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Encoding Performance of Orthogonal Space-Time Coded Continuous Phase Modulation System

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Abstract
The orthogonal space-time coded continuous phase modulation (OST-CPM) system shows attractive performance over fading MIMO channels. In this paper, the Chernoff bound on pair-wise error probability (PWEP) is studied for two transmit antennas over spatially correlated quasi-static Rayleigh-fading channel. The maximum likelihood sequence detection (MLSD) algorithm is applied to the OST-CPM system. Approximate bound for high signal-to-noise ratio (SNR) is derived to evaluate the encoding performance in correlated channel. The effects of correlation coefficient matrices on the coding performance are simulated. Both analytical and simulation results show that the coding performance of this system decreases as the fading coefficients between the antennas increases. And the penalty on the coding performance increases a lot in fully correlated channel.

Keywords: *OST-CPM, correlated channel, Chernoff bound, pair-wise error probability*

1. Introduction

Space-time coding has shown considerable promise for reliable transmission over wireless fading channels by efficiently employing diversity [1] -[5] . Almost all existing space-time code designs consider linear modulation scheme. However the continuous phase modulation (CPM) has the characteristics of constant envelope and phase continuity. Therefore CPM has become an attractive scheme for the data transmission over both bandwidth and power limited links such as mobile satellite communications [6] . Combining CPM with space-time coding (STC-CPM) can provide better performance in wireless communications [7] -[11] . Wang et al.[8] extended Alamouti's orthogonal encoding criterion to CPM signals and designed orthogonal space-time coded CPM (OST-CPM) systems with n_T=2, where different CPM schemes are applied for the two transmit antennas. The optimum detection method for OST-CPM systems is maximum likelihood sequence detection (MLSD) algorithm. Previous work on OST-CPM systems has been restricted ot the ideal case of independent and identically distributed (i.i.d.) channels, i.e., uncorrelated fading channels. However insufficient antenna spacing and lack of scattering cause the individual antennas to be correlated for a real-world channel model [12] .

Chernoff bounds on pair-wise error probabilities (PWEP) have been used to design space-time codes and analyze the performance of multiple-input-multiple-output (MIMO) wireless communication systems. The effect of spatial correlation on the PWEP of the space-time code is studied in slow Rayleigh fading MIMO channels [13] [14] . Most of the work reseached on OST-CPM is on encoding and decoding design, and no similar work has been done with the impact of spatial correlation.

In this paper, we derive Chernoff bound expressions for MLSD algorithm of the OST-CPM system in a correlated quasi-static Rayleigh-fading channel. Approximate Chernoff bound and achievable diversity gain are derived for high scattering signal-to-noise ratio (SNR). We consider spatically correlated fading on the transmit or receive side or both. The frame error rate (FER) performance of OST-2CPM systems in fully and partially correlated and i.i.d. channels are simulated and compared with each other.

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2. System Model

In this paper, a wireless communication system with n_T transmit and n_R receive antennas is considered. For simplification we just focus on the baseband equivalent block diagram of the OST-CPM system, which is shown in

Figure 1.

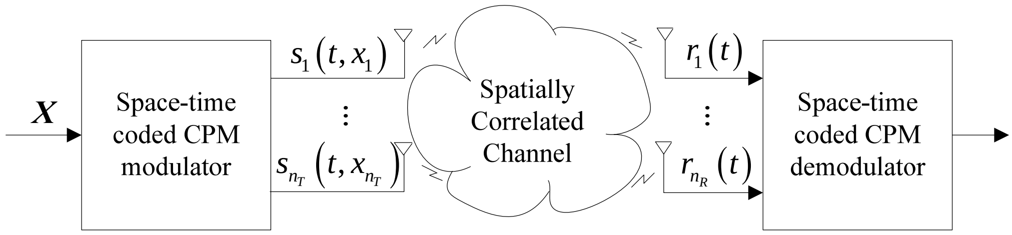

Figure 1. Simplified base-band equivalent block diagram of OST-CPM system

We define an n_T×L space-time codeword matrix $\boldsymbol{X}$ as the input of space-time coded modulator, obtained by arranging the transmitted sequence in an array as

$$\boldsymbol{X} = \left[\boldsymbol{x}_1, \boldsymbol{x}_2, \cdots, \boldsymbol{x}_{n_T}\right]^T = \begin{bmatrix} x_{1,1} & x_{1,2} & \cdots & x_{1,L} \\ x_{2,1} & x_{2,2} & \cdots & x_{2,L} \\ \vdots & \vdots & \ddots & \vdots \\ x_{n_T,1} & x_{n_T,2} & \cdots & x_{n_T,L} \end{bmatrix} \tag{1}$$

where the i-th row $\boldsymbol{x}_i$ =[$x_{i,1}$, $x_{i,2}$, …, $x_{i,L}$] is the data sequence transmitted from the i-th transmit antenna. Then the OST-CPM signals are simultaneously transmitted from the n_T transmit antennas.

The transmitted signal at the i-th antenna can be represented as

$$s_i(t, \boldsymbol{x}_i) = \sqrt{\frac{E_s}{T}} \exp\left\{ j\phi(t, \boldsymbol{x}_i) \right\}, i = 1, 2, \cdots, n_T \tag{2}$$

$$\phi(t, \boldsymbol{x}_i) = 2\pi h \sum_{n=0}^{k} x_{i,n} q(t - nT), kT \le t < (k+1)T \tag{3}$$

where E_s is the symbol energy, T is the symbol period, ϕ (t, x_i) is the carried phase information, h is the modulation index. $q(t) = \int_{-\infty}^{t} g(\tau) d\tau$ is the phase smoothing response function. $g(t)$ is the pulse shaping function, which is nonzero only at the limited time period $0 \le t \le LT$ (L is the modulation memory). We take the OST-CPM system with two transmit antennas for example. The two CPM-modulated signals are transmitted by the two transmit antennas simultaneously. Due to the orthogonal coding design for the CPM signals, the rows of the matrix

$$\begin{bmatrix} s_1(t, \boldsymbol{X}) & s_1(t+T, \boldsymbol{X}) \\ s_2(t, \boldsymbol{X}) & s_2(t+T, \boldsymbol{X}) \end{bmatrix} \tag{4}$$

are orthogonal for each t for the MSLD algorithm to be studied in the next section [8]

The MIMO channel with n_T transmit and n_R receive antennas can be represented by an n_T×n_R channel matrix $\boldsymbol{H}$. At time t, the channel matrix $\boldsymbol{H}$ is given by

$$\boldsymbol{H}_t = \begin{bmatrix} h_{1,1}(t) & h_{1,2}(t) & \cdots & h_{1,n_R}(t) \\ h_{2,1}(t) & h_{2,2}(t) & \cdots & h_{2,n_R}(t) \\ \vdots & \vdots & \ddots & \vdots \\ h_{n_T,1}(t) & h_{n_T,2}(t) & \cdots & h_{n_T,n_R}(t) \end{bmatrix} \quad (5)$$

where $h_{i,j}(t)$ is the correlated fading coefficient for the path from the i-th transmit antenna to the j-th receive antenna, and is modeled as a complex Gaussian random variable with zero mean and variance σ^2. For quasi-static fading channels, the fading coefficients are constant during a frame and vary from one frame to another, which means that the symbol period is small compared to the channel coherence time. So we can ignore the time t of the fading coefficient $h_{i,j}(t)$ to $h_{i,j}$.

A slightly less general but more useful model considers correlations on transmit and receive sides separately. The transmit correlation matrix is represented as $\boldsymbol{R}_T$ while receive correlation matrix is as $\boldsymbol{R}_R$, which can be represented as

$$\boldsymbol{R}_R = \boldsymbol{K}_R \boldsymbol{K}_R^H \quad (6)$$

where $\boldsymbol{K}_R$ is a $n_R \times n_R$ matrix and

$$\boldsymbol{R}_T = \boldsymbol{K}_T \boldsymbol{K}_T^H \quad (7)$$

where is $\boldsymbol{K}_T$ a $n_T \times n_T$ matrix and superscript H denotes the Hermitian (transpose conjugate) of a matrix. Matrices $\boldsymbol{K}_R$ and $\boldsymbol{K}_T$ are $n_R \times n_R$ and $n_T \times n_T$ lower triangular matrices with positive diagonal elements. They can be obtained from the respective correlation matrices $\boldsymbol{R}_R$ and $\boldsymbol{R}_T$ by Cholesky decomposition. The coefficient matrix in correlated channels, denoted by $\boldsymbol{H}$, is represented as [15]

$$\boldsymbol{H} = \boldsymbol{K}_R \boldsymbol{H}_w \boldsymbol{K}_T \quad (8)$$

where $\boldsymbol{H}_w$ is a i.i.d. channel fading coefficients matrix.

At the receiver, the signal at each receive antenna is a noisy superposition of the n_T transmitted signals degraded by channel fading. At time t, the received signals at the j-th antenna can be written as

$$r_j(t) = \sum_{i=1}^{n_T} h_{i,j} s_i(t, \boldsymbol{x}_i) + n_j(t), \quad j = 1, 2, \cdots, n_R \quad (9)$$

where $n_j(t)$ is the noise component of the j-th receive antenna at time t, which is modeled as a zero-mean complex Gaussian random variable with one sided power spectral density N_0 per dimension.

3. Performance Analysis

The optimum decoding and demodulation of OST-CPM system is maximum likelihood sequence detection (MLSD) algorithm. We assume that the decoder at the receiver has ideal channel state information (CSI) on the MIMO channel. The decoded codeword matrix $\hat{\boldsymbol{X}}$ can be expressed as

$$\hat{\boldsymbol{X}} = \arg\min_{\boldsymbol{X}} \left\{ \sum_{j=1}^{n_R} \int_0^{LT} \left| r_j(t) - \sum_{i=1}^{n_T} h_{i,j} s_i(t, \boldsymbol{x}_i) \right|^2 dt \right\} \quad (10)$$

The pair-wise error probability (PWEP) is the probability that the decoder selects as its estimate an erroneous matrix $\hat{\boldsymbol{X}} = \left[\hat{\boldsymbol{x}}_1, \hat{\boldsymbol{x}}_2, \cdots, \hat{\boldsymbol{x}}_{n_T}\right]^T$ when the transmitted matrix is in fact $\boldsymbol{X} = \left[\boldsymbol{x}_1, \boldsymbol{x}_2, \cdots, \boldsymbol{x}_{n_T}\right]^T$. In maximum likelihood decoding algorithm, that occurs if

$$\sum_{j=1}^{n_R}\int_0^{LT}\left|r_j(t) - \sum_{i=1}^{n_T} h_{i,j} s_i(t, \boldsymbol{x}_i)\right|^2 dt \geq \sum_{j=1}^{n_R}\int_0^{LT}\left|r_j(t) - \sum_{i=1}^{n_T} h_{i,j} s_i(t, \hat{\boldsymbol{x}}_i)\right|^2 dt \tag{11}$$

The above inequality is equivalent to

$$\sum_{j=1}^{n_R}\int_0^{LT} 2\mathrm{Re}\left\{n_j^*(t)\sum_{i=1}^{n_T} h_{i,j}\left[s_i(t,\hat{\boldsymbol{x}}_i) - s_i(t,\boldsymbol{x}_i)\right]\right\}dt \geq \sum_{j=1}^{n_R}\int_0^{LT}\left|\sum_{i=1}^{n_T} h_{i,j}\left[s_i(t,\hat{\boldsymbol{x}}_i) - s_i(t,\boldsymbol{x}_i)\right]\right|^2 dt \tag{12}$$

where *Re*{·} means the real part of a complex number.

It has been assumed that ideal CSI is available at the receiver. That is to say a realization of the fading coefficient matrix ***H*** is given. Therefore the term on the left side of (12) is a zero mean Gaussian random variable and the term on the right side is a constant equal to $d^2\left[\boldsymbol{S}(\boldsymbol{X}), \boldsymbol{S}(\hat{\boldsymbol{X}})\right]$. And $d^2\left[\boldsymbol{S}(\boldsymbol{X}), \boldsymbol{S}(\hat{\boldsymbol{X}})\right]$ is a modified Euclidean distance between the two OST-CPM signal matrices ***S***(***X***) and $\boldsymbol{S}(\hat{\boldsymbol{X}})$.

$$d^2\left[\boldsymbol{S}(\boldsymbol{X}), \boldsymbol{S}(\hat{\boldsymbol{X}})\right] = \sum_{j=1}^{n_R}\int_0^{LT}\left|\sum_{i=1}^{n_T} h_{i,j}\left[s_i(t,\hat{\boldsymbol{x}}_i) - s_i(t,\boldsymbol{x}_i)\right]\right|^2 dt \tag{13}$$

The PWEP conditioned on ***H*** can be approximated by [8]

$$P\left(\boldsymbol{X} \to \hat{\boldsymbol{X}} \mid \boldsymbol{H}\right) = Q\left(\sqrt{\frac{E_s}{2N_0} d^2\left[S(\boldsymbol{X}), S(\hat{\boldsymbol{X}})\right]}\right) \tag{14}$$

The modified Euclidean distance $d^2\left[\boldsymbol{S}(\boldsymbol{X}), \boldsymbol{S}(\hat{\boldsymbol{X}})\right]$ can be rewritten as

$$d^2\left[\boldsymbol{S}(\boldsymbol{X}), \boldsymbol{S}(\hat{\boldsymbol{X}})\right] = tr\left(\boldsymbol{H}\boldsymbol{B}\left(\boldsymbol{X}, \hat{\boldsymbol{X}}\right)\boldsymbol{H}^H\right) \tag{15}$$

where $\boldsymbol{B}\left(\boldsymbol{X}, \hat{\boldsymbol{X}}\right)$ is a $n_T \times n_T$ signal distance matrix with entries

$$\boldsymbol{B}\left(\boldsymbol{X}, \hat{\boldsymbol{X}}\right) = \begin{bmatrix} \int_0^{LT}\left|\Delta_1(t)\right|^2 dt & \int_0^{LT}\Delta_1^*(t)\Delta_2(t)dt & \cdots & \int_0^{LT}\Delta_1^*(t)\Delta_{n_T}(t)dt \\ \int_0^{LT}\Delta_1(t)\Delta_2^*(t)dt & \int_0^{LT}\left|\Delta_2(t)\right|^2 dt & \cdots & \int_0^{LT}\Delta_2^*(t)\Delta_{n_T}(t)dt \\ \vdots & \vdots & \ddots & \vdots \\ \int_0^{LT}\Delta_1(t)\Delta_{n_T}^*(t)dt & \int_0^{LT}\Delta_2(t)\Delta_{n_T}^*(t)dt & \cdots & \int_0^{LT}\left|\Delta_{n_T}(t)\right|^2 dt \end{bmatrix} \tag{16}$$

and $\Delta_i(t)$ is the difference signal

$$\Delta_i(t) = s_i(t, \boldsymbol{x}_i) - s_i(t, \hat{\boldsymbol{x}}_i) \tag{17}$$

For simplicity, we note $\Gamma = d^2\left[\boldsymbol{S}(\boldsymbol{X}), \boldsymbol{S}(\hat{\boldsymbol{X}})\right]$. By using Graig's formula for the Gaussian Q function

$$Q(x) = \frac{1}{\pi}\int_0^{\frac{\pi}{2}} \exp\left\{-\frac{x^2}{2\sin^2\theta}\right\} d\theta, \tag{18}$$

we can rewrite the conditional PWEP (14) as

$$P\left(\boldsymbol{X} \to \hat{\boldsymbol{X}} \mid \boldsymbol{H}\right) = \frac{1}{\pi}\int_0^{\frac{\pi}{2}} \exp\left\{-\frac{E_s}{4N_0 \sin^2\theta}\Gamma\right\} d\theta \tag{19}$$

In order to calculate the average PWEP, we average (19) with respect to the distribution of Γ. The average PWEP can be represented in terms of the moment generating function (MGF) of Γ, which is given by

$$M_\Gamma(s) = \int_0^\infty e^{s\Gamma} P_\Gamma(\Gamma) d\Gamma \tag{20}$$

Thus the average PWEP can be represented as

$$\begin{aligned} P\left(\boldsymbol{X} \to \hat{\boldsymbol{X}}\right) &= \frac{1}{\pi}\int_0^{\frac{\pi}{2}} E\left[\exp\left(-\frac{E_s}{4N_0\sin^2\theta}\Gamma\right)\right] d\theta \\ &= \frac{1}{\pi}\int_0^{\frac{\pi}{2}}\int_0^\infty \exp\left(-\frac{E_s}{4N_0\sin^2\theta}\Gamma\right) P_\Gamma(\Gamma)\, d\Gamma d\theta \\ &= \frac{1}{\pi}\int_0^{\frac{\pi}{2}} M_\Gamma\left(-\frac{E_s}{4N_0\sin^2\theta}\right) d\theta \end{aligned} \tag{21}$$

Therefore the PWEP can be represented as [16]

$$P\left(\boldsymbol{X} \to \hat{\boldsymbol{X}}\right) = \frac{1}{\pi}\int_0^{\frac{\pi}{2}} \prod_{i=1}^{n_T}\prod_{j=1}^{n_R}\left(1 + \frac{E_s}{4N_0\sin^2\theta}\mu_i\lambda_j\right)^{-1} d\theta \le \prod_{i=1}^{n_T}\prod_{j=1}^{n_R}\left(1 + \frac{E_s}{4N_0}\mu_i\lambda_j\right)^{-1} \tag{22}$$

where μ_i and λ_j are the eigenvalues of matrices $\boldsymbol{B}\left(\boldsymbol{X}, \hat{\boldsymbol{X}}\right)R_T$ and R_R respectively. Equation (22) is the Chernoff bound for the PWEP. When the SNR is high, the upper bound can be simplified as [3]

$$P\left(\boldsymbol{X} \to \hat{\boldsymbol{X}}\right) \le \left(\frac{E_s}{4N_0}\right)^{-r\hat{r}} \prod_{i=1}^{r}\prod_{j=1}^{\hat{r}}\left(\mu_i\lambda_j\right)^{-1} \tag{23}$$

where r and $\hat{r}$ are the ranks of $\boldsymbol{B}\left(\boldsymbol{X}, \hat{\boldsymbol{X}}\right)R_T$ and R_R respectively. $\mu_i (i = 1, 2, \cdots, r)$ and $\lambda_j (j = 1, 2, \cdots, \hat{r})$ are the nonzero eigenvalues of $\boldsymbol{B}\left(\boldsymbol{X}, \hat{\boldsymbol{X}}\right)R_T$ and R_R.

Then we will analyze the encoding performance variation of the OST-CPM system when the MIMO channels are correlated. In independent MIMO fading channels, the correlation coefficient matrices R_T and R_R are unit matrices with n_T and n_R nonzero eigenvalues respectively, which are defined as $\mu_i'(i=1,2,\cdots,n_T)=\lambda_j(j=1,2,\cdots,n_R)=1$. $\boldsymbol{B}(\boldsymbol{X},\hat{\boldsymbol{X}})$ and R_T are independent and R_T has full rank, thus the rank of $\boldsymbol{B}(\boldsymbol{X},\hat{\boldsymbol{X}})R_T$ is the same as the one of $\boldsymbol{B}(\boldsymbol{X},\hat{\boldsymbol{X}})$. The product of all the nonzero eigenvalues μ_i (i=1,2,…,r) of $\boldsymbol{B}(\boldsymbol{X},\hat{\boldsymbol{X}})R_T$ equals to the product of the nonzero eigenvalues $\mu_k''(k=1,2,\cdots,r)$ of $\boldsymbol{B}(\boldsymbol{X},\hat{\boldsymbol{X}})$ and the ones $\mu_i'(i=1,2,\cdots,n_T)$ of R_T.

$$\prod_{i=1}^{r}\mu_i=\prod_{k=1}^{r}\mu_k''\cdot\prod_{i=1}^{n_T}\mu_i'=\prod_{k=1}^{r}\mu_k'' \tag{24}$$

and

$$\prod_{i=1}^{n_T}\mu_i'=1,\ \prod_{j=1}^{n_R}\lambda_j=1 \tag{25}$$

For independent MIMO channels, the PWEP upper bound (23) of OST-CPM could be simplified to

$$P\left(\boldsymbol{X}\to\hat{\boldsymbol{X}}\right)\le\left(\frac{E_s}{4N_0}\right)^{-rn_R}\prod_{k=1}^{r}\left(\mu_k''\right)^{-1} \tag{26}$$

When the MIMO channel is partially correlated, matrices R_T and R_R are full rank. The PWEP upper bound could be represented as

$$P_{partial}\left(\boldsymbol{X}\to\hat{\boldsymbol{X}}\right)\le\left(\frac{E_s}{4N_0}\right)^{-rn_R}\prod_{i=1}^{n_T}\prod_{k=1}^{r}\prod_{j=1}^{n_R}\left(\mu_i'\mu_k''\lambda_j\right)^{-1} \tag{27}$$

For full rank matrices R_T and R_R with Toeplitz form, all the eigenvalues $\mu_i'(i=1,2,\cdots,n_T)$ and $\lambda_j(j=1,2,\cdots,n_R)$ are nonzero and the multiplication of all eigenvalues is smaller than 1. That is to say

$$\prod_{i=1}^{n_T}\left(\mu_i'\right)^{-1}>1,\ \prod_{j=1}^{n_R}\left(\lambda_j\right)^{-1}>1 \tag{28}$$

Consequently we can derive that

$$\left(\frac{E_s}{4N_0}\right)^{-rn_R}\prod_{k=1}^{r}\left(\mu_k''\right)^{-1}<\left(\frac{E_s}{4N_0}\right)^{-rn_R}\prod_{i=1}^{n_T}\prod_{k=1}^{r}\prod_{j=1}^{n_R}\left(\mu_i'\mu_k''\lambda_j\right)^{-1} \tag{29}$$

The PWEP upper bound in equation (27) is increased with independent MIMO channels. Finally we have deduced that the encoding performance of the OST-CPM system in partially correlated channels is degraded than the performance in independent channels.

4. Simulation Results

In this section, some simulation results are presented to evaluate the FER performances of the MLSD over quasi-static Rayleigh fading channels with spatial correlation.

OST-2CPM systems of full response CPM signals with two transmit antennas are considered. The number of receive antennas is one and two. The modulation index h is choosen to be 0.25 and 0.5. The transmit and receive antenna correlation matrices are given by $R_T = \begin{bmatrix} 1 & \rho_t \\ \rho_t & 1 \end{bmatrix}, R_R = \begin{bmatrix} 1 & \rho_r \\ \rho_r & 1 \end{bmatrix}$, for the two transmit and two receive antennas systems.

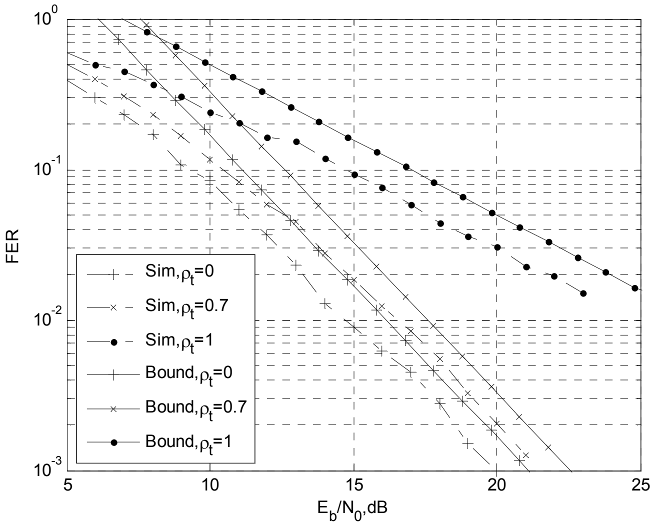

Figure 2. FER of OST-2CPM (h=0.25) with 2 transmit antennas and 1 receive antennas

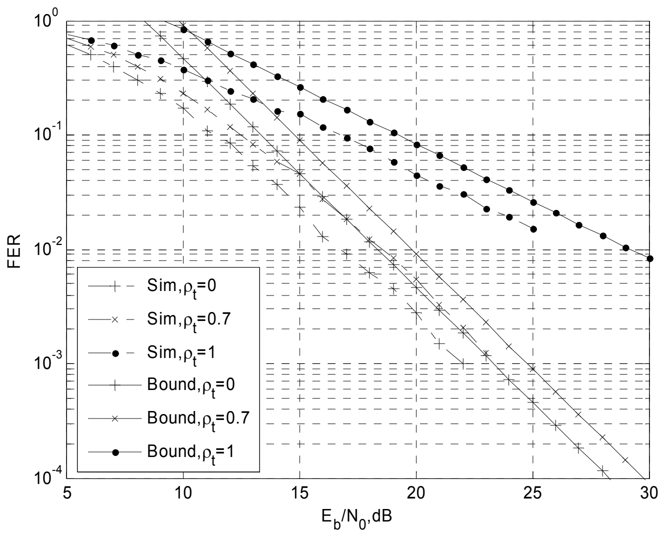

Figure 3. FER of OST-2CPM (h=0.5) with 2 transmit antennas and 1 receive antennas

Figure 2,
Figure 3 and
Figure 4 show the FER performances of the MLSD algorithm for the OST-2CPM system with different modulation index *h*. Simulation results show that the error performance over a correlated channel is degraded by approximately 2dB at a FER of 10^{-2} when the correlated index ρ_t increases from 0 to 0.7 for the one-receive OST-2CPM system. And the penalty on the code performance increases to over 9dB at the same FER in fully correlated channel. For the

two receive antennas OST-2CPM system the FER is degraded by approximately 3dB at a FER of 10^{-2} when the correlated index ρ_t increases from 0 to 0.7. It can also be observed from the two figures that the diversity gain decreases when the channel is fully correlated. The upper bounds can be closer when the diversity is high, e.g., if the number of the transmit or receive antennas is increased.

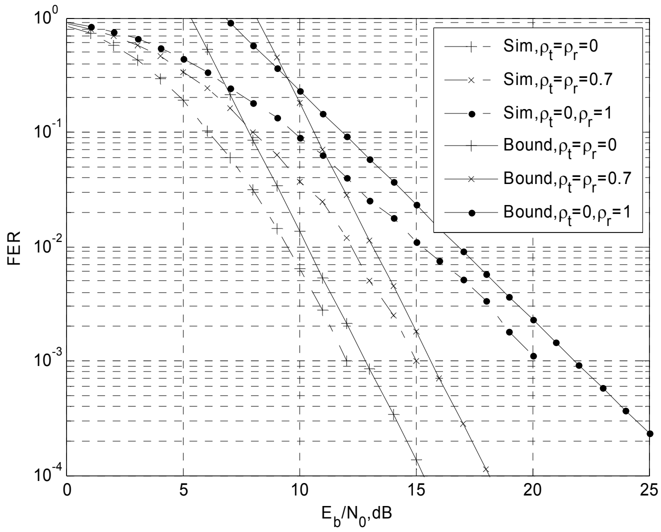

Figure 4. FER of OST-2CPM (h=0.5) with 2 transmit antennas and 2 receive antennas

5. Conclusion

The effect of the spatial correlation on MLSD of OST-2CPM system over Rayleigh fading channels was investigated in the paper. The PWEP upper bound of OST-2CPM systems was deduced to evaluate the coding performance over quasi-static fading channels with spatial correlation. Simulation results show that the FER performance of this system decreases as the signal correlation between the antennas increases. The simulation results well match the theoretical analysis.

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