lambda}(\hat{\mathbf{H}}\mathbf{U})^*\right) = \sum_{i=1}^{D}\lambda_i\|\mathbf{c}_i\|^2, \qquad (13)$$

where $\mathbf{c}_i$ is the i^{th} column of $\tilde{\mathbf{H}}\mathbf{U}$, and

$$\begin{aligned} \sum_{i=1}^{D}\|\mathbf{c}_i\|^2 &= trace\left((\hat{\mathbf{H}}\mathbf{U})(\hat{\mathbf{H}}\mathbf{U})^*\right) \\ &= trace\left(\hat{\mathbf{H}}\mathbf{U}\mathbf{U}^*\hat{\mathbf{H}}^*\right) \\ &= trace\left(\hat{\mathbf{H}}\hat{\mathbf{H}}^*\right) \\ &= \|\hat{\mathbf{H}}\|^2. \end{aligned} \qquad (14)$$

At this point, let us assume that we have a full-rank space-time code which means that all the eigenvalues of the matrix $\mathbf{B}\mathbf{B}^*$ are positive (i.e., nonzero). We denote the minimum of $\lambda_1,\ldots,\lambda_D$ by $\hat{\lambda}$ and note that

$$\sum_{i=1}^{D}\lambda_i\|\mathbf{c_i}\|^2 \ge \hat{\lambda}\|\hat{\mathbf{H}}\|^2. \qquad (15)$$

Hence, we can obtain the following upper bound to be used in simplification of the PEP upper bound,

$$e^{\left(-\frac{\hat{\rho}}{2\sigma^2}\|\hat{\mathbf{H}}\mathbf{B}\|^2\right)} \le e^{\left(-\frac{\rho}{4D}\hat{\lambda}\|\hat{\mathbf{H}}\|^2\right)}.$$

We also note that

$$\|\hat{\mathbf{H}}\|^2 = \sum_{n=1}^{N}\sum_{d=0}^{D-1}|\hat{h}_{n,d}|^2$$

where $\hat{h}_{n,d}$ is the estimated channel coefficient at row n and column d of $\hat{H}$. First, by making the change of variables, $\frac{h_{n,d}}{\sqrt{2\sigma^2}} = \alpha_{n,d}e^{\theta_{n,d}}$ ($dh_{n,d} = \sqrt{2\sigma^2}\alpha_{n,d}d\alpha_{n,d}d\theta_{n,d}$), then defining $u_{n,d} = \alpha^2_{n,d}$ ($du_{n,d} = 2\alpha_{n,d}d\alpha_{n,d}$), and using $\int_0^{2\pi}d\theta_{n,d} = 2\pi$, we obtain

$$\begin{aligned} P(\mathbf{S}\to\widehat{\mathbf{S}}) \quad \le \quad & \psi\int_0^\infty\cdots\int_0^\infty e^{\left(-(\hat{\rho}\hat{\lambda}+1)\sum_{n=1}^{N}\sum_{d=0}^{D-1}u_{n,d}\right)} \\ & \left(\frac{\sum_{n=1}^{N}\sum_{d=0}^{D-1}u_{n,d}}{ND!}\right)^{ND(M-1)}\prod_{n=1}^{N}\prod_{d=0}^{D-1}du_{n,d}. \end{aligned}$$

where $\psi = M\frac{1}{(2\sigma^2)^{(ND)/2}}\left(\frac{1}{ND!}\right)^{M-1}$. In order to simplify further, we can write the double summation as a single summation as follows

$$\sum_{n=1}^{N}\sum_{d=0}^{D-1} u_{n,d} = \sum_{r=1}^{ND} v_r, \tag{16}$$

where $v_r = u_{n,d}$ with the index $r = dN + n$. This allows us to use the following expansion

$$\left(\sum_{r=1}^{ND} v_r\right)^{ND(M-1)} = \sum_{r_1=1}^{ND}\cdots\sum_{r_{ND(M-1)}=1}^{ND} v_{r_1}\cdots v_{r_{NM-NL_T}}, \tag{17}$$

where the terms v_r with l_r multiplicities in $v_{r_1}\cdots v_{r_{NM-NL_T}}$ can be written as $\prod_{r=1}^{ND}(v_r)^{l_r}$ such that

$$\sum_{r=1}^{ND} l_r = ND(M-1). \tag{18}$$

Then we can write the PEP bound as,

$$P(\mathbf{S}\to\widehat{\mathbf{S}}) \leq \psi\int_0^{\infty}\cdots\int_0^{\infty} e^{-\sum_{r=1}^{ND}(\hat{\rho}\hat{\lambda}+1)v_r} \sum_{r_1=1}^{ND}\cdots\sum_{r_{ND(M-1)}=1}^{ND}(v_r)^{l_r}dv_1\cdots dv_{ND}.$$

Changing the order of summation and integration and using

$$\int_0^{\infty} x^m e^{-ax}dx = \frac{m!}{a^{m+1}},$$

results in

$$P(\mathbf{S}\to\widehat{\mathbf{S}}) \leq \psi\sum_{r_1=1}^{ND}\cdots\sum_{r_{ND(M-1)}=1}^{ND}\prod_{r=1}^{ND}\frac{l_r!}{\left(\hat{\rho}\hat{\lambda}+1\right)^{(l_r+1)}}. \tag{19}$$

Finally, considering high SNRs and expression (18), we obtain

$$P(\mathbf{S}\to\hat{\mathbf{S}}) \leq \frac{\psi}{\hat{\lambda}^{MND}}\left(\sum_{r_1=1}^{ND}\cdots\sum_{r_{ND(M-1)}=1}^{ND}\prod_{r=1}^{ND} l_r!\right)\hat{\rho}^{-MND}. \tag{20}$$

This result shows that a diversity order of MND (i.e., full diversity available in the system) can be achieved when only one transmit antenna is selected based on the instantaneous SNR at the receiver. Although the diversity advantage is the same as the full-complexity system, the coding gain decreases with antenna selection. Since selecting only one transmit antenna achieves full spatial diversity, we expect the same diversity order to be achieved when more than one transmit antenna is selected as well. The coding gain depends on the number of antennas and the eigenvalues of the square of the codeword difference matrix, $\mathbf{BB}^*$. Obviously, the coding gain with antenna selection will be lower than that of full-complexity system.

Although the PEP analysis for using more than one antennas in actual transmission ($L_T > 1$) is not shown due to space limitations, it is obvious that full diversity can be achieved and an upper PEP bound at high SNRs can be simply written as

$$P(\mathbf{S}\to\widehat{\mathbf{S}}) \leq \kappa\tilde{\rho}^{-MND}, \tag{21}$$

where κ is independent of SNR and a function of M, N, D, L_T, μ and eigenvalues of $\mathbf{B}$.

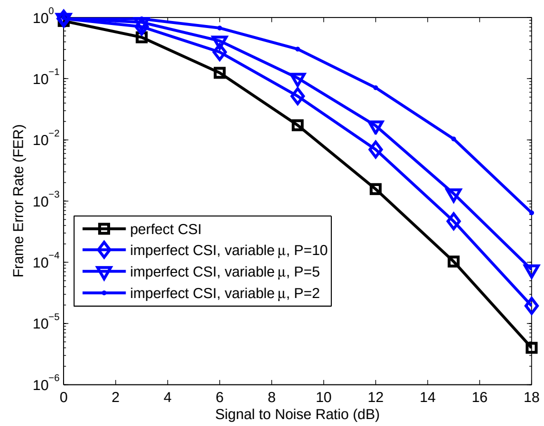

Figure 2: FER plots for transmit antenna selection, $M = 3, N = 1, L_T = 2, D = 2$, μ depends on number of pilots and SNR.

4. SIMULATION RESULTS

In this section, error rates of STC systems with joint transmit and receive antenna selection using imperfect CSI are illustrated. We note that in the presence of channel estimation errors, the decoding metric should be as described in [12] which is slightly different than the metric for perfect CSI scenario. We have observed that, the performance results with both metric are almost the same.

The frame error rate (FER) plots of STC based on $(5,7)_8$ convolutional coding [4] over frequency selective fading channels with transmit antenna selection $M = 3, N = 1, L_T = 2, D = 2$ are depicted in Figure 2 when the correlation coefficient μ is modeled as a variable. As in practical receivers, the channel estimation error in these simulations assumed to decrease with increasing SNR, and the constant P which is related to the number of pilots and the method of channel estimation. We observe that the full diversity order 6 is achieved with perfect CSI and it remains the same with imperfect channel estimation. When P decreases, FER increases although the diversity does not change.

The FER plots of STC based on $(5,7)_8$ convolutional coding with transmit antenna selection $M = 3, N = 1, L_T = 2, D = 3$ with fixed correlation coefficient μ are depicted in Figure 3. This figure STC is rank-deficient due to increased number of ISI taps [4] and it cannot achieve full diversity $MND = 9$ with perfect and imperfect CSI. This figure illustrates that at low to mid SNRs, the performance degradation due to channel estimation errors is insignificant, however, at high SNRs the error floors occur. When the correlation μ is larger than 0.9995 ($\sigma_e^2 < 0.001$), the FER is almost the same as the FER with perfect CSI scenario. When μ is smaller than

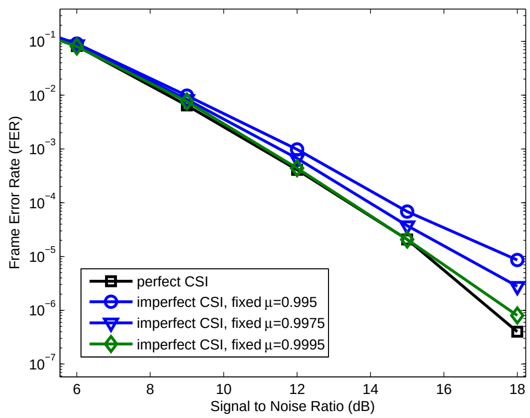

Figure 3: FER plots for transmit antenna selection, $M = 3, N = 1, L_T = 2, D = 3$, μ has a fixed value.

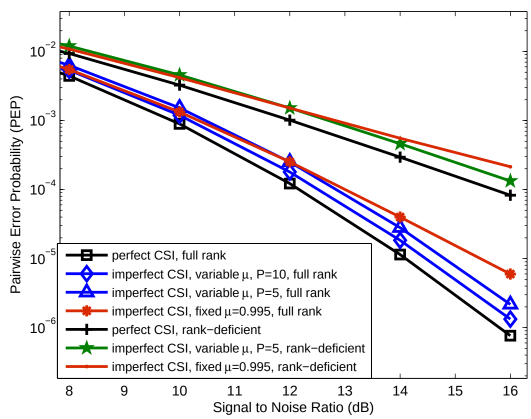

Figure 4: PEP plots for transmit antenna selection, $M = 4, N = 1, L_T = 2, D = 2$, considering fixed and variable μ.

0.9975, the degradation becomes significant, thus, in practical systems, there can be restrictions on the mean square error of channel estimators.

The exact PEP plots of the STCs based on generalized and standard delay diversity scheme [3] with transmit antenna selection $M = 4, L_T = 2, N = 1, D = 2$ are depicted in Figure 4. After extensive simulations for several cases, we verified the theoretical results, compared to perfect CSI scenario, we observe that the diversity order is preserved with imperfect CSI. The performance degradation for both full rank and rank-deficient codes under channel estimation errors (fixed and variable μ cases) are similar.

5. CONCLUSIONS

In this paper, the effect of imperfect channel estimates on the performance of space time coded systems with transmit antenna selection over frequency selective fading channel is presented. Only the receiver is assumed to have the imperfect CSI and the antenna selection is based on maximum estimated received powers. The pairwise error probability analysis and the numerical examples have shown that the diversity order achievable with perfect CSI is not reduced when imperfect channel estimates are used in antenna selection and space time decoding. The performance degradation caused by channel estimation can be seen as reduction in SNR.

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