Performance Analysis for Optimum Transmission and Comparison with Maximal Ratio Transmission for MIMO Systems with Cochannel Interference

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Abstract – This paper presents the performance analysis of optimum transmission (OT) and maximum-ratio transmission (MRT) for multiple-input/multiple-output (MIMO) systems subject to cochannel interference (CCI). We consider digital cellular mobile radio systems with quadrature amplitude modulation (QAM) transmission operating over both nonfading and flat Rayleigh fading channels. Our analysis accounts for pulse waveform and modulation of the desired signal, as well as cochannel interference (CCI). The phase delay between the desired signal and CCI, which can result in symbol timing and phase offsets for each receiving antenna element, are taken into account. A Gaussian interference model is also considered for a comparison with the precise interference model in the MRT based MIMO systems.

I. INTRODUCTION

The most adverse effect from mobile radio systems suffer is mainly multipath fading and co-channel interference (CCI), which ultimately limit the quality of service offered to the users. Space diversity combining with a single antenna at the transmitter and multiple antennas at the receiver (SIMO) provides an attractive means to combat multipath fading of the desired signal and reduces the relative power of co-channel interfering signals. A practical and simple diversity combining technique is maximal ratio combining (MRC), which is optimal in the presence of spatially white Gaussian noise. MRC mitigates fading, however, it ignores CCI. Optimum combining (OC), in which the combiner weights of OC need to be adjusted to maximize the output signal to interference-plus-noise ratio (SINR), can resolve the problem in the presence of CCI.

The performance of OC was studied for both nonfading [1] and fading [2]-[8] communication systems. Whereas publications in the area dealt with SIMO, applications in more recent years have become increasingly sophisticated, thereby relying on the more general multiple-input/multiple-output (MIMO) antenna systems which promise significant increases in system performance and capacity. With no CCI, the performance of MIMO systems based on Maximum Ratio Transmission (MRT) in a Rayleigh fading channel was studied in [9]-[11]. In the presence of CCI, the outage performances based on Maximum Ratio Transmission (MRT) [12] and Optimum Transmission (OT) [13][14] were studied. In general, the analyses of above SIMO and MIMO systems is on obtaining closed-form expressions, and adopt the following assumptions: 1) The number of interferers exceeds the number of antenna elements, and the antenna array is unable to cancel every interfering signal [14]. At this point, the interference is approximated by Gaussian noise. 2) The phase of each interferer relative to the desired signal for each diversity branch is neglected, and thus phase tracking and symbol synchronization are not only perfect for the desired signal, but also for CCI [2]-[14]. 3) Average powers of interferers are assumed to be equal, which is valid in the case these interferers are approximately at the same distance from the receiver [2]-[14]. 4) The effect of thermal noise is neglected, which is reasonable for interference limited systems [5][14]. Based on the above assumptions, the SINR distribution is derived and enables analytical computation of performance measures.

Multiple interference meets the conditions of the central limit theorem, hence, it can be assumed Gaussian (nonfading). The noise approximation model is simplistic, but was shown to be inaccurate for the case of a few dominant interferers. Moreover, all of the early work mentioned above did not consider pulse waveform and the modulation characteristics of the signal of interest, as well as CCI. The effects of ISI produced by CCI due to symbol timing were neglected. In [6][7], the bit error rate (BER) of PSK operating in several different flat fading environments was analyzed using the precise cochannel interference model. This paper studies the exact average BER of OT and provides the comparison with MRT using the precise cochannel interference model when the desired signal and interference are subject to nonfading and Rayleigh fading. Both the cases of a single and six interferers are analyzed. With the multiple ISI-like CCI sources, the simulation is expected to be very tedious and time-exhausting in MIMO systems. Therefore, we estimate the error probability for each fading channel fast and accurately using Gauss quadrature rules (GQR).

II. SYSTEM MODELS

We consider a MIMO QAM system equipped with $T$ antenna elements at the transmitter and $R$ antenna elements at the receiver as shown in Fig. 1. It is assumed that there exist totally $L$ cochannel interferers from the neighboring cells. The transmitted QAM baseband signal from the desired signal can be expressed in the form

$$s_{D,k}(t) = \sum_{n} c_{0,n} g_{n}(t-nT_{c})w_{k}^{n}$$

where $w_{k}^{n}$ represents the transmit weight on the $k$th antenna ($k=1,...,T$). Since the CCI transmit weights are not controlled by the desired receiver, the transmit weights of CCI

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can be neglected. Thus, the $i$th transmit cochannel interference can be combined as

$$s_{i,t}(t) = \sum_{i} c_{i,n} g_{i}(t-nT_s)$$

(2)

where $c_{i,n} = a_{i,n} + j b_{i,n}$ is the sequence of complex data symbols and $T_s$ is the symbol interval. The data symbols $a_{i,n}$ and $b_{i,n}$ on the in-phase and quadrature paths define the signal constellation of the QAM signal with $M$ points. In the constellation, we take $a_{i,n}, b_{i,n} = \pm 1, \pm 3, \ldots, \pm \sqrt{M+1}$. The transmitter filter gives a pulse $g_{i}(t)$ having the square-root raised-cosine spectrum with a rolloff factor. The desired symbol is indexed by $i = 0$, and the CCI sources by $i = 1, \ldots, L$ for (CC1).

The channel is assumed to be spatially independent flat Rayleigh fading. The complex channel gain between the $i$th transmit antenna and $m$th receive antenna for the desired signal can be represented by $h_{D,i,m} = \lambda_{i,m} e^{j\theta_{i,m}}$, where $\lambda_{i,m}$ is the envelope with Rayleigh distribution having variance $\sigma_\theta^2 = E[\lambda_{i,m}^2]$. The complex gain channel between the $i$th receive antenna and $m$th receive antenna can be represented by $h_{R,i,m} = \lambda_{i,m} e^{j\theta_{i,m}}$ with variance $\sigma_\theta^2 = E[\lambda_{i,m}^2]$. Phase $\theta_{i,m}$ and $\theta_{i,m}$ have a uniform distribution in $[0, 2\pi]$. With zero-mean information symbols, the average power of the $i$th cochannel interferer received by the $m$th diversity antenna is derived as $\sigma_i^2 \sigma_j^2 / T_s$, where $\sigma_i^2 = E[|\lambda_{i,m}^2|]$ represents the data symbol variance for all cochannel sources. For a $M$-QAM system, $\sigma_i^2 = 2(M-1)/3$. The input noise $n_{i,m}(t)$ is a zero-mean AWGN with two-sided power spectral density of $N_0 / W$. Thus, the noise power measured in the Nyquist band is $N_0 / T_s$. 

The power of the transmit weight vector is restricted to be one for the desired signal. Therefore, the average value of the SNR on each diversity branch is defined by $\sigma_i^2 \sigma_j^2 / N_0$. Equal average power is assumed for all received interferers, and therefore, we set $\sigma_i^2 = \sigma_j^2$ for $i = 1, \ldots, L$. The signal-to-interference power ratio (SIR) per diversity branch can be denoted by $SIR = \sigma_i^2 / \sigma_j^2$.

At the receiver, we assume that the frequency and symbol synchronization are perfect for the desired signal. As shown in Fig. 1, after matching and sampling at $t = nT_s$, the signal received at the $m$th diversity antenna is given by

$$r_m(\text{IT}_m) = c_{0,m} \sum_{l=1}^{T} w_{h_{D,l},m} \sum_{k=1}^{L} h_{D,l,m} g_{k}(t-nT_s) + v_m(\text{IT}_m)$$

(3)

where $\tau_i$ is the relative symbol timing offset between the desired signal and the $i$th interferer, is assumed to be the same for all diversity branch based on the principle of microscopic diversity; pulse response $g(t)$ having the raised-cosine spectrum is the combined transmitter filter $g_{i}(t)$ and receiver filter $g_{i}(t)$ which have the same response; the filtered noise is $v_{i,m}(t) = n_{i,m}(t) \otimes g_{i}(t)$ and the power is calculated as $\sigma_i^2 = N_0$ where $\otimes$ denotes the convolution operation. Since noise is wide stationary (WSS) and the power is independent of sampling instance, we have $E[v_m(\text{IT}_m)] = \sigma_i^2 = N_0$. The signal from the $m$th receive branch is weighted by a complex weight $w_{m}$. The output of the combiner has the form

$$\hat{c}_{0,m} = c_{0,m} \sum_{l=1}^{T} w_{m}^{*} \sum_{k=1}^{L} h_{D,l,m} g_{k}(t-nT_s) \ldots \sum_{m=1}^{R} w_{m}^{*} v_m(\text{IT}_m)$$

(4)

$$+ \sum_{l=1}^{L} \sum_{m=1}^{R} w_{m}^{*} h_{D,l,m} c_{l,n} g_{n}(t-nT_s - \tau_i)$$

For convenience, the MIMO signal can be expressed in a matrix form. The channel gain for the desired user can be defined as a $R \times T$ matrix

$$H_D = \begin{bmatrix} h_{D,1,1} & h_{D,1,2} & \cdots & h_{D,1,T} \\ h_{D,2,1} & h_{D,2,2} & \cdots & h_{D,2,T} \\ \vdots & \vdots & \ddots & \vdots \\ h_{D,L,1} & h_{D,L,2} & \cdots & h_{D,L,T} \end{bmatrix}$$

(5)

and the $L$ cochannel interferers can be written in a $R \times L$ matrix form as

$$H_I = \begin{bmatrix} h_{I,1,1} & h_{I,1,2} & \cdots & h_{I,1,L} \\ h_{I,2,1} & h_{I,2,2} & \cdots & h_{I,2,L} \\ \vdots & \vdots & \ddots & \vdots \\ h_{I,L,1} & h_{I,L,2} & \cdots & h_{I,L,L} \end{bmatrix}$$

(6)

The $T \times 1$ weight vector at the transmitter and the $R \times 1$ weight vector at the receiver are defined as $w = [w_1^T, w_2^T, w_3^T, \ldots, w_L^T]^T$ and $w = [w_1^T, w_2^T, w_3^T, \ldots, w_L^T]^T$. The output defined in Eq. (4) is

$$\hat{c}_{0,m} = c_{0,m} w^H w_D \hat{c}_g + w^H H_I [c g] + w^H v$$

(7)

where $w = [c_1, c_2, c_3, \ldots, c_L]^T$, a $L \times 1$ vector, represents ISI due to all interfering signals with $g_i = [g_{(N-\tau_i)}, g_{-\tau_i}, \ldots, g_{-\tau_i}]/2$, which are the $2N+1$ truncated samples of the raised-cosine pulse due to the delay offset $\tau_i$ from the $i$th interferer, and $c_i = [c_i(-\tau_i), \ldots, c_i(0)], \ldots, c_i(NT_i)]$, which is the symbol sequence of the $i$th interferer. The vector $\mathbf{v} = [v_1(\text{IT}_1), v_1(\text{IT}_2), \ldots, v_1(\text{IT}_T)]$ represents $R$ discrete filtered noise sources at the receiver. The weight vectors $w$ and $w$ can be determined using MRT and OT methods.

A. MRT weight for MIMO

In a MIMO system employing MRT scheme, signals are combined in such a way that the overall output signal-to-noise (SNR) of the system is maximized. Based on the Maximum-ratio-combining (MRC) scheme, we have $w = (H^H w)^H$, where $^*$ denotes the complex conjugate operation. It follows that the SNR is given by
where \((\cdot)^T\) is the conjugate transpose operator. Maximizing SNR can be accomplished by choosing the weight vector \(w\), that maximizes the quadrature form \(w^H H w\). It is known that \(w^H H w\) can be maximized by finding the maximum eigenvalue of \(H^H H\), subject to the constraint \(w^H w = 1\). The resulting SINR \(\frac{\sigma^2}{\lambda_{\max}}\) corresponding to the largest eigenvalue, \(\lambda_{\max}\), of the quadrature form \(H^H H\). The corresponding maximum SINR is given by \((\sigma^2/N_0)\lambda_{\max}\). Choosing this receive antenna vector results in \(\left\|w^H w_r\right\| = w^H H w = \lambda_{\max}\).

B. Optimal weight for MIMO with CCI

In the presence of CCI, the optimal strategy is to choose the transmission and combining weights to maximize the signal-to-interference plus noise ration (SINR). We can find the optimum weight \(w\), given that \(w\) is known. The difficulty, however, is how to determine the optimum \(w\). Those optimum weights can be determined based on the mean square error

\[\text{MSE} = E[|e_i - \hat{c}_i|^2]\]

to minimize interference-plus-noise conditioned on the fixed desired signal [13]. The receiving weight vector that minimizes the SINR in terms of MSE is given by the well known relation

\[w_r = R^{-1}(H_Dw)^*\]  \hspace{1cm} (9)

where \(R\) is a \(r \times r\) Hermitian covariance matrix of CCI and can be expressed as [13][14]

\[R = (1 - \beta/4)H^H H + \frac{N_0}{\sigma^2} I\]  \hspace{1cm} (10)

with roll of factor \(\beta\), when the random relative carrier and symbol timing offsets are considered [4]. \(I\) is the identity matrix of dimension \(r\). The factor \((1 - \beta/4)\) was not considered in [13][14]. With the constraint \(w^H w = 1\). The transmitting weight vector \(w = V_{\text{max}}\) denotes the unitary eigenvector corresponding to the largest eigenvalue, \(\lambda_{\max}\), of the quadratic form \(H^H H\). The resulting SINR [13]

\[\text{SNR} = \frac{\sigma^2}{\lambda_{\max}}\left\|w^H w_r\right\|^2 = \frac{\sigma^2}{\lambda_{\max}}\left\|w^H H w\right\|^2 = \frac{\left\|w^H H w\right\|^2}{\frac{w^H H w}{\sigma^2 + (1 - \beta/4)\left\|H^H H\right\|^2}} = \frac{w^H H w}{\frac{w^H H w}{\sigma^2 + (1 - \beta/4)\left\|H^H H\right\|^2}}\]  \hspace{1cm} (11)

By substituting (9) into (11), it follows that the SINR for a given \(w\), can be written as

\[\text{SNR} = H^H H\]  \hspace{1cm} (12)

Therefore, the maximum SINR can be achieved when the weight vectors \(w = R^{-1}(H_Dw)^*\) given that \(w = V_{\text{max}}\).

III. ERROR PROBABILITY ESTIMATION

The combined signal in Eq. (7) as

\[\tilde{c}_t = (a_0 + j b_0) c + (\xi + j \eta) + \omega_t\]  \hspace{1cm} (13)

where \(c = w^H H w\), which is equal to \(\lambda_{\max}\) is the largest eigenvalue of the matrix \(H^H H\) for MRT and \(H^H H\) for OT. With defining \(g_{i,n} = g(nT_s + \tau_i)\), the combined ISI in the in-phase rail due to total CCI can be denoted by

\[\xi = \sum_{i=1}^{L} \left( \sum_{n=1}^{N} a_{i,n} p_{i,n} - \sum_{n=1}^{N} b_{i,n} q_{i,n} \right)\]  \hspace{1cm} (14)

where we define the sampled pulse response of the \(i\)th CCI source as

\[p_{i,n} = \sum_{m=1}^{r} \lambda_{i,m} (w_{i,n}^T \cos \theta_{i,m} - w_{i,n}^T \sin \theta_{i,m}) g_{i,n}\]

\[q_{i,n} = \sum_{m=1}^{r} \lambda_{i,m} (w_{i,n}^T \sin \theta_{i,m} + w_{i,n}^T \cos \theta_{i,m}) g_{i,n}\]  \hspace{1cm} (15)

with \(h_{i,m} = \lambda_{i,m} e^{j \phi_{i,m}} = \lambda_{i,m} (\cos \theta_{i,m} + j \sin \theta_{i,m})\) and \(w^T = w_{i,n}^T + j w_{i,n}^T\). The ISI corresponding to the quadrature channel is denoted by \(\eta\). With a slight change in indexing the signal, we denote above pulse responses in the in-phase and quadrature channels, respectively, as

\[\xi = \sum_{i=1}^{L} \left( \sum_{n=1}^{N} a_{i,n} p_{i,n} - \sum_{n=1}^{N} b_{i,n} q_{i,n} \right)\]  \hspace{1cm} (16)

The \(m\)th weighted discrete-time noise is expressed as \(\omega_{m,i} = w_{m,i} v_{n,i}(T_s)\). The power spectra of the filtered noise \(v_{n,i}(t)\) is \(v_{m,i}(T_s) = F G(f)\) and hence resulting in the output power (variance) \(\sigma_{\omega_m}^2 = (w_{m,i}^T v_{n,i}(T_s))^2 N_0\), where \(G(f)\) has a raised cosine spectral characteristic. Since the noise is uncorrelated between diversity branches, the variance of the combined output noise, \(\omega\), is expressed as

\[\sigma_{\omega}^2 = N_0 \sum_{m=1}^{r} \sigma_{\omega_m}^2 = N_0 \sum_{m=1}^{r} (w_{m,i}^T v_{n,i}(T_s))^2.\]  \hspace{1cm} (17)

We define \(\omega = \omega_{m,i} + j \omega_{m,i}\), where \(\omega_{m,i}\) and \(\omega_{m,i}\) have equal variance, \(\sigma^2 = \sigma_{\omega}^2 / 2\). Since the distribution density functions of quantities \(\xi\) and \(\eta\) are symmetric to zero and are identical, it has been shown that the average symbol error probability \(P_M\) can be bounded tightly by

\[P_M = 2E[P, (\xi)] = 2 \left(1 - \frac{1}{\sqrt{P}}\right) \text{erfc} \left(\frac{g_0 + \xi}{\sqrt{2}\sigma}\right).\]  \hspace{1cm} (18)

The average bit error rate (BER) \(P_e = P_u/2\) for 4-QAM. Because \(\xi\) is a random variable whose distribution is not known explicitly, the evaluation of \(E[g(\xi)]\) is performed by computing the conditional error probability of each of all possible sequences of CCI, and then averaging over all those sequences [15][16]. We anticipate that the exhaustive semi-analytic method in Eq. (18) and direct Monte Carlo are inapplicable for fading cases in which more than, at least, 10,000,000 runs for error probability up to the order of \(10^{-6}\). 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 \(\xi\), up to an order that depends on the accuracy required.

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

\[E[g(\xi)] = \int_{-\infty}^{\infty} g(x)f(x)dx = \sum_{i=1}^{N} w_i g(x_i),\]  \hspace{1cm} (19)

a linear combination of values of the function \(g(\cdot)\), where
\(f(x)\) denotes the probability function of the random variable \(\xi\) with range \([a, b]\). The weights (or coefficients) \(w_i\), and the abscissas \(x_i\), \(i = 1, 2, \ldots, N\) can be calculated from the knowledge of the first \(2N+1\) moments of \(\xi\). A recursive algorithm which can be used to determine the moments of all order of \(\xi\) was discussed in [15]. We compute the average in (19) by means of the classic GQR’s suggested in [15] [16].

IV. SIMULATION RESULTS

We only exhibit the simulation results of 4-QAM with the rolloff factor \(\beta = 0.5\). A single dominant CCI and six interferers are considered individually. The average value of SIR is set to 10dB for simulation. In our test, the GQR method can generate the same results by using the analytic method in Eq. (18), when the channel is fixed. We first consider the performance of Maximum-ratio-transmission (MRT), when the precise CCI and Gaussian noise-like CCI models are employed. When all channels are unfaded, Fig. 2 show plots of the bit error probability (BER) versus average \(S/N\) per diversity branch. In general, for a given average \(S/N\), the transmit power in each of antennas is smaller for \(T > R\), whereas the total combined noise power at the receiver is higher for \(T < R\). Therefore, the effects of these two factors compensate for each other which makes the performance of bit error rate (BER) is symmetric in \(T\) and \(R\) in the absence of CCI, as the results discussed in [10].

It is observed that the results obtained by using precise interference model are considerably better than that obtained by using the Gaussian model. Those curves appear different for \(L = 1\), but they become close when \(L = 6\) or \(R\) is higher. Based on a central limit theorem, by increasing the number of interference and number of receiver antennas, the precise interference model can approach to the Gaussian model. Unlike the Gaussian CCI case, we can see that using the precise CCI model results in slightly better performance for \(T > R\). This is attributed to the fact that fewer interfering signals received by antennas cause a smaller degradation than Gaussian-like CCI. On the other hand, the curves of BER when the Gaussian CCI and precise CCI become close for \(T = R\) system, since CCI received by several antennas can be approached by Gaussian noise-like CCI.

When both the desired signal and CCI are subject to fading, the results are exhibited in Fig. 3. The average BER becomes very high, since fading effects increase the chance of taking on a lower instantaneous SIR. The curves of the Gaussian CCI and the precise CCI appear different with the increase of the transmit and receive antennas, which reduces the fading effect of the desired signal. The performance is slightly better for \(T > R\), (e.g. \(T = 3\) \(R = 2\)) when precise CCI model is used due to the CCI characteristics. However, similar to the case with fading CCI, the performance is better for \(T < R\) (e.g. \(T = 2\) \(R = 3\)) when Gaussian CCI model is used. Therefore, the improvement provided by the precise CCI model is evident when \(T > R\). Because of fading CCI, the performance with \(L = 1\) is worse than that with \(L = 6\). Moreover, we note that unlike the nonfading case, the BER with \((T, R) = (2, 2)\) is the lowest compared to \((T, R) = (3, 1)\) or \((1, 3)\).

Next, we consider the OT scheme and compare its results with the MRT scheme in the MIMO system. The number of receiver antennas must be greater than two in order to cancel CCI. For the nonfading case, Fig. 4 shows that OT performs worse than MRT due to significant noise enhancement under certain channel conditions of CCI, while the MIMO intends to cancel CCI. We find that the OT scheme is unable to show the superiority over the MRT scheme for higher value of SIR (e.g. \(SIR > 7\)dB), particularly when \(T > 1\). The use of \(R = 3\) can improve the raised curve of BER for OT. In fact, the maximum SINR is unable to guarantee the minimum BER, if the CCI is not Gaussian distributed. The joint antenna weights, which are derived for SINR maximization, is capable of minimizing the total power of interference and noise, while the power of CCI is reduced and the power of noise is enlarged. As a result, the BER becomes relatively high, since CCI causes much less impairment than the Gaussian noise given the same power. Contrarily, the MRT scheme provides good CCI elimination and achieve satisfied performance in this nonfading case.

When the desired signal and CCI are subject to fading, the probability of low instantaneous SIR is considerably increased, and then OT can demonstrate its superiority in cancelling CCI at higher SIR. Fig. 5 shows that OT can improve the performance significantly and outperform the MRT for \(L = 1\) at high SNR. Due to the noise enhancement effect, the performance of OT with two antennas is much worse than that with three antennas, given the a fixed number of antenna elements (e.g. \(T + R = 4\)) between the transmitter and receiver. Unlike the MRT case, the performance of OT with \(T = 2\), \(R = 2\) is worse than that with \(T = 1\), \(R = 3\) in the case of \(L = 1\) due to noise enhancement. However, the performance with \(T = 2\), \(R = 2\) is still better in the case of \(L = 6\), since CCI cannot be eliminated and has a similar behavior to the MRT case.

V. CONCLUSION

The results of this study are expected to lead to a better understanding of the effects of interference, and then to optimize spectrum reuse and coverage in MIMO systems. We successfully apply the Gaussian Quadrature rule (GQR) to calculate the accurate probability of error in the presence of ISI caused by cochannel interferers in the MIMO system based MRT and OT schemes. The simulation results show that the Gaussian interference model will become inaccurate for high order of antennas in the MRT based MIMO system, when the desired signal and CCI suffer Rayleigh fading. Moreover, the optimal technique using OT can offer performance gain at high SNR under fading conditions when the number of interferers is smaller. However, the OT scheme, in general, does not provide significant performance improvement over the MRT scheme when the number of interferers is large.

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