Mean Mutual Information Per Coded Bit based Precoding in MIMO-OFDM Systems

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***Abstract*—This work proposes a per-subband multiple input multiple output (MIMO) precoder selection technique for point-to-point MIMO orthogonal frequency division multiplexing (OFDM) based bit interleave coded modulation (BICM) systems with the soft-output minimum mean square error (MMSE) receiver. Given a pre-designed precoder codebook, the codeword/precoder that maximizes the mean of the mutual information per coded bit (MMIB) on all subcarriers within a subband is selected. The main advantages of this technique are the following: i) the precoder selection metric is explicitly related to BICM performance, thus it outperforms the previously proposed precoding techniques; ii) with commonly used unitary precoding codebooks, this technique works for an arbitrary number of transmit streams unlike the minimum singular value based method which does not work when the number of input streams is the same as the number of transmit antennas; iii) when multiple packets are transmitted and one precoder is used for these transmitted packets, an algorithm that combines the MMIB of each packet is proposed using an upper bound on the average packet error rate.**

I. Introduction

MIMO-OFDM is a spectrum-efficient technology and has been incorporated into many of the wireless standards such as IEEE 802.11, WiMaX and 3GPP LTE. The packaged technologies in standardized MIMO-OFDM systems also include error control coding (ECC) and BICM. In general, ECC is required to achieve frequency diversity across the OFDM subcarriers. Also ECC can be used to achieve the spatial diversity even for spatial multiplexing when the coded bits are transmitted across the transmit antennas (this is the so called vertical encoding). ECC thus helps to improve the link robustness against fading, interference and noise. To bridge ECC and modulation, a popular paradigm, i.e., BICM is used, which randomizes the encoded bit sequence by interleaving before modulation [1].

In this paper, point-to-point spatial multiplexing is considered, where the transmitter and the receiver both have multiple antennas. When a representation of the MIMO channel quality information (CQI) is available at the transmitter of a MIMO-OFDM system, precoding can be applied at the transmitter. A typical form of the precoding techniques is linear precoding, which applies a matrix precoder to the spatial streams. The standard procedure to determine the precoder for MIMO systems, however, is to draw a codeword at the receiver from a pre-computed codebook (available at both the transmitter and the receiver), then feed back the codeword index to the transmitter [2]. This is also called limited feedback based MIMO precoding. Techniques on how to determine the precoder for spatial multiplexing has been proposed in literature, e.g., [3] [4].

The soft output MMSE receiver is considered in this paper [5]. Though it gives suboptimal performance compared to the maximum likelihood (ML) receiver, it has lower implementation complexity. Given linear receivers and spatial multiplexing, prior work proposed minimum singular value based precoding [3], maximizing minimum capacity per packet precoding (maximizing the minimum information theoretic capacity of each spatially transmitted packet), and minimum signal to noise ratio based precoding [4]. Though these methods are primarily used for uncoded narrow band MIMO systems, extension to MIMO-OFDM systems is somewhat straightforward. Simply, we can select and feed back a different precoder for each subcarrier based on these methods. However, the signaling overhead for per-subcarrier based feedback is quite significant if the channel coherence time is short.

A technique to reduce the signaling overhead is introduced in [6], in which an interpolation-based precoding technique is proposed and it significantly reduces the signaling overhead. However, the main drawback of interpolation-based precoding is the increase in receiver complexity caused by the interpolation operation. As a simpler alternative, subband-based precoding has been adopted in the 3GPP LTE standards. However, how to determine a common precoder within each subband becomes a question. A simple method is to use the existing precoder selection method operating on the average channel on this subband. The main drawback of this method is that it does not consider the overall BICM error rate performance.

Recent research progress on MIMO-OFDM BICM systems indicate that link quality can be represented by the mean mutual information of all encoded bits over an equivalent log-likelihood ratio (LLR) channel. This link quality metric has been used in the link adaptation context to determine the modulation order and coding rate [7] [8]. For fixed rate systems (e.g. voice streaming), we propose to use the mean mutual information per coded bit as the metric to determine the precoding codeword. Our mean mutual information per coded bit (MMIB) based precoder selection technique is compared with the minimum singular value (MSV) based precoding and maximizing minimum capacity (MMC) per packet precoding. Unlike the MSV precoding technique, our proposed method

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can be applied to the scenario in which the number of transmitted streams is the same as the number of transmit antennas, with unitary precoding codebooks. Further, when multiple packets are transmitted, a rule that combines the MMIB of each packet is proposed based on a packet error rate upper bound. Performance gain of the MMIB precoding over MSV and MMC has been observed through extensive simulations.

The mean mutual information per coded bit based precoding technique with an MMSE receiver proposed in this paper can be extended to precoding systems with the ML receiver. The mean mutual information per coded bit analysis is available for the scenario of two input streams for non-precoded MIMO systems [9]. Due to space constraints, we do not elaborate on this extension and the related performance evaluation.

The standard matrix notations are used in this paper, where $(\cdot)^T$ denotes transpose and $(\cdot)^H$ denotes conjugate transpose.

II. System Model

In this section, the system model of MIMO-OFDM BICM is presented. We also give an introduction to mean mutual information over an LLR channel.

A. BICM Resource Allocation Model

We consider a point-to-point MIMO-OFDM link where the transmitter has M_t transmit antennas and the receiver has M_r receive antennas. The total number of OFDM subcarriers is denoted by N_T. There are N_{pac} packets to be transmitted, and these are divided into M_s streams ($N_{pac} \leq M_s$). The m-th stream occupies $M_{s,m}$ spatial streams. The symbols in each of these spatial streams are then transmitted over N_{sc} subcarriers[1]. The m-th packet is thus transmitted over $N_{sc}M_{s,m}$ space-frequency points in the absence of precoding, and each sub-carrier carries M_s symbol streams. Spatial precoding can be applied to each sub-carrier on these M_s streams – this is the scenario considered in this paper. The above is a brief generalized description of the 3GPP LTE standards [10]. For example, when $M_s = 4$, two packets are transmitted and each of them takes two spatial streams. For simplicity, we assume that the QAM modulation orders used for all packets are the same and are fixed to be Q-QAM where $Q = 2, 4, 16, 64$. The set of QAM constellations is denoted by χ. We also assume the ECC rates for all packets are the same, i.e., R. When different packets have different modulation order and coding rate, the precoder design problem is more complicated, but is extendable from the formulation in Section III.

To simplify the presentation in this paper, the wireless channel is assumed to be constant over J_{ofdm} OFDM symbols and varies independently from block to block. We assume further that the m-th packet consists of B_m coded bits and it spans N_{symb} OFDM symbols, where it is assumed that $N_{symb} \leq J_{ofdm}$ for simplicity. Note that we assume that all packets use the same set of subcarriers and OFDM symbols. The transmitted bits for the m-th packet are denoted by $c_{m,v}$ ($v = 0, ..., B_m - 1$). The permutation in the subcarrier domain is denoted by π. Thus the p-th logical subcarrier ($0 \leq p \leq N_{sc} - 1$) is mapped to $\pi(p)$-th physical subcarrier. We assume that a subband based subcarrier allocation is used for each packet. This means that N_{sub} contiguous subcarriers are grouped together and a total of $N_{sb} = N_{sc}/N_{sub}$ subbands that may spread over the entire OFDM band are allocated to the packet. Finally, the precoder is selected from a codebook $\mathcal{C}$ with a cardinality denoted by $|\mathcal{C}|$. The codebook design is out of the scope of this work. We use the recommended codebook in 3GPP LTE in this paper [10].

[1] N_{sc} is always smaller than N_T because some sub-carriers are not used for carrying data.

B. Per Subcarrier Signal Model

The signal model on the k-th physical subcarrier can be written as

$$\mathbf{x}[k] = \mathbf{H}[k]\mathbf{F}[k]\mathbf{s}[k] + \mathbf{n}[k], \quad (1)$$

where $\mathbf{H}[k]$ of size $M_r \times M_t$ is the MIMO channel on the k-th subcarrier and $\mathbf{F}[k]$ of size $M_t \times M_s$ is the precoder on the k-th subcarrier. The signal $\mathbf{s}[k]$ of size $M_s \times 1$ is the transmitted QAM symbol vector. Due to BICM, the elements of $\mathbf{s}[k]$ are independently distributed with zero mean and unit variance. The noise vector $\mathbf{n}[k]$ has a dimension $M_r \times 1$. Its elements follow i.i.d. Gaussian distribution with zero mean and a variance of σ_n^2.

The receiver uses the soft output MMSE detector [5] [11]. Define a matrix $\mathbf{A}[k]$ of size $M_s \times M_r$ as the following

$$\mathbf{A}[k] = (\mathbf{I}_{M_s}\sigma_n^2 + \mathbf{F}[k]^H\mathbf{H}[k]^H\mathbf{H}[k]\mathbf{F}[k])^{-1}\mathbf{F}[k]^H\mathbf{H}[k]^H, \quad (2)$$

where $\mathbf{I}_{M_s}$ is the identity matrix of size $M_s \times M_s$.

The following operation is performed on each subcarrier

$$\mathbf{y}[k] = \mathbf{A}[k]\mathbf{x}[k]. \quad (3)$$

Define a matrix $\mathbf{R}[k] = (\sigma_n^{-2}\mathbf{F}[k]^H\mathbf{H}[k]^H\mathbf{H}[k]\mathbf{F}[k] + \mathbf{I}_{M_s})^{-1}$. For the i^{th} stream on the k-th subcarrier, the signal to noise and interference ratio can be written as

$$SINR_i[k] = \frac{1}{R_{i,i}[k]} - 1, \quad (4)$$

where $R_{i,i}[k]$ denotes the $(i,i)^{th}$ element of the matrix $\mathbf{R}[k]$ [12].

We denote the soft output of the MMSE detector for the u^{th} bit of the i^{th} stream on the k-th subcarrier by $\Lambda_{u,i}[k]$. The soft-output of the MMSE detector is given as follows [5]:

$$\begin{aligned}\Lambda_{u,i}[k] &= \log\Big(\sum_{a\in\chi_1^u} e^{-|y_i[k]/(1-R_{i,i}[k])-a|^2(\frac{1}{R_{i,i}[k]}-1)}\Big) \\ &- \log\Big(\sum_{a\in\chi_0^u} e^{-|y_i[k]/(1-R_{i,i}[k])-a|^2(\frac{1}{R_{i,i}[k]}-1)}\Big), \quad (5)\end{aligned}$$

where χ_1^u denotes the set of QAM constellation points with the u^{th} bit being 1 in its binary representation and χ_0^u denotes the set of QAM constellation points with the u^{th} bit being 0 in its binary representation. The quantity a denotes any point in the restricted constellations χ_1^u or χ_0^u. The quantity $y_i[k]$ denotes the i^{th} element of the vector $\mathbf{y}[k]$. This equation is the LogAPP (Logarithmic A Posteriori Probability) calculation of the soft outputs.

TABLE I

MEAN MUTUAL INFORMATION PER SYMBOL FOR BPSK, QPSK, 16QAM AND 64QAM

Modulation	$\mathcal{I}_Q^{symb}(SINR)$
BPSK	$J(\sqrt{8SINR})$
QPSK	$J(\sqrt{4SINR})$
16QAM	$\frac{1}{2}J(a_3\sqrt{SINR}) + \frac{1}{4}J(b_3\sqrt{SINR}) + \frac{1}{2}J(c_3\sqrt{SINR})$
64QAM	$\frac{1}{3}J(a_4\sqrt{SINR}) + \frac{1}{3}J(b_4\sqrt{SINR}) + \frac{1}{3}J(c_4\sqrt{SINR})$
For LogAPP demapping, $a_3 = 0.8$, $b_3 = 2.17$, $c_3 = 0.965$, $a_4 = 1.47$, $b_4 = 0.529$, $c_4 = 0.366$.	

C. Mean Mutual Information Over Log Likelihood Ratio Channel

The mutual information over log likelihood ratio channel is defined as the information theoretic mutual information between the coded bits ($c_{m,v}$) and the log likelihood ratio ($L(c_{m,v})$) extracted by the detector. The mutual information per coded bit for the m-th packet is given in [13] as the following

$$\mathcal{I}(c_{m,v}, L(c_{m,v})) = \frac{1}{2}\sum_{c_{m,v}=\{0,1\}} \int_{-\infty}^{+\infty} p_{LLR}(z|c_{m,v}) \cdot \log_2\left(\frac{2p_{LLR}(z|c_{m,v})}{p_{LLR}(z|c_{m,v}=0)+p_{LLR}(z|c_{m,v}=1)}\right) dz. \quad (6)$$

The mean mutual information per coded bit or symbol can be written as

$$\mathcal{I}_Q^{symb} = \frac{1}{\log_2(Q)} \sum_{v=1}^{\log_2(Q)} \mathcal{I}(c_{m,v}, L(c_{m,v})), \quad (7)$$

where Q denotes the size of the QAM constellation. The LLR per coded bit is Gaussian distributed with the mean of the PDF of LLR μ_{LLR} being half of the variance of the PDF of LLR σ_{LLR}^2. Therefore, we have [13]

$$\mu_{LLR} = \frac{\sigma_{LLR}^2}{2}. \quad (8)$$

When BPSK modulation is used, we have $\mu_{LLR} = 4SINR$. For BPSK modulation, the mutual information per coded bit can be written as

$$\begin{aligned} \mathcal{I}_{BPSK}^{symb}(SINR) &= 1 - \int_{-\infty}^{+\infty} \frac{1}{\sqrt{2\pi\sigma_{LLR}^2}} \\ &\cdot \exp\left(-\frac{|z-\sigma_{LLR}^2/2|^2}{2\sigma_{LLR}^2}\right) \\ &\cdot \log_2(1+\exp(-z))dz \\ &= J(\sigma_{LLR}) \\ &= J(\sqrt{8SINR}). \end{aligned} \quad (9)$$

Direct numerical integration for the mutual information is difficult. Therefore, numerical approximation has been approached as a means to calculate the mutual information. Based on [8], we can approximate the $J(\cdot)$ function as the following

$$J(x) = \begin{cases} a_1x^3 + b_1x^2 + c_1x \\ (0 < x < 1.6363), \\ 1 - \exp(a_2x^3 + b_2x^2 + c_2x + d_2) \\ (1.6363 \le x < \infty), \end{cases} \quad (10)$$

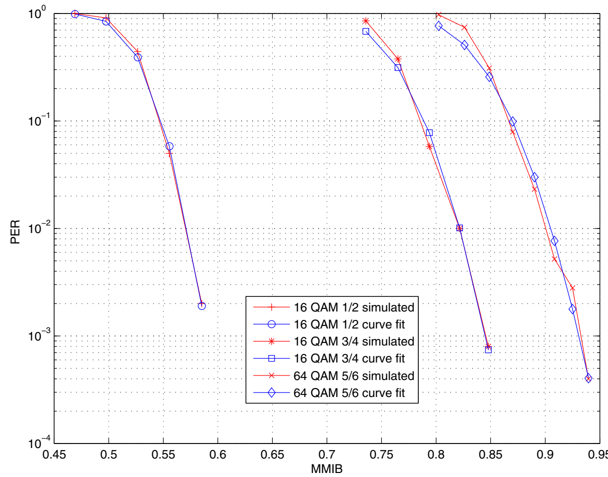

Fig. 1. Comparison of the simulated packet error rate versus the curve fitted packet error rate for different modulations and coding rates in AWGN channel using LTE Turbo code.

and the coefficients of the $J(\cdot)$ function are that $a_1 = -0.0421061$, $b_1 = 0.209252$, $c_1 = -0.00640081$, $a_2 = 0.00181491$, $b_2 = -0.142675$, $c_2 = -0.0822054$, $d_2 = 0.0549608$. For higher modulation order than BPSK, we have Table I that includes the results on mutual information per symbol [7].

Considering MIMO-OFDM modulation, for the m-th packet, the mean mutual information per coded bit is denoted by

$$\mathcal{I}_m^{Packet} = \frac{1}{M_{s,m}N_{sc}} \sum_{i=\sum_{t=1}^{m-1}M_{s,t}+1}^{\sum_{t=1}^{m}M_{s,t}} \sum_{p=1}^{N_{sc}} \mathcal{I}_Q^{symb}(SINR_i[\pi(p)]). \quad (11)$$

The mean mutual information per coded bit (MMIB) has been shown to be a good metric to represent the link quality [7] [8]. The packet error rate given a QAM modulation order and coding rate has been shown to be solely parameterized as a monotonically decreasing function of the mutual information [7]. We define this function as $\text{PER}_m = f_{\text{mcs}}(\mathcal{I}_m^{Packet})$, where PER_m denotes the packet error rate of the m-th packet and mcs denotes the modulation and coding scheme. The function $f_{\text{mcs}}(\cdot)$ can be approximately parameterized as the following [7]

$$f_{\text{mcs}}(x) = \frac{1}{2}\left[1 - erf\left(\frac{x - \text{coef}_1}{\sqrt{2}\text{coef}_2}\right)\right]. \quad (12)$$

the parameters coef_1 and coef_2 are summarized in Table II. The simulated packet error rates using LTE Turbo code versus the computed values using equation (12) are shown in Fig. 1. Note that the cost function is defined as the summation of the differences of the error rates in the $\log_{10}$ scale. Also, the function $f_{\text{mcs}}(\cdot)$ is approximately convex. The mean mutual information over a packet is used as the design metric for the precoder selection.

III. MEAN MUTUAL INFORMATION BASED (MMIB) PRECODING

The mean packet error rate over all N_{pac} packets is considered as the design objective. The design parameters are the N_{sb}

TABLE II
COEFFICIENTS FOR $f_{\text{MCS}}(\cdot)$ WITH QPSK, 16QAM AND 64QAM FOR LTE TURBO ENCODER

Modulation	Code Rate	coef$_1$	coef$_2$
QPSK	1/3	0.3648	0.0239
16QAM	1/2	0.5203	0.0224
16QAM	3/4	0.7501	0.0308
64QAM	5/6	0.8269	0.0337

precoders numbered as $\mathbf{F}_1, ..., \mathbf{F}_{N_{sb}}$ on different subcarrier subbands. Therefore the optimization problem is formulated as

$$\min_{\mathbf{F}_1,...,\mathbf{F}_{N_{sb}} \in \mathcal{C}} \frac{1}{N_{pac}} \sum_{m=1}^{N_{pac}} f\text{mcs}(\mathcal{I}_m^{Packet}(\mathbf{F}_1, ..., \mathbf{F}_{N_{sb}})). \quad (13)$$

A brute-forth search over all possible precoders on different subbands in the codebook has intractable complexity when the number of subbands is not small. We resort to simplifying this optimization problem using a bound on packet error rate.

We define the mean mutual information on the j^{th} subband for the m-th packet as the following

$$\mathcal{I}_{m,j}^{Subband}(\mathbf{F}_j) = \frac{1}{M_{s,m}N_{sub}} \sum_{i=\sum_{t=1}^{m-1} M_{s,t}+1}^{\sum_{t=1}^{m} M_{s,t}} \sum_{p=(j-1)*N_{sub}+1}^{N_{sub}*j} \mathcal{I}_Q^{symb}(SINR_i(\mathbf{F}_j)[\pi(p)]), \quad (14)$$

where the $SINR_i(\mathbf{F}_j)[\pi(p)]$ as a function of $\mathbf{F}_j$ is defined in (4).

Hence,

$$\begin{aligned} & \tfrac{1}{N_{pac}} \textstyle\sum_{m=1}^{N_{pac}} f\text{mcs}(\mathcal{I}_m^{Packet}(\mathbf{F}_1, ..., \mathbf{F}_{N_{sb}})) \\ = & \tfrac{1}{N_{pac}} \textstyle\sum_{m=1}^{N_{pac}} f\text{mcs}(\tfrac{1}{N_{sb}} \sum_{j=1}^{N_{sb}} \mathcal{I}_{m,j}^{Subband}(\mathbf{F}_j)) \\ \leq & \tfrac{1}{N_{pac}N_{sb}} \textstyle\sum_{m=1}^{N_{pac}} \sum_{j=1}^{N_{sb}} f\text{mcs}(\mathcal{I}_{m,j}^{Subband}(\mathbf{F}_j)). \quad (15) \end{aligned}$$

The last step follows the convexity of the function $f\text{mcs}(\cdot)$. Using this upper bound on PER, we reformulate the optimization problem as

$$\min_{\mathbf{F}_1,...,\mathbf{F}_{N_{sb}} \in \mathcal{C}} \frac{1}{N_{pac}N_{sb}} \sum_{m=1}^{N_{pac}} \sum_{j=1}^{N_{sb}} f\text{mcs}(\mathcal{I}_{m,j}^{Subband}(\mathbf{F}_j)). \quad (16)$$

This is equivalent to obtaining the minimum of the cost function for each $\mathbf{F}_j$ individually

$$\min_{\mathbf{F}_j \in \mathcal{C}} \frac{1}{N_{pac}} \sum_{m=1}^{N_{pac}} f\text{mcs}(\mathcal{I}_{m,j}^{Subband}(\mathbf{F}_j)). \quad (17)$$

This criteria is fundamentally different from [3] [4]. When the number of packet is one, the selection metric boils down to the following:

$$\max_{\mathbf{F}_j \in \mathcal{C}} \mathcal{I}_{m,j}^{Subband}(\mathbf{F}_j). \quad (18)$$

This is a simple function to compute and we do not need to use the curve fitting result in Section II-C. The MMIB based precoder selection algorithm is summarized in Table III.

The complexity of the MMIB based method at the receiver is roughly $\Theta(N_{sc}M_s|\mathcal{C}|) + \Theta(N_{sc}M_s^3|\mathcal{C}|) \approx \Theta(N_{sc}M_s^3|\mathcal{C}|)$ (taking into account of the MMIB computation in Table I and the matrix inversion at each subcarrier to compute the effective SINR), where Θ denotes the asymptotic tight bound of the computational complexity. For the MSV based approach, we first compute the average MIMO channel on each subband. Then singular value decomposition is applied to the average channel. The MSV based approach has a complexity which is roughly $\Theta(N_{sb}M_s^2M_r|\mathcal{C}|)$. When N_{sub} (the number of subcarriers in each subband) is not large, the complexity of the MMIB method is not significantly higher than the MSV method. Also we should note that $M_s < M_t$ is required for the MSV method with unitary precoding codebooks, however, we can have $M_s = M_t$ for the MMIB method (conditioned on that $M_s \leq M_r$).

TABLE III
ALGORITHM OF MMIB BASED PRECODER SELECTION:

Step 1: at the receiver, for each subband, per-subcarrier SINR is calculated for each stream using equation (4) for every precoding codeword in the codebook. **Step 2:** calculate the MMIB for the j^{th} subband taken by the m-th packet for every drawn precoding codeword using equation (14). **Step 3:** using equation (17), choose the desired precoding codeword for the j^{th} subband. **Step 4:** feed back the index of the chosen precoding codeword to the transmitter.

IV. SIMULATION RESULTS

Simulations over 3GPP Extended Pedestrian A (EPA) channel model [10] are conducted. The transmission strategy follows the 3GPP LTE standards, which uses MIMO-OFDM BICM. The total number of subcarriers is 2048. All packets are of 98 byte long. Rate 1/2 Turbo coding and 16QAM modulation are used for all packets. The consecutive 240 subcarriers (20 resource blocks) are allocated to each packet. The packet is then zero-padded to fit this resource allocation requirement. On each resource block (12 consecutive subcarrier [10], i.e., a subband that consists of one resource block only), a precoder is assigned.

First, simulations are done for a system that employs two transmit antennas and two receive antennas. Only one packet is sent and vertically encoded across the two transmit antennas. For this 2x2 system, the LTE precoding codebook that consists of two unitary codewords is used [10]. The simulation results are summarized in Fig. 2. We observe that approximately 0.8 dB and 0.4 dB gains are achieved by using the mean mutual information based precoding technique compared with the open loop spatial multiplexing scheme and the MMC precoding scheme respectively at PER = 0.2.

Then a system that employs four transmit antennas and two spatial streams is simulated. This system also uses vertical encoding and only sends one packet within each scheduling block. The receiver only needs two receive antennas to separate the two transmitted streams. Thus M_r is set to be two. We again uses the 3GPP LTE codebook (defined for the case of two streams and four transmit antennas) for the precoder selection. The simulation results are summarized in Fig. 3. We observe that approximately 0.5 dB and 0.8 dB gains are

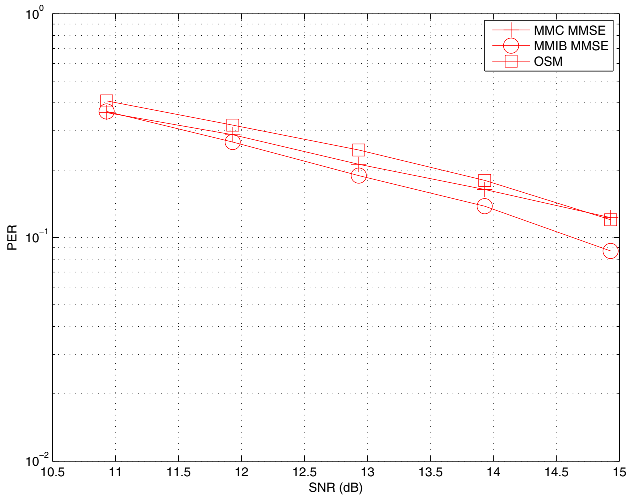

Fig. 2. Comparison of MMIB precoding, MMC precoding and open loop spatial multiplexing (OSM) for 2x2 16 QAM modulation over EPA channel.

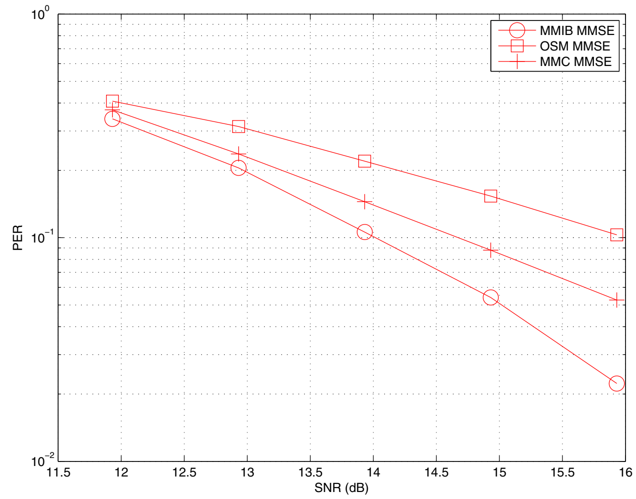

Fig. 4. Comparison of MMIB precoding, MMC precoding and open loop spatial multiplexing (OSM) for 4x4 16 QAM modulation over EPA channel.

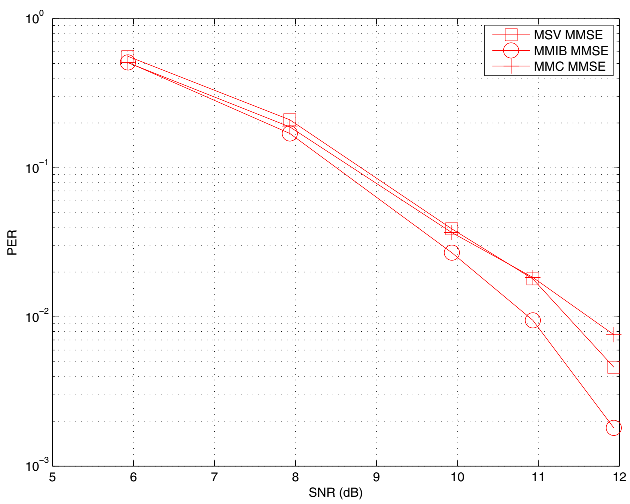

Fig. 3. Comparison of MMIB precoding, MMC precoding and MSV precoding for 2x4 16 QAM modulation over EPA channel.

achieved by MMIB precoding compared with MSV precoding and MMC precoding respectively at PER = 0.01.

For the last set of simulations, a 4x4 system is considered where four transmit antennas are used at the transmitter and four receive antennas are used at the receiver. Two packets each occupying two spatial streams are transmitted. Again, the 3GPP LTE codebook for precoding, which consists of unitary precoding codewords, is employed. The simulation results are summarized in Fig. 4. We can find that 2 dB and 0.7 dB gains are achieved for MMIB precoding compared with open loop spatial multiplexing and MMC precoding respectively at PER = 0.1.

V. Conclusion

In this paper, we proposed a mean mutual information based MIMO precoding technique that uses the mean mutual information per coded bit as the precoder selection metric for MIMO-OFDM systems. We compared the performance of the proposed precoding technique with open loop spatial multiplexing, minimum singular value based precoding and maximizing minimum capacity per packet precoding and observed that 0.4 - 2 dB gain can be achieved in different MIMO scenarios using the proposed method.

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