Capacity and Performance of MIMO-BICM System With Soft-output MMSE OSIC Detector

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Abstract—In this paper, we investigate the capacity and performance of the soft-output minimum men square error (MMSE) ordering successive interference cancellation (OSIC) detector for multiple-input multiple-output (MIMO) bit-interleaved coded modulation (BICM) system. We consider two methods to perform interference cancellation, which are maximum *a posteriori* probability (MAP) hard decision interference cancellation (HDIC) and soft interference cancellation (SIC). A reduced-complexity implementation for all of them is also proposed. Furthermore, we derived link level capacity (LLC) under ideal fast-fading conditions. Simulation results show that the soft-output MMSE OSIC detector outperforms MMSE and all of them are strongly superior to conventional MMSE OSIC detectors.

I. INTRODUCTION

Recently, soft-output minimum mean square error (MMSE) ordering successive interference cancellation (OSIC) detector [1], [2] has been applied to coded multiple-input multiple-output (MIMO) system under block-fading channels. By taking decision errors of previously detected spatial data streams into account, this detector significantly outperforms the MMSE and conventional MMSE OSIC detector which simply assumes perfect interference cancellation. However, in the detector reported in [1] and [2], the same probability is assigned to those modulation constellation points, which have identical Euclidean distance to the hard decision point corresponding to the transmitted symbol. This simple processing limits the performance improvement of the detector.

On the other hand, it has been shown that MIMO bit-interleaved coded modulation (BICM) is suitable for fast-fading environments [3], [4]. In this paper, we investigate the link level capacity (LLC) and performance of soft-output MMSE OSIC detector for MIMO-BICM under fast-fading MIMO channels. Contrarily to the detector proposed in [1] and [2], we proposed to compute the residual interference cancellation error covariance matrix based on the *a posteriori* probability associated with each previously detected spatial data stream. We consider two interference cancellation methods. In the first one, the maximum *a posteriori* probability (MAP) hard decision of previously detected spatial data stream is cancelled out, which we called as hard decision interference cancellation (HDIC) throughout this paper. In the second one, we proposed to cancel out the soft decision (mean value) of previously detected spatial data stream, which we called as soft interference cancellation (SIC). We also provided a complexity-reducing implementation for all of them. Furthermore, we derive the LLC under ideal fast-fading conditions. Extensive simulation results show the proposed soft-output MMSE OSIC MIMO detector is superior to the existing scheme and much better than MMSE detector and conventional MMSE OSIC MIMO detector, in which perfect interference cancellation is assumed.

The remaining parts of this paper are organized as follows. The system model is described in Section II. The soft-output MMSE OSIC MIMO detection algorithm and reduced-complexity implementation are discussed in Section III. In Section IV, we derive the LLC of MMSE OSIC MIMO detector. Simulation results and discussion are presented in Section V. Finally, we conclude this paper in Section VI.

II. SYSTEM DESCRIPTIONS

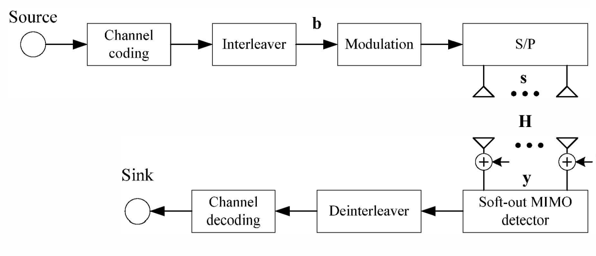

Figure 1. System model

A MIMO-BICM system with N_T transmitter antennas and N_R receiver antennas illustrated in Figure 1 can be modeled as follows

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$$\mathbf{y} = \mathbf{H}\mathbf{s} + \mathbf{n} \tag{1}$$

where $\mathbf{y} = [y_1, \cdots, y_{N_R}]^T \in \mathbb{C}^{N_R}$ is the $N_R \times 1$ received symbol vector. $\mathbf{H} = \left[\mathbf{h}_1, \cdots, \mathbf{h}_{N_T}\right] = \left[h_{i,j}\right]_{N_R \times N_T}$ is the $N_R \times N_T$ MIMO channel matrix whose element $h_{i,j}$ represents complex fading coefficient between the jth transmitter antenna and the ith receiver antenna. In this paper, we adopt the following spatially correlated MIMO channel model [5]

$$\mathbf{H} = \mathbf{H}_w \mathbf{R}_t^{1/2} \tag{2}$$

where the elements of $\mathbf{H}_w$ are independent zero mean circularly symmetric complex Gaussian (ZMCSCG) random variables with unit variance and $\left[\mathbf{R}_t\right]_{N_T \times N_T}$ is the transmitter correlation matrix. In this paper, $h_{i,j}$ is assumed to be flat fast-fading, which varies randomly from one transmitted symbol vector to another. It is also assumed to be perfectly known at the receiver, i.e., perfect channel estimation, but unknown at the transmitter. Meanwhile, $\mathbf{n}$ is additive white Gaussian noise (AWGN) vector whose elements are modeled as samples of independent complex Gaussian random variables with mean zero and variance $\sigma_n^2/2$ per dimension. $\mathbf{s} = [s_1, \cdots, s_{N_T}]^T \in \mathbb{C}^{N_T}$ is $N_T \times 1$ transmitted symbols vector whose element $s_i = \mathrm{map}\left(\mathbf{b}_i\right) \in \mathcal{A}$ is taken from complex modulation constellation $\mathcal{A}$ of size M (such as M-PSK or M-QAM) and $E\left[\mathbf{s}\mathbf{s}^H\right] = E_S \mathbf{I}_{N_T}$. The function $\mathrm{map}\left(\mathbf{b}_i\right)$ denotes the Gray mapping from channel coded bits vector $\mathbf{b}_i = \left[b_i^1, \cdots, b_i^{\log_2(M)}\right]^T$ to one of the complex modulation symbols belonging to $\mathcal{A}$.

III. SOFT-OUTPUT MMSE OSIC MIMO DETECTOR

A. Basic principle

Let $k_i \in \{1, 2, \cdots, N_T\}$ be the index of the ith data stream to be detected according to the maximal post-detection signal-to-noise ratio (SNR) ordering rule [6], we have the corresponding interference-cancelled received signal vector $\mathbf{y}_{k_i}$ as

$$\mathbf{y}_{k_i} = \mathbf{h}_{k_i} s_{k_i} + \underbrace{\sum_{j=k_{i+1}}^{k_{N_T}} \mathbf{h}_j s_j}_{\mathbf{I}_U} + \underbrace{\sum_{j=k_1}^{k_{i-1}} h_j \left(s_j - \tilde{s}_j\right)}_{\mathbf{I}_D} + \mathbf{n} \tag{3}$$

where $\mathbf{I}_U$ is the interferences from undetected data stream and $\mathbf{I}_D$ is the interference due to the decision error of previously detected data streams, and $\tilde{s}_j$ represents the estimation of s_j. The MMSE spatial filter $\mathbf{W}_i$ that is applied to suppress $\mathbf{I}_U$, $\mathbf{I}_D$ and $\mathbf{n}$ is given as

$$\mathbf{W}_i = \left[\mathbf{H}_{k_i:k_{N_T}} \mathbf{H}_{k_i:k_{N_T}}^H + \frac{1}{E_S}\mathbf{R}_{\mathbf{I}_D} + \frac{\sigma_n^2}{E_S}\mathbf{I}_{N_R}\right]^{-1} \mathbf{h}_{k_i} \tag{4}$$

where $\mathbf{H}_{k_i:k_{N_T}} = \left[\mathbf{h}_{k_i}, \mathbf{h}_{k_{i+1}}, \cdots, \mathbf{h}_{k_{N_T}}\right]$ is the sub-matrix of $\mathbf{H}$ by deleting the $k_1, k_2, \cdots, k_{i-1}$ columns. The covariance matrix $\mathbf{R}_{\mathbf{I}_D}$ is given as

$$\mathbf{R}_{\mathbf{I}_D} = E\left[\mathbf{I}_D \mathbf{I}_D^H\right] = \mathbf{H}_{k_1:k_{i-1}} \mathbf{Q}_i \mathbf{H}_{k_1:k_{i-1}}^H \tag{5}$$

where $\mathbf{Q}_i = \left[q_{u,v}\right]_{(i-1)\times(i-1)}$ is the residual interference cancellation error covariance matrix whose element is given as

$$q_{u,v} = \begin{cases} E\left[\left|s_{k_u} - \breve{s}_{k_u}\right|^2 \middle| \breve{s}_{k_u}\right], u = v \\ E\left[\left(s_{k_u} - \breve{s}_{k_u}\right)\middle| \breve{s}_{k_u}\right] E\left[\left(s_{k_v} - \breve{s}_{k_v}\right)^* \middle| \breve{s}_{k_v}\right], u \neq v \end{cases} \tag{6}$$

Note that $q_{u,v}^i = q_{u,v}^{i+1}, i \in \{2, 3, \cdots, N_T - 1\}$ and $u, v \in \{1, 2, \cdots, i-1\}$. Thus, the approximate equivalent Gaussian channel expression of the estimate of s_{k_i} can be given as [1], [2]

$$\tilde{s}_{k_i} = \mu_{k_i} s_{k_i} + \eta_{k_i} \tag{7}$$

where $\mu_{k_i} = \mathbf{W}_i^H \mathbf{h}_{k_i}$ and $\eta_{k_i} \sim \mathcal{CN}\left(0, \sigma_{\eta_{k_i}}^2 = \left(\mu_{k_i} - \mu_{k_i}^2\right) E_S\right)$. Then, the LLR of each coded bit can be obtained from (7) as in [3].

B. Computation of the element in $\mathbf{Q}_i$

In the scheme proposed in [1]and [2], the element of $\mathbf{Q}_i$ is computed through symbol error probability estimation and modulation constellation dimension decomposition. However, the same probability is assigned to those modulation constellation points, which have identical Euclidean distance to the hard decision point of the transmitted symbol. This simple processing limits the performance improvement of the detector. Here, we proposed another method which computes the residual interference cancellation error covariance matrix based on the *a posteriori* probability associated with each previously detected spatial data stream

Considering the approximate equivalent Gaussian expression of (7), the *a posteriori* probability of $\alpha_l \in \mathcal{A}$ conditioned on $\tilde{s}_{k_i}$ and μ_{k_i} is given as

$$p\left(s_{k_i} = \alpha_l \middle| \tilde{s}_{k_i}, \mu_{k_i}\right) = \frac{p\left(\tilde{s}_{k_i} \middle| s_{k_i} = \alpha_l, \mu_{k_i}\right) p\left(s_{k_i} = \alpha_l\right)}{p\left(\tilde{s}_{k_i}\right)} \tag{8}$$

Assuming that each modulation constellation symbol is transmitted with equal probability, it is straightforward that

$$\begin{aligned} p\left(s_{k_i}=\alpha_l\middle|\tilde{s}_{k_i},\mu_{k_i}\right) &\propto p\left(\tilde{s}_{k_i}\middle|s_{k_i}=\alpha_l,\mu_{k_i}\right) \\ &=\frac{1}{\pi\sigma_{\eta_{k_i}}^2}\exp\left(-\left|\tilde{s}_{k_i}-\mu_{k_i}\alpha_l\right|^2\Big/\sigma_{\eta_{k_i}}^2\right) \end{aligned} \tag{9}$$

While the constraint $\sum_{\alpha_l\in\mathcal{A}} p\left(s_{k_i}=\alpha_l\middle|\tilde{s}_{k_i},\mu_{k_i}\right)=1$ stands, the *a posteriori* probability $p\left(s_{k_i}=\alpha_l\middle|\tilde{s}_{k_i},\mu_{k_i}\right),\alpha_l\in\mathcal{A}$ can therefore be computed.

1) ML HDIC

For ML HDIC, the estimation of $\breve{s}_{k_i}$ can be obtained as

$$\breve{s}_{k_i}=\hat{s}_{k_i}=\underset{\alpha_l\in\mathcal{A}}{\arg\max}\, p\left(s_{k_i}=\alpha_l\middle|\tilde{s}_{k_i},\mu_{k_i}\right) \tag{10}$$

And the element of $\mathbf{Q}_i$ can be computed through the following relationships

$$\begin{aligned} E\left[\left|s_{k_u}-\breve{s}_{k_u}\right|^2\middle|\tilde{s}_{k_u}\right] &= \sum_{\alpha_l\in\mathcal{A}}\left|\alpha_l-\breve{s}_{k_u}\right|^2 p\left(s_{k_i}=\alpha_l\middle|\tilde{s}_{k_i},\mu_{k_i}\right) \\ E\left[\left(s_{k_u}-\breve{s}_{k_u}\right)\middle|\tilde{s}_{k_u}\right] &= \sum_{\alpha_l\in\mathcal{A}}\left(\alpha_l-\breve{s}_{k_u}\right) p\left(s_{k_i}=\alpha_l\middle|\tilde{s}_{k_i},\mu_{k_i}\right) \\ E\left[\left(s_{k_u}-\breve{s}_{k_u}\right)^*\middle|\tilde{s}_{k_u}\right] &= \sum_{\alpha_l\in\mathcal{A}}\left(\alpha_l-\breve{s}_{k_u}\right)^* p\left(s_{k_i}=\alpha_l\middle|\tilde{s}_{k_i},\mu_{k_i}\right) \end{aligned} \tag{11}$$

2) SIC

Contrarily to HDIC, the soft estimation (mean value) of the previously detected symbol is cancelled out in SIC based MMSE OSIC MIMO detector, i.e.,

$$\breve{s}_{k_i}=\bar{s}_{k_i}=\sum_{\alpha_l\in\mathcal{A}}\alpha_l p\left(s_{k_i}=\alpha_l\middle|\tilde{s}_{k_i},\mu_{k_i}\right) \tag{12}$$

Therefore, $E\left[\left(s_{k_u}-\breve{s}_{k_u}\right)\middle|\tilde{s}_{k_u}\right]=0$ for SIC and $\mathbf{Q}_i$ is a strictly diagonal matrix.

C. *Complexity-reducing implemention*

Clearly, most the of computation complexity are resulted from the matrix inverse of MMSE filter in (4). An efficient method to reduce the complexity of matrix inverse is to adopt the matrix inverse lemma [7]. In order to adopt this lemma, we have to ignore the off-diagonal elements of the matrix $\mathbf{Q}_i$ in HDIC based soft-output MMSE OSIC MIMO detector. Fortunately, simulation results show that this approximation will introduce negligible performance loss (less that 0.3dB). On the other hand, because the matrix $\mathbf{Q}_i$ is diagonal matrix, there are not any approximations needed in the SIC based soft-output MMSE OSIC MIMO detector. Therefore, we can rewrite (4) as

$$\begin{aligned} \mathbf{W}_i &= \Bigg[\underbrace{\sum_{m=k_i}^{k_{N_T}}\mathbf{h}_m\mathbf{h}_m^H+\frac{1}{E_S}\sum_{m=k_1}^{k_{i-1}}\mathbf{h}_m\mathbf{h}_m^H q_{m,m}^i+\frac{\sigma_n^2}{E_S}\mathbf{I}_{N_R}}_{\mathbf{A}_i}\Bigg]^{-1}\mathbf{h}_{k_i} \\ &= \mathbf{A}_i^{-1}\mathbf{h}_{k_i} \end{aligned} \tag{13}$$

As mentioned in the previous Section, we have $q_{j,j}^i=q_{j,j}^{i+1}$. Therefore, it follows that,

$$\begin{aligned} \mathbf{A}_i &= \mathbf{A}_{i-1}+\mathbf{h}_{k_i}\mathbf{h}_{k_i}^H\left(q_{i,i}^i/E_S-1\right), i=2,\cdots,N_T \\ \mathbf{A}_1 &= \mathbf{H}\mathbf{H}^H+\frac{\sigma_n^2}{E_S}\mathbf{I}_{N_R} \end{aligned} \tag{14}$$

By applying the matrix inverse lemma [7], $\mathbf{A}_i^{-1}$ can be given as,

$$\mathbf{A}_i^{-1}=\mathbf{A}_{i-1}^{-1}-\frac{\mathbf{A}_{i-1}^{-1}\mathbf{h}_{k_i}\mathbf{h}_{k_i}^H\mathbf{A}_{i-1}^{-1}}{E_S\Big/\left(q_{i,i}^i-E_S\right)+\mathbf{h}_{k_i}^H\mathbf{A}_{i-1}^{-1}\mathbf{h}_{k_i}}, i=2,\cdots,N_T \tag{15}$$

As $\mathbf{h}_{k_i}^H\mathbf{A}_{i-1}^{-1}=\left(\mathbf{A}_{i-1}^{-1}\mathbf{h}_{k_i}\right)^H$, only the computation of $\mathbf{A}_{i-1}^{-1}\mathbf{h}_{k_i}$ is needed. Consequently, only one direct matrix inverse is required in the detection of each transmitted symbols vector. With this approximation, $3N_R^2+N_R+1$ multiplications/divisions and $2N_R^2+1$ additions/subtractions are needed to compute $\mathbf{A}_i^{-1}, i=2,\cdots,N_T$. On the other hand, a numerically more stable way to compute the matrix inverse is via singular value decomposition (SVD) [8], which needs $13N_R^3+N_R^2$ multiplications and $13N_R^3-N_R^2$ additions. Therefore, the complexity can be reduced drastically.

IV. LINK-LEVEL CAPACITY

Link-level capacity (LLC) is defined as the mutual information (MI) of each bit level [3], [4]. According to (7), the conditional MI for the λ-th bit level is given as

$$\begin{aligned} I\left(b^\lambda;\tilde{s}_{k_i}\middle|\mu_{k_i}\right) &= H\left(b^\lambda\right)-H\left(b^\lambda\middle|\tilde{s}_{k_i},\mu_{k_i}\right) \\ &= 1-E_{b^\lambda|\tilde{s}_{k_i},\mu_{k_i}}\left(\log_2\frac{\sum_{\tilde{b}^\lambda\in\{0,1\}}p\left(\tilde{s}_{k_i}\middle|\tilde{b}^\lambda,\mu_{k_i}\right)}{p\left(\tilde{s}_{k_i}\middle|b^\lambda,\mu_{k_i}\right)}\right) \end{aligned} \tag{16}$$

where $H(\cdot)$ is the entropy function [9] and

$$p\left(\tilde{s}_{k_i}\middle|b^\lambda,\mu_{k_i}\right)=\frac{1}{\left|\mathcal{A}_b^{i,k}\right|}\sum_{\alpha_l\in\mathcal{A}_b^{i,k}}p\left(\tilde{s}_{k_i}\middle|\alpha_l,\mu_{k_i}\right) \tag{17}$$

where $\mathcal{A}_b^{i,k}$ denotes the modulation constellation points set which has the ith bit equal to $b\in\{0,1\}$, and $\left|\mathcal{A}_b^{i,k}\right|$ denotes

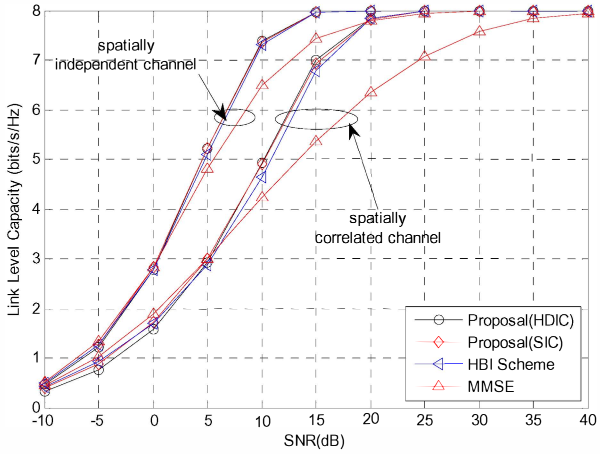

Figure 2. LLC versus average SNR per Rx antenna for QPSK

the cardinality of $\mathcal{A}_b^{i,k}$.

Assuming ideal interleaving, the channel can be considered ergodic and the LLC is obtained by averaging the MI for each bit level over channel [3], [4], which is given by

$$I_{\mathrm{BICM\_OSIC}} = \sum_{k=1}^{N_T} \sum_{i=1}^{\log_2(M)} E_{\mu_{k_i}} \left(I\left(b^{\lambda}; \tilde{s}_{k_i} \middle| \mu_{k_i}\right) \right)$$
$$= N_T \log_2(M)$$
$$- \sum_{k=1}^{N_T} \sum_{i=1}^{\log_2(M)} E_{\mu_{k_i}} \left(E_{b^{\lambda}|\tilde{s}_{k_i},\mu_{k_i}} \left(\log_2 \frac{\sum_{\tilde{b}^{\lambda}\in\{0,1\}} p\left(\tilde{s}_{k_i} \middle| \tilde{b}^{\lambda}, \mu_{k_i}\right)}{p\left(\tilde{s}_{k_i} \middle| b^{\lambda}, \mu_{k_i}\right)} \right) \right) \tag{18}$$

It is worthy of noting that the formula of (18) can not be applied to conventional MMSE OSIC detector which assumes perfect interference cancellation. As the interference cancellation error is neglected, The Gaussian approximation of MMSE filter output will lead to unreliable *a posteriori* probability. Then the corresponding MI is also unreliable. Moreover, it is very difficult to obtain accurate probability distribution about the output of conventional MMSE OSIC detector. Therefore, we don't evaluate the LLC with respect to conventional MMSE OSIC detector in this paper.

V. NUMERICAL RESULTS

In this section, we present simulation results to compare the LLC and bi-error rate (BER) performance of the proposed detector with HDIC and SIC. For the sake of comparison, we also provide the LLC and BER of the conventional hard decision scheme, the MMSE based approximation as in [10] and the exiting scheme proposed in [1] and [2], which we denoted as HBI scheme in our simulation results.

In our simulations, we adopt the 1/2 rate binary convolutional code (CC) with polynomials (133, 171) in octal notation and the coded bits are interleaved with a block interleaver of depth 100. At the receiver, the soft-in APP decoder [12] implemented in Matlab™ Simulink® is used. The QPSK and 16-QAM modulations are used in the simulations and the simulations are performed for $N_T = N_R = 4$ MIMO systems. The applied transmitter correlation matrix is given as following

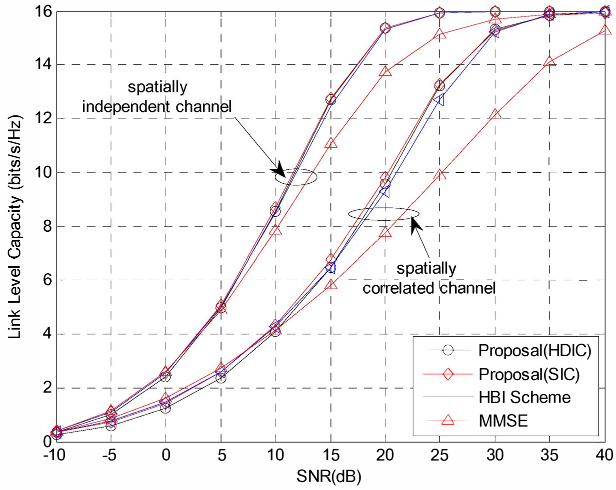

Figure 3. LLC versus average SNR per Rx antenna for 16-QAM

$$\mathbf{R}_t = \begin{bmatrix} 1 & 0.4290+0.7766j & -0.3642+0.5490j & -0.4527-0.0015j \\ 0.4290-0.7766j & 1 & 0.4290+0.7766j & -0.3642+0.5490j \\ -0.3642-0.5490j & 0.4290-0.7766j & 1 & 0.4290+0.7766j \\ -0.4527+0.0015j & -0.3642-0.5490j & 0.4290-0.7766j & 1 \end{bmatrix}$$

which is taken from [11].

A. LLC evaluation

The LLC results of different detectors are obtained through Monte-Carlo integration as in [3] and [4]. Fig. 2 and Fig. 3 present the LLC results of different detectors under spatially independent and correlated MIMO channels for QPSK and 16-QAM respectively. As the reason mentioned in previous Section, we have not provide the LLC of conventional MMSE OSIC MIMO detector. It can be seen that the achievable LLC of the proposed detector is larger than that of HBI scheme and MMSE detector, especially for spatially correlated MIMO channel condition. For our proposed detector, HDIC and SIC can achieve nearly the same LLC for all considered modulation constellations and channel conditions.

B. BER performance

To confirm the performance prediction provided by LLC comparison, we provide the BER performance comparison in this Section. Fig. 4 and Fig. 5 show the BER performance of different detectors under spatially independent and correlated MIMO channels for QPSK and 16-QAM modulation respectively. As shown in these figures, our proposed detector outperforms all the existing schemes under all

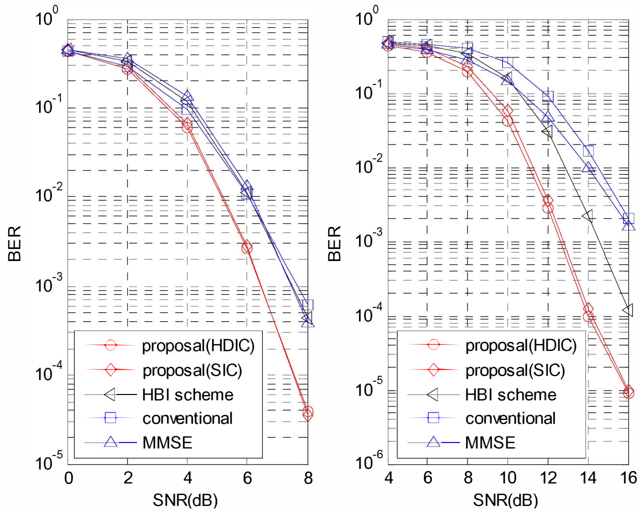

Figure 4. Performance comparison of different detectors (QPSK, left: independent channel, right: correlated channel)

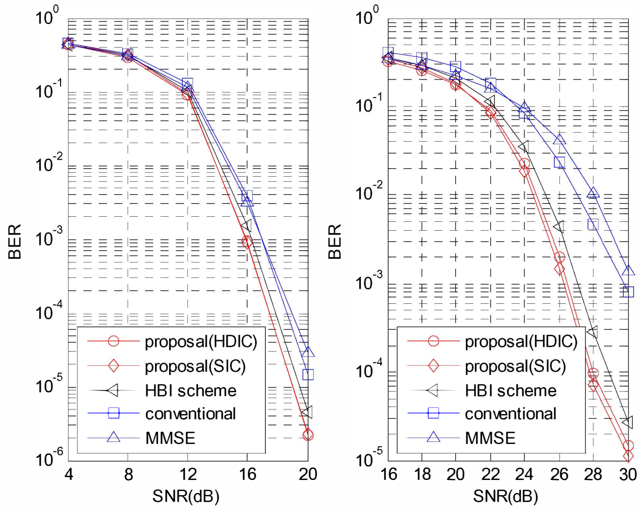

Figure 5. Performance comparison of different detectors (16-QAM, left: indpendent channel, right: correlated channel)

channel conditions. Compared with HBI scheme, the performance gain of our proposal is more outstanding for spatially correlated MIMO channel condition and QPSK modulation. The reason is that the error introduced by the uniform probability assignment of HBI scheme becomes small when the size of modulation constellation increases. Even though, the gain is about 1dB for 16-QAM in spatially independent channel when the BER equals to 10^{-5}. Furthermore, for our proposed detector, the two schemes have not important performance difference. Considering the maximal value search is needed for HDIC scheme, which leads to extra complexity, the performance of SIC is slightly better than that of HDIC. It is reasonable to choose SIC as a potential candidate for practical implementation.

VI. Conclusion

In this paper, we investigate the LLC and performance of MIMO-BICM system with soft-output MMSE OSIC detector. We propose to compute the residual interference cancellation error covariance matrix based on the *a posteriori* probability associated with each previously detected spatial data stream. We consider both HDIC and SIC and provide reduced-complexity implementation. Furthermore, we derive the LLC of soft-output MMSE OSIC detector under ideal fast-fading conditions. Simulation results show the advantages of our proposed detector.

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