Performance Analysis of MRT-MIMO for OFDM Systems over Fading Channels with Cochannel Interference

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Abstract: This article discusses the performance of Orthogonal Frequency Division Multiplexing (OFDM) in the presence of cochannel interference (CCI) and fading, while multiple-input-multiple output (MIMO) antenna technology is used. The transmitter and receiver weights of MIMO antenna arrays are adjusted jointly according to Maximum-Ratio transmission (MRT) criterion. Inter-carrier interference (ICI) results from the other sub-channels in the same data block of the same user is neglected. However, the effects of cross channel ICI produced by CCI due to carrier frequency offsets (CFO) are considered in the precise interference model. The error probability is calculated fast and accurately using a semi-analytical technique along with the Gauss quadrature rule (GQR) approach based on the method of moments, which can approximate the statistical distribution of the ICI.

Key words: Orthogonal Frequency Division Multiplexing (OFDM); Multi-input multiple-output (MIMO); Maximum-Ratio transmission (MRT); Cochannel Interference (CCI).

I. Introduction

The bit rates achieved in cellular and local area wireless communications systems have increased rapidly. The use of more complex modulation formats such as Orthogonal Frequency Division Multiplexing (OFDM) and higher order QAM has satisfied this requirement. The OFDM technology has been proposed for a range of standards and is widely applied to several existing high-speed wireless transmissions, such as Wireless Local Area Network (WLAN), Worldwide Interoperability for Microwave Access (WiMAX) systems and Long Term Evolution (LTE).

The most adverse effect from mobile radio systems suffer is mainly multipath fading and interference, which ultimately limit the quality of service offered to the users. With standard OFDM, very narrow transmissions can suffer from narrowband fading and interference. The OFDMA technology, which incorporates elements of time division multiple access (TDMA), allows subsets of the subcarriers to be allocated dynamically among the different users on the channel. The ability to schedule users by frequency provides resistance to frequency-selective fading. Therefore, it can achieve more frequency diversity. Moreover, as long as the cyclic extension (CP) is longer than the memory of the channel, successive OFDM symbols do not interfere with each other, and the receiver can be made very simple since no special processing, such as equalization, is needed to remove the effect of intersymbol interference (ISI). Therefore, OFDM has been shown its robustness in the presence of multipath dispersive propagation in high-speed wireless communications. In order to further increase the available system capacity, spatial processing using antenna arrays can be employed for supporting multiple users mobile radio system. The multi-input multiple-output (MIMO) design, an attractive technology to combat narrowband fading and interference, is suitable for OFDM transmission. Thus, MIMO-OFDM systems, which combine OFDM with MIMO, can promise significant increases in system performance.

In OFDM, uncorrected frequency errors will result in a loss of orthogonality among subcarriers and an inter-carrier interference (ICI). The signal frequency must be tracked continuously [1]. The ICI is different from the co-channel interference (CCI). The CCI is caused by reused channels in other cells which are allowed to transmit simultaneously on the same Resource Block (RB), while ICI results from the other sub-channels in the same data block of the same user. Even if only one user is in communication, ICI might occur, yet the co-channel interference will not happen. Several methods can suppress ICI induced by carrier frequency offsets (CFO)[1]-[7]. However, the effect of ICI may result from CCI, since CCI experiences different transmitter and channel. This ICI-like effect of CCI usually was neglected by most of previous researches [7]. For a TDMA system, the effects of cross channel inter-symbol interference (ISI) produced by CCI due to symbol timing offset were considered and applied to performance analysis of MRT-MIMO systems [8]. In fact, the ICI phenomenon in OFDM is the frequency-domain dual of intersymbol interference (ISI) that plagues single-carrier TDMA transmission over frequency-selective channels. However, frequency-selective ICI is assumed to be independent fading from subcarrier to subcarrier.

This paper presents the performance of OFDM with MIMO antenna technology in the presence of CCI over frequency-selective channels. The MIMO scheme is based on maximum ratio transmission (MRT) due to its simplicity. Perfect frequency tracking is assumed, and therefore ICI associated with synchronization error for the desired user vanishes. The focus is on the ICI effect generated by CCI due to the carrier frequency offsets (CFO). Similar to the ISI case, to evaluate the average BER for high-order QAM modulation under fading conditions by Monte Carlo simulations is exhaustive and time-consuming. We use a semi-analytical technology, Gauss quadrature rule (GQR), which can approximate the probability density function (pdf) of ICI-like CCI.

II. System Modeling

We first consider the OFDM system without the use of MIMO. For *M*-QAM, the transmitted complex symbol is given by $X(n) = a(n) + jb(n)$ having variance $\sigma_X^2 = 2(M-1)/3$, where $a(n)$, $b(n) = \pm 1, \pm 3, \ldots, \pm(\sqrt{M}+1)$ represent symbols on the in-phase and quadratue paths. With N subcarriers, the modulated signals can be expressed as a vector $\mathbf{X} = [X(0), X(1), \ldots, X(N-1)]^T$. The symbols are modulated with N subcarrier by passing a inverse discrete Fourier transform (IDFT) processor. The discrete Fourier transform (DFT) and inverse discrete Fourier transform (IDFT) in an $N \times N$ matrix form is given by

$$\mathbf{F} = \begin{bmatrix} 1 & 1 & \cdots & 1 \\ 1 & e^{\frac{-j2\pi}{N}} & \cdots & e^{\frac{-j2\pi(N-1)}{N}} \\ \vdots & \cdots & \cdots & \vdots \\ 1 & e^{\frac{-j2\pi(N-1)}{N}} & \cdots & e^{\frac{-j2\pi(N-1)^2}{N}} \end{bmatrix} \tag{1}$$

and

$$\mathbf{F}^H = \frac{1}{N}\begin{bmatrix} 1 & 1 & \cdots & 1 \\ 1 & e^{\frac{j2\pi}{N}} & \cdots & e^{\frac{j2\pi(N-1)}{N}} \\ \vdots & \cdots & \cdots & \vdots \\ 1 & e^{\frac{j2\pi(N-1)}{N}} & \cdots & e^{\frac{j2\pi(N-1)^2}{N}} \end{bmatrix}. \tag{2}$$

The IDFT signal can be written as $\mathbf{x} = [\mathbf{F}^H\mathbf{X}]^T$. The sampled signal $\mathbf{x} = [x(0), x(1), \ldots, x(N-1)]$ then passes through the parallel-to-series (P/S) circuit. The Cyclic Prefix (CP) of length P is added and then the transmit OFDM signal can be rewritten as $\mathbf{x}_{cp} = [x(-P), \ldots, x(-1), x(0), x(1), \ldots, x(N-1)]$.

The multipath channel can be expressed as

$$h(t) = \sum_{l=0}^{L} h(l)\delta[t - \tau(l)] \tag{3}$$

where L represents the number of paths and $\tau(l)$ is the value of the lth delay. Each delay beam $h(l)$ has Rayleigh-fading distribution and can be characterized as a complex envelope $h(l) = \alpha(l) + j\beta(l)$, where $\alpha(l)$ and $\beta(l)$ are zero-mean, i.i.d Gaussian random variables with variance σ_l^2. The number of paths L is less than the length of CP. At the receiver, the CP of received signal is removed, then the signals are passed to a DFT processor. The multipath channel can be expressed as a circular matrix

$$\mathbf{h} = \begin{bmatrix} h(0) & 0 & \dots & 0 & h(L) & \dots & h(1) \\ h(1) & h(0) & \vdots & \vdots & \vdots & \vdots & \\ & h(1) & h(0) & \vdots & \vdots & \vdots & h(L) \\ h(L) & \vdots & h(1) & \vdots & \vdots & \vdots & 0 \\ 0 & h(L) & \vdots & \vdots & h(0) & \vdots & \vdots \\ \vdots & \vdots & \vdots & \vdots & h(1) & h(0) & 0 \\ 0 & \dots & 0 & h(L) & \dots & h(1) & h(0) \end{bmatrix}_{N \times N}. \tag{4}$$

After the S/P and DFT processes, the received signal can be written as $\mathbf{Fhx} = \mathbf{FhF}^H\mathbf{X} = \mathbf{HX}$. Each subcarrier through multipath can be transfered into a single path with a diagonal matrix $\mathbf{H} = \mathbf{FhF}^H$.

$$\mathbf{H} = \begin{bmatrix} \eta(0) & 0 & \dots & \dots & 0 \\ 0 & \eta(1) & \vdots & \vdots & \vdots \\ \vdots & \vdots & \eta(2) & \vdots & \vdots \\ \vdots & \vdots & \vdots & \vdots & 0 \\ 0 & \dots & \dots & 0 & \eta(N-1) \end{bmatrix} \tag{5}$$

Parameters $\eta(0), \eta(1), \dots, \eta(N-1)$ are the channel responses corresponding to all subcarrier signals. They are shown to be independent from subcarrier to subcarrier. Each has a Rayleigh distribution with a common mean variance of $2\sum_{l=0}^{L}\sigma_l^2$.

Next, we consider the system with the use MIMO antenna technology. The MIMO system equipped with K transmit and R receive antennas over a CCI channel is shown in Fig. 1. Each subcarrier signal transmitted by the kth antenna is multiplied by a controllable complex weight vector $\mathbf{w}_t$. For convenience, the MIMO signal can be expressed in a matrix form. The channel gain of the nth subcarrier signal can be defined as a $R \times K$ matrix

$$\mathbf{H}(n) = \begin{bmatrix} \eta_{1,1}(n) & \eta_{2,1}(n) & \dots & \eta_{K,1}(n) \\ \eta_{1,2}(n) & \eta_{2,2}(n) & \dots & \eta_{K,2}(n) \\ \vdots & \vdots & \vdots & \vdots \\ \eta_{1,R}(n) & \eta_{2,R}(n) & \dots & \eta_{K,R}(n) \end{bmatrix}_{R \times K} \tag{6}$$

where $\eta_{k,m}(n)$ represents the channel parameter of the *n*th subcarrier between the *k*th transmit antenna and the *m*th receive antenna. The $K \times 1$ weight vector at the transmitter and the $R \times 1$ weight vector at the receiver are defined as $\mathbf{w}_t(n) = [w_1^t(n), w_2^t(n), w_3^t(n), \ldots, w_K^t]^T$ with $\|\mathbf{w}^t(n)\|^2 = 1$ (i.e. average transmit power is restricted to be constant) and $\mathbf{w}_r(n) = [w_1^r(n), w_2^r(n), w_3^r(n), \ldots, w_R^r]^T$, respectively. By exploiting the correlation between adjacent subcarrier channels, it is possible to use the same weight for a number of subcarriers.

In a MIMO system employing a maximum-ratio transmission (MRT) scheme, signals are combined in such a way that the overall output signal-to-noise (SNR) of the system is maximized, where CCI is ignored. The input noise is a zero-mean white Gaussian noise with double-sided power spectral density of N_0 W/Hz. Based on the Maximum-ratio-combining (MRC) scheme, we have $\mathbf{w}_r(n) = [\mathbf{H}(n)\mathbf{w}_t(n)]^*$, where * denotes the complex conjugate operation. It follows that the output SNR is given by

$$(SNR)_o = \frac{\sigma_X^2 \|\mathbf{w}_r^T(n)\mathbf{H}(n)\mathbf{w}_t(n)\|^2}{\sigma_v^2 \|\mathbf{w}_r^T(n)\|^2} = \frac{\sigma_X^2}{N_o}\mathbf{w}_t^H(n)\mathbf{H}^H(n)\mathbf{H}(n)\mathbf{w}_t(n) \tag{7}$$

where $(\cdot)^H$ is the conjugate transpose operator. Maximizing SNR can be accomplished by choosing the weight vector $\mathbf{w}_t(n)$ that maximizes the quadrature form $\mathbf{w}_t^H(n)\mathbf{H}^H(n)\mathbf{H}(n)\mathbf{w}_t(n)$ subject to the constraint $\mathbf{w}_t^H(n)\mathbf{w}_t(n) = 1$.

It is known that $\mathbf{w}_t^H(n)\mathbf{H}^H(n)\mathbf{H}(n)\mathbf{w}_t(n)$ can be maximized by finding the maximum eigenvalue of $K \times K$ Hermitian matrix $\mathbf{H}^H(n)\mathbf{H}(n)$. Based on this fact, we can choose the transmitting weight vector as $\mathbf{w}_t(n) = \mathbf{V}_{max}(n)$, the unitary eigenvector corresponding to the largest eigenvalue, $\Omega_{max}(n)$, of the quadrature form $\mathbf{H}^H(n)\mathbf{H}(n)$. The corresponding maximum SNR is given by $(\sigma_X^2 / N_0)\Omega_{max}(n)$. Choosing this receive antenna vector results in $\|\mathbf{w}_r^T(n)\|^2 = \mathbf{w}_r^H(n)\mathbf{w}_r(n) = \mathbf{w}_t^H(n)\mathbf{H}^H(n)\mathbf{H}(n)\mathbf{w}_t(n) = \Omega_{max}(n)$.

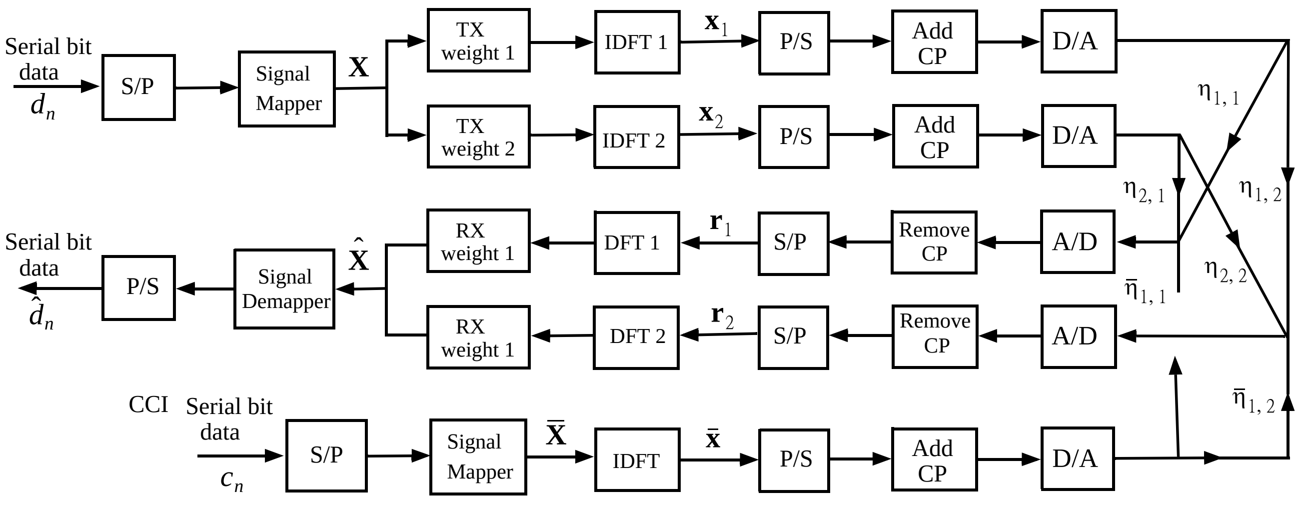

Fig. 1 MRT-MIMO OFDM (two transmit antennas and two receive antenna, one CCI)

Since the CCI transmit weights are not controlled by the desired receiver, the transmit weights of CCI can be neglected. Similarly, the channel complex gain for I cochannel interferers on the nth subchannel can be written in a $R \times I$ matrix form as

$$\bar{\mathbf{H}}(n) = \begin{bmatrix} \bar{\eta}_{1,1}(n) & \bar{\eta}_{2,1}(n) & \ldots & \bar{\eta}_{I,1}(n) \\ \bar{\eta}_{1,2}(n) & \bar{\eta}_{2,2}(n) & \ldots & \bar{\eta}_{I,2}(n) \\ \vdots & \vdots & \vdots & \vdots \\ \bar{\eta}_{1,R}(n) & \bar{\eta}_{2,R}(n) & \ldots & \bar{\eta}_{I,R}(n) \end{bmatrix}_{R \times I}, \quad (8)$$

where $\bar{\eta}_{i,m}(n)$ represents the channel parameter on the mth receive antenna for the ith interfering signal. The symbol and channel of a CCI is indexed by a bar.

The average signal-to-noise ratio (SNR) of the OFDM received signal can be derived as

$$SNR = \frac{\sigma_X^2}{N_0}\sum_{l=1}^{L}|h_l|^2 = \frac{2\sigma_X^2}{N_0}\sum_{l=1}^{L}\sigma_l^2 \quad (9)$$

The average signal-to-interference ratio (SIR) is derived as

$$SIR = \sum_{l=1}^{L}\sigma_l^2 \Big/ I\sum_{l=1}^{L}\bar{\sigma}_l^2 \quad (10)$$

where σ_l and $\bar{\sigma}_l$ represent the standard deviations of the Gaussian random variables on the lth delay beam for the desired signal and interfering signals. Equal powers are assumed for the interfering signals.

In OFDM systems, ICI mainly is caused by carrier frequency offset (CFO), due to imperfect OSC or Doppler effects. This may result in circulation of the phase in time domain for each subcarrier. Assuming that CFO generated by the desired signal can be improved, the effect of ICI may result from CCI since CCI transmitter is not controlled by the system transmitting the desired signal. The intercarrier interference (ICI) due to the CFO can be expressed as

$$C(n) = \frac{\sin(\pi(-n+\varepsilon))}{N\sin(\pi(-n+\varepsilon)/N)} e^{j\pi\varepsilon(N-1)/N} e^{-j\pi(-n/N)} \quad (11)$$

where $n = 0, 1, \ldots, N-1$ and ε is the frequency offset. Unlike the ISI case, ICI is assumed to be independent fading from subcarrier to subcarrier. Hence, the resulting ICI term $C(n)$ has to be adjusted by different channel parameters $\bar{\eta}_{i,m}(n)$.

III. Error Probability Estimation

The price paid for using Monte Carlo methods is the run time required for executing the simulation in the presence of ICI. If the system and the channel model are complicated and the BER is low, the required run time is sometimes so long that the use of Monte Carlo techniques becomes impractical for all but the most important simulations, especially in the fading case. We use the semi-analytic approach for performance estimation. Such an approach avoids a Gaussian characterization of interference, which is not realistic because interference has the effects of waveform, modulation and fading. Semi-analytic error probability estimation can be sped up considerably by combining the transmitter, the channel, the receiver into one single response.

1.1 Precise Interference Model

For the MIMO-OFDM case with a precise CCI model, the estimate of the symbol on the *l*th subcarrier can be expressed as

$$\hat{X}(l) = [a(l) + jb(l)]\rho_l + (\xi + j\zeta) + \nu(l) \tag{12}$$

where $\rho_l = \mathbf{w}_r^T(l)\mathbf{H}(l)\mathbf{w}_t(l)$ which is equal to $\Omega_{max}(l)$, the largest eigenvalue of the matrix $\mathbf{H}^H(l)\mathbf{H}(l)$. We define the symbol of the *n*th subchannel for the *i*th CCI as $X_i(n) = a_i(n) + jb_i(n)$. The combined ICI in the in-phase rail can be denoted by

$$\xi = \sum_{i=1}^{I}\left[\sum_{n=0}^{N-1} a_i(n)p_i(n) - \sum_{n=0}^{N-1} b_i(n)q_i(n)\right] \tag{13}$$

and ICI in the quadrature rail is

$$\zeta = \sum_{i=1}^{I}\left[\sum_{n=0}^{N-1} a_i(n)q_i(n) + \sum_{n=0}^{N-1} b_i(n)p_i(n)\right] . \tag{14}$$

With defining $C(n)$ in (11), the ICI term from the *n*th subchannel of the *i*th interfering signal on the *m*th receive antenna is multiplied by the channel response $\bar{\eta}_{i,m}(n)$. The response becomes $g_{i,m}(n) = \bar{\eta}_{i,m}(n)C(n)$. Defining $g_{l,m}(n) = g_{l,m}^I(n) + g_{l,m}^Q(n)$ results in the responses $p_i(n)$ and $q_i(n)$

$$p_i(n) = \sum_{m=1}^{R} w_{I,m}^r(l)g_{l,m}^I(n) - w_{Q,m}^r(l)g_{l,m}^Q(n) \tag{15}$$

$$q_i(n) = \sum_{m=1}^{R} w_{I,m}^r(l)g_{l,m}^Q(n) + w_{Q,m}^r(l)g_{l,m}^I(n) \tag{16}$$

where $w_m^r(l) = w_{I,m}^r(l) + jw_{Q,m}^r(l)$ represents the *m* receive weights of the *l*th subcarrier.

The weighted discrete-time noise on the *m*th receive antenna has the output power (variance) $[[w_{I,m}^r(l)]^2 + [w_{Q,m}^r(l)]^2]N_0$. Since the noise is uncorrelated between diversity paths, the variance of the combined output noise, $\nu(l)$, is expressed as

$$\sigma_\nu^2 = N_0 \sum_{m=1}^{R} [w_{I,m}^r(l)]^2 + [w_{Q,m}^r(l)]^2 . \tag{17}$$

We define $v(l) = v_I(l) + jv_Q(l)$ where $v_I(l)$ and $v_Q(l)$ have equal variance, $\sigma^2 = \sigma_v^2/2$. Since the distribution density functions of quantities ξ and ζ are symmetric to zero and are identical, it has been shown that the average symbol error probability P_M can be bounded tightly by

$$P_M = 2E[P_I(\xi)] = 2\left(1 - \frac{1}{\sqrt{M}}\right) E\left[\text{erfc}\left(\frac{\rho_l + \xi}{\sqrt{2}\sigma}\right)\right]. \tag{18}$$

Because ξ is a random variable whose distribution is not known explicitly, the evaluation of $E[e(\xi)]$ is performed by computing the conditional error probability of each of all possible sequences of CCI, and then averaging over all those sequences. For (18), $e(\cdot)$ is given by $\text{erfc}(\cdot)$.

The semi-analytical technique in (18) is computationally very efficient compared to the Monte-Carlo method. However, this approach is cumbersome and may be computationally infeasible if a large number of cross-channel ICI symbols (e.g. with high order of modulation) are included or/and more than one interferer are present, especially when dealing with low error rates and MIMO systems. Some techniques can be used for evaluation of numerical approximations to $E[e(\xi)]$. One efficient approach called the Gaussian quadrature rule (GQR) approximation will be addressed for the numerical evaluation of (18), which depends on knowing the moments of ξ, up to an order that depends on the accuracy required.

Using the Gaussian quadrature rule, the averaging operation in (18) can be approximated by

$$E[e(\xi)] = \int_a^b e(x) f_\xi(x) dx \cong \sum_{i=1}^{N} w_i\, e(x_i), \tag{19}$$

a linear combination of values of the function $e(\cdot)$, where $f_\xi(x)$ denotes the probability function of the random variable ξ. The weights (or coefficients) w_i, and the abscissas x_i, $i = 1, 2, \ldots, P$ can be calculated from the knowledge of the first $2P + 1$ moments of ξ. We compute the average in (19) by means of the classic GQR's. The precise BER results are obtained by using a combination of analysis and simulation under fading conditions.

For the ICI term ξ in (18), we can assume that there are N_1 terms in the first summation and N_2 terms in the second for each interferer. We assume that $N_s = I(N_1 + N_2)$. The random variable ξ is the sum of N_s ICI terms for the multiple CCI case. The ICI ξ can be rewritten as

$$\xi = \sum_{j=1}^{N_s} I(j)x(j) = \sum_{j=1}^{N_s} y(j) \tag{20}$$

where $I(j)$ represents a discrete random variable, $a_i(n)$ or $b_i(n)$, whose moments are given and $x(j)$ is a sequence of known constants $p_i(n)$ or $q_i(n)$. It is suggested that we reorder the sequence $y(j)'s$ so that $max|y(j)| \geq max|y(j+1)|$ i.e. $|x(j)| \geq |x(j+1)|$, $1 \leq i \leq N_s - 1$. This reordering lets the moments of the dominant terms be computed first and rolloff error be minimized. A recursive algorithm which can be used to determine the moments of all order of ξ.

1.2 Gaussian Interference Model

To simplify the analysis and make it both computationally and mathematically tractable, an alternative approach, Gaussian interference model, for representing the cochannel interference is often used. A Gaussian model assumed that all interfering signals have no CFO effect relative to the desired signal and did not consider cross-channel ICI effects. In this model, the interference contribution is represented by a Gaussian noise with mean and variance equal to the mean and variance of the sum of the interfering signals. In our simulation, the accuracy is assessed by comparing their BER performances with precise BER results.

Using the Gaussian interference model, the MRT scheme is optimum for the MIMO system. The average power of each interferer received by the *m*th receive antenna element is

$$E[|\bar{\eta}_{i,m}(n)X_i(n)|^2] = 2\sum_{l=1}^{L}\bar{\sigma}_l^2 E[X_i(n)] = 2\sum_{l=1}^{L}\bar{\sigma}_l^2 N_I \tag{21}$$

where $X_i(n)$ is assumed to be Gaussian distributed and has power spectrum density N_I. Thus, the SIR ratio per diversity branch can be defined as

$$SIR = \sigma_X^2\sum_{l=1}^{L}(\sigma_l^d)^2 \Big/ IN_I\sum_{l=1}^{L}(\sigma_l^c)^2 \tag{22}$$

The output power of combined interference is then given by

$$\sigma_{\omega}^2 = \sum_{m=1}^{R}\sum_{i=1}^{I}\bar{\eta}_{i,m}(n)[(w_{I,m}^r)^2 + (w_{Q,m}^r)^2]N_I. \tag{23}$$

The total output power of the interference plus noise is $\sigma_{\mu}^2 = \sigma_{\omega}^2 + \sigma_v^2$, where σ_v^2 is given in (17). The symbol error probability for fading Gaussian interference is given by

$$P_M = 2\left(1 - \frac{1}{\sqrt{M}}\right)\left[\operatorname{erfc}\left(\frac{\rho_I}{\sqrt{2}\sigma}\right)\right] \tag{24}$$

where $\sigma^2 = \sigma_{\mu}^2/2$ represents the variance in each rail. Unlike the precise CCI model, the interfering signal becomes uncorrelated from branch to branch under this assumption. As a result, the Gaussian interference model usually overestimates the effect of CCI in nonfading channel. The accuracy of the Gaussian interference model usually depends on the statical characteristics of the channel and the MIMO scheme.

IV. Simulation Results

Computer simulations are conducted to demonstrate the performance of the MRT-MIMO OFDM with 4-QAM modulation. Herein, we consider the case with a single cochannel interferer and the case with three interferers. We use a 10-order GQR that needs 20 moments of 82 joint interference complex terms. Both the desired signal and CCI are subject to frequency-selective fading. Independent Rayleigh fading from subcarrier to subcarrier is assumed. Parameter ε of CFO is fixed at 0.3 and the SIR is set into 10dB.

The simulation results of a MRT-MIMO system with three receive antennas for the interference with and without CFO are shown in Figs 2. Using the MRT approach, CCI is not eliminated and then the error rate is irreducible due to the residual CCI. The performance with a CFO is worse and appears a higher irreducible floor due to the cumulative ICI-like CCI. A comparison of the precise CCI model against the Gaussian CCI model is shown in Figs 3. It is seen that the curves of Gaussian CCI and the precise CCI appear different with the increase of the transmit and receive antennas. The Gaussian CCI model always overestimates the performance. We note that the performance with $I = 3$ is better than that with $I = 1$. This is due to the fact that with total interference power equally distributed among three cochannel interferers, the probability that at least one of the interferers is strongly faded is greater in the case of multiple interferers, thus leading to a smaller error rate. In general, the assumption of zero-CFO interference is optimistic, while the Gaussian interference is too pessimistic.

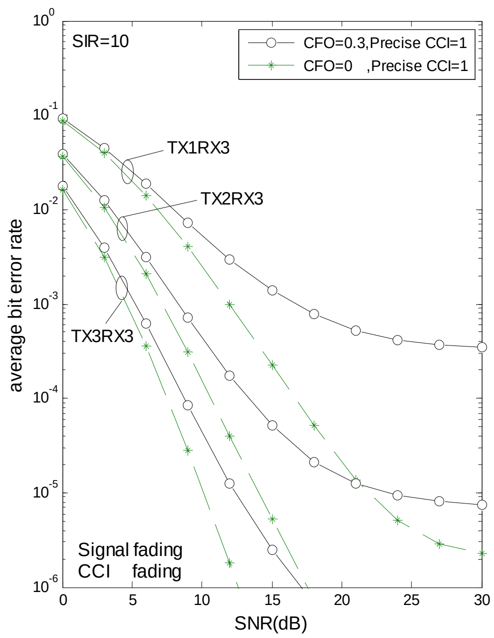

Fig. 2 Average BER for MRT-MIMO OFDM with CFO = 0 and 0.3. (SIR = 10dB)

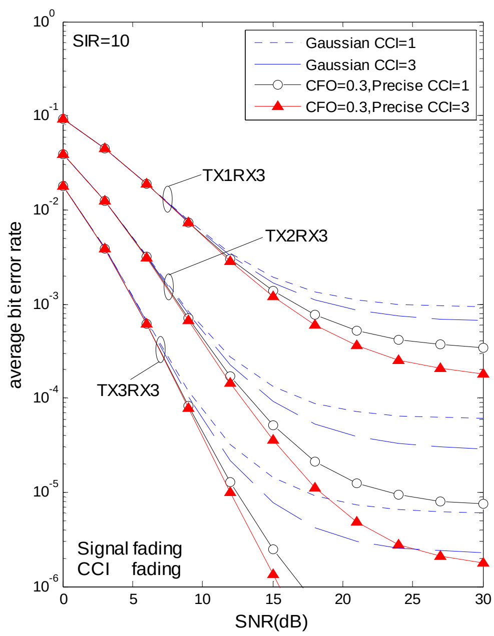

Fig. 3 Average BER for MRT-MIMO OFDM using precise CCI and Gaussian CCI models. (SIR = 10dB)

V. Conclusions

In this paper, we have analyzed the performance of MRT-MIMO based OFDM systems subject to cochannel interference (CCI) operating frequency-selective fading channels. The use of precise CCI model and Gaussian Quadrature rule (GQR) provides significant improvement in the performance analysis. The results of this study are expected to lead to a better understanding of the effects of interference, and then to optimize spectrum reuse in MIMO-OFDM systems.

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