Performance of MIMO Techniques to Achieve Full Diversity and Maximum Spatial Multiplexing

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***Abstract*— In this paper, we analyze the performance of bit interleaved coded multiple beamforming (BICMB). We provide interleaver design criteria such that the resulting system achieves full spatial multiplexing of $\min(N,M)$ and full spatial diversity of NM for a system with N transmit and M receive antennas. We combined BICMB with OFDM in order to combat ISI caused by the frequency selective channels. The resulting system, named as BICMB-OFDM, achieves full spatial multiplexing of $\min(N,M)$, while maintaining full spatial and frequency diversity of NML for a $N \times M$ system over L-tap frequency selective channels when an appropriate convolutional code is used. Both systems analyzed in this paper assume perfect channel state information at the transmitter. Simulation results show that BICMB, and BICMB-OFDM provide substantial performance gains when compared to the other spatial multiplexing systems.**

I. Introduction

It is known that multi-input multi-output (MIMO) systems provide significant capacity increase [1]. MIMO systems also achieve a high diversity order. Some of the high diversity order achieving systems do not require channel state information (CSI) at the transmitter (e.g., space-time codes [2]). A technique that provides high diversity and coding gain with the help of CSI at the transmitter is known as beamforming. Singular value decomposition (SVD) based beamforming separates the MIMO channel into parallel subchannels. Therefore, multiple streams of data can be transmitted easily. Single beamforming (i.e., sending one symbol at a time) was shown to achieve the maximum diversity in space with a substantial coding gain compared to space-time codes [3]. If more than one symbol at a time are transmitted, then the technique is called multiple beamforming. For uncoded multiple beamforming systems, it was shown that while the data rate increases, one loses the diversity order with the increasing number of streams used over flat fading channels [4].

Bit interleaved coded modulation (BICM) was introduced as a way to increase the code diversity [5], [6]. BICM has been deployed with OFDM, and MIMO OFDM systems to achieve high diversity orders [7], [8], [9], [10]. In Section II-A, we analyze bit interleaved coded multiple beamforming (BICMB). We show that with the inclusion of BICM to the system, one does not lose the diversity order with multiple beamforming even when all the subchannels are used. That is, in Section II-B we show that BICMB achieves full diversity NM, and full spatial multiplexing $\min(N,M)$ for a system with N transmit and M receive antennas. In this letter, spatial multiplexing is defined as the number of symbols transmitted simultaneously over N transmit antennas. In order to guarantee full diversity, we provide design criteria for the interleaver.

If there is frequency selectivity in the channel, then as in Section III-A, we combined BICMB with OFDM in order to combat ISI. In Section III-B we show that BICMB-OFDM achieves full diversity NML, and full spatial multiplexing $\min(N,M)$ for a system with N transmit and M receive antennas over L-tap frequency selective channels, when an appropriate convolutional code is used.

II. Bit Interleaved Coded Multiple Beamforming (BICMB)

A. System Model

BICMB is a combination of BICM and multiple beamforming. The output bits of a binary convolutional encoder are interleaved and then mapped over a signal set $\chi \subseteq \mathbb{C}$ of size $|\chi| = 2^m$ with a binary labeling map $\mu : \{0,1\}^m \to \chi$. The minimum Hamming distance of the convolutional encoder, d_{free}, should satisfy $d_{free} \geq S$. The interleaver is designed such that the consecutive coded bits are

1) mapped over different symbols,
2) transmitted over different subchannels that are created by beamforming.

The reasons for the interleaver design are given in Section II-B. Gray encoding is used to map the bits onto symbols. During transmission, the code sequence $\underline{c}$ is interleaved by π, and then mapped onto the signal sequence $\underline{x} \in \chi$.

Beamforming separates the MIMO channel into parallel subchannels. The beamforming vectors used at the transmitter and the receiver can be obtained by the SVD [11] of the MIMO channel. Let H denote the quasi-static, flat fading $N \times M$ MIMO channel. Then the SVD of H can be written as

$$H = U\Lambda V^H = [u_1\, u_2 \ldots u_N]\Lambda[v_1\, v_2 \ldots v_M]^H \qquad (1)$$

where U and V are $N \times N$ and $M \times M$ unitary matrices, respectively, and Λ is an $N \times M$ diagonal matrix with singular values of H, $\lambda_i \in \mathbb{R}$ and non-negative, on the main diagonal with decreasing order. If S symbols are transmitted at the same time, then the system input-output relation at the k^{th} time instant can be written as

$$\begin{aligned} \underline{\mathrm{y}}_k &= \underline{\mathrm{x}}_k[u_1\, u_2 \ldots u_S]^H H[v_1\, v_2 \ldots v_S] + \underline{\mathrm{n}}_k[v_1\, v_2 \ldots v_S] \\ y_{k,s} &= \lambda_s x_{k,s} + n_{k,s}, \text{ for } s = 1, 2, \ldots, S \end{aligned} \tag{2}$$

where $\underline{\mathrm{n}}_k$ is $1 \times M$ additive white Gaussian noise with zero-mean and variance $N_0 = N/SNR$. Note that, the total power transmitted is scaled as N. The channel elements h_{nm} are modeled as zero-mean, unit-variance complex Gaussian random variables. Consequently, the received signal-to-noise ratio is SNR.

For an $N \times M$ uncoded multiple beamforming system, if S symbols are transmitted at a time, then it was shown that the diversity order for the uncoded multiple beamforming is equal to $(N - S + 1)(M - S + 1)$ [4].

The bit interleaver of BICMB can be modeled as $\pi : k' \to (k, s, i)$ where k' denotes the original ordering of the coded bits $c_{k'}$, k denotes the time ordering of the signals $x_{k,s}$ transmitted, s denotes the subchannel used to transmit $x_{k,s}$, and i indicates the position of the bit $c_{k'}$ on the symbol $x_{k,s}$.

Let χ_b^i denote the subset of all signals $x \in \chi$ whose label has the value $b \in \{0, 1\}$ in position i. Then, the ML bit metrics can be given by using (12), [5], [6]

$$\gamma^i(y_{k,s}, c_{k'}) = \min_{x \in \chi^i_{c_{k'}}} |y_{k,s} - x\lambda_s|^2. \tag{3}$$

The ML decoder at the receiver can make decisions according to the rule

$$\underline{\hat{\mathbf{c}}} = \arg\min_{\underline{\mathbf{c}} \in \mathcal{C}} \sum_{k'} \gamma^i(y_{k,s}, c_{k'}). \tag{4}$$

B. Pairwise Error Probability Analysis

In this section we are going to show that by using BICM, and the given interleaver design criteria, coded multiple beamforming can achieve full spatial diversity order of NM while transmitting $S \leq \min(N, M)$ symbols at a time. Assume the code sequence $\underline{\mathrm{c}}$ is transmitted and $\underline{\hat{\mathrm{c}}}$ is detected. Then, using (3), the PEP of $\underline{\mathrm{c}}$ and $\underline{\hat{\mathrm{c}}}$ given CSI can be written as

$$\begin{aligned} &P(\underline{\mathrm{c}} \to \underline{\hat{\mathrm{c}}} | H) = \\ &P\left(\sum_{k'} \min_{x \in \chi^i_{c_{k'}}} |y_{k,s} - x\lambda_s|^2 \geq \sum_{k'} \min_{x \in \chi^i_{\hat{c}_{k'}}} |y_{k,s} - x\lambda_s|^2\right) \end{aligned} \tag{5}$$

where $s \in \{1, 2, \ldots, S\}$.

For a convolutional code with rate k_0/n_0, the minimum Hamming distance between $\underline{\mathrm{c}}$ and $\underline{\hat{\mathrm{c}}}$, $d(\underline{\mathrm{c}} - \underline{\hat{\mathrm{c}}})$, is d_{free}. Assume $d(\underline{\mathrm{c}} - \underline{\hat{\mathrm{c}}}) = d_{free}$ for $\underline{\mathrm{c}}$ and $\underline{\hat{\mathrm{c}}}$ under consideration for PEP analysis. Then, $\chi^i_{c_{k'}}$ and $\chi^i_{\hat{c}_{k'}}$ are equal to one another for all k' except for d_{free} distinct values of k'. Therefore, the inequality on the right hand side of (5) shares the same terms on all but d_{free} summation points. Hence, the summations can be simplified to only d_{free} terms for PEP analysis.

Note that for binary codes and for the d_{free} points at hand, $\hat{c}_{k'} = \bar{c}_{k'}$. For the d_{free} bits let's denote $\tilde{x}_{k,s} = \arg\min_{x \in \chi^i_{c_{k'}}} |y_{k,s} - x\lambda_s|^2$, and $\hat{x}_{k,s} = \arg\min_{x \in \chi^i_{\bar{c}_{k'}}} |y_{k,s} - x\lambda_s|^2$. It is easy to see that $\tilde{x}_{k,s} \neq \hat{x}_{k,s}$ since $\tilde{x}_{k,s} \in \chi^i_{c_{k'}}$ and $\hat{x}_{k,s} \in \chi^i_{\bar{c}_{k'}}$ where $\chi^i_{c_{k'}}$ and $\chi^i_{\bar{c}_{k'}}$ are complementary sets of constellation points within the signal constellation set χ. Also, $|y_{k,s} - x_{k,s}\lambda_s|^2 \geq |y_{k,s} - \tilde{x}_{k,s}\lambda_s|^2$ and $x_{k,s} \in \chi^i_{c_{k'}}$.

For convolutional codes, due to their trellis structure, d_{free} distinct bits between any two codewords occur in consecutive trellis branches. Let's denote d such that d_{free} bits occur within d consecutive bits. The bit interleaver can be designed such that d consecutive bits are mapped onto distinct symbols (interleaver design criterion 1). This guarantees that there exist d_{free} distinct pairs of $(\tilde{x}_{k,s}, \hat{x}_{k,s})$, and d_{free} distinct pairs of $(x_{k,s}, \hat{x}_{k,s})$. The PEP can be rewritten as

$$\begin{aligned} &P(\underline{\mathrm{c}} \to \underline{\hat{\mathrm{c}}} | H) = \\ &P\left(\sum_{k, d_{free}} |y_{k,s} - \tilde{x}_{k,s}\lambda_s|^2 - |y_{k,s} - \hat{x}_{k,s}\lambda_s|^2 \geq 0\right) \\ &\leq Q\left(\sqrt{\frac{d^2_{min} \sum_{s=1}^{S} \alpha_s \lambda_s^2}{2N_0}}\right) \end{aligned} \tag{6}$$

where $\sum_{k, d_{free}}$ denotes that the summation is taken with index k over d_{free} different values of k, and d_{min} is the minimum Euclidean distance between two constellation points.

If the interleaver is designed such that the consecutive coded bits of length equal to the interleaver depth are transmitted on the same subchannel, then the performance is dominated by the worst singular value. In other words, the error event on the trellis occurs on the consecutive branches spanned by the worst subchannel, and $\alpha_S = d_{free}$. This results in a diversity order of $(N - S + 1)(M - S + 1)$ as in uncoded multiple beamforming. However, the interleaver can be designed such that the consecutive coded bits are transmitted on different subchannels (interleaver design criterion 2). Criterion 2 guarantees that $\alpha_s \geq 1$, for $s = 1, 2, \ldots, S$. This way, on the trellis, within the d_{free} bits under consideration, coded bits that are transmitted on better subchannels can provide better error correcting on the neighboring bits that are transmitted on worse subchannels. Using an upper bound for the Q function $Q(x) \leq (1/2)e^{-x^2/2}$, PEP can be upper bounded as

$$P(\underline{\mathrm{c}} \to \underline{\hat{\mathrm{c}}}) \leq E\left[\frac{1}{2}\exp\left(-\frac{d^2_{min} \sum_{s=1}^{S} \alpha_s \lambda_s^2}{4N_0}\right)\right]. \tag{7}$$

Let's denote $\alpha_{min} = \min\{\alpha_s : s = 1, 2, \ldots, S\}$. Then,

$$\frac{\sum_{s=1}^{S} \alpha_s \lambda_s^2}{S} \geq \frac{\alpha_{min} \sum_{s=1}^{S} \lambda_s^2}{S} \geq \frac{\alpha_{min} \sum_{s=1}^{N} \lambda_s^2}{N}. \tag{8}$$

Note that, $\Theta \overset{\triangle}{=} \sum_{s=1}^{N} \lambda_s^2 = \|H\|_F^2 = \sum_{n,m} |h_{n,m}|^2$ is a chi-squared random variable with $2NM$ degrees of freedom (the elements of H, $h_{n,m}$, are complex Gaussian random variables). Then, PEP can be upper bounded by

$$P(\underline{c} \to \underline{\hat{c}}) \leq E\left[\frac{1}{2}\exp\left(-\frac{d_{min}^2 \alpha_{min} S}{4N_0 N}\Theta\right)\right]. \quad (9)$$

The expectation in (9) is taken with respect to Θ with pdf $f_\Theta(\theta) = \theta^{(NM-1)} e^{-\theta/2} / 2^{NM}(NM-1)!$ [12]. Consequently,

$$P(\underline{c} \to \underline{\hat{c}}) \leq \frac{1}{2^{NM+1}}\left(\frac{d_{min}^2 \alpha_{min} S}{4N_0 N} + \frac{1}{2}\right)^{-NM} \quad (10)$$

$$\approx \frac{1}{2^{NM+1}}\left(\frac{d_{min}^2 \alpha_{min} S}{4N^2} SNR\right)^{-NM} \quad (11)$$

for high SNR. As can be seen from (11) the diversity order of BICMB at high SNR is NM. Consequently, BICMB achieves full diversity order independent of the number of spatial streams transmitted.

A very low complexity decoder for BICM can be implemented as in references [13], [14]. The same decoder can be used for BICMB as well: Instead of using the single-input single-output (SISO) channel value of BICM-OFDM for the decoder ([13] and [14]), one should use λ_s. Hence, BICMB provides a full spatial multiplexing, full diversity, and easy-to-decode system.

III. Bit Interleaved Coded Multiple Beamforming with OFDM (BICMB-OFDM)

A. System Model

In order to combat the ISI in frequency selective channels, we combined BICMB with OFDM and named the system as BICMB-OFDM. The system model is similar to BICMB with few minor differences as given in this section. The interleaver is designed such that the consecutive coded bits are

1) interleaved within one MIMO-OFDM symbol to avoid extra delay requirement to start decoding at the receiver,
2) mapped over different symbols,
3) transmitted over different subcarriers of an OFDM symbol,
4) transmitted over different subchannels that are created by beamforming.

By adding cyclic prefix (CP), OFDM converts the frequency selective channel into parallel flat fading channels for each subcarrier. Let $H(k)$ denote the quasi-static, flat fading $N \times M$ MIMO channel observed at the k^{th} subcarrier, and $\underline{h}_{nm} = [h_{nm}(0) \quad h_{nm}(1) \quad \cdots \quad h_{nm}(L-1)]^T$ represent the L-tap frequency selective channel from the transmit antenna n to the receive antenna m. Each tap is assumed to be statistically independent and modeled as zero mean complex Gaussian random variable with variance $1/L$. If S symbols are transmitted on the same subcarrier over N transmit antennas, then the system input-output relation at the k^{th} subcarrier can be written as

$$y_s(k) = \lambda_s(k) x_s(k) + n_s(k) \quad (12)$$

for $s = 1, 2, \ldots, S$, and $k = 1, 2, \ldots, K$ where $n_s(k)$ is the additive white complex Gaussian noise. Note that, the total power transmitted and the noise power are scaled such that the received signal-to-noise ratio is SNR.

B. Pairwise Error Probability Analysis

Using a similar analysis of Section II-B, and the interleaver design criteria of Section III-A, PEP can be upper bounded as

$$\begin{aligned} P(\underline{c} \to \underline{\hat{c}}) =& E\left[P(\underline{c} \to \underline{\hat{c}} | H(k), \forall k)\right] \\ \leq& E\left[\frac{1}{2}\exp\left(-\frac{d_{min}^2 \sum_{k, d_{free}} \lambda_s^2(k)}{4N_0}\right)\right]. \end{aligned} \quad (13)$$

Assuming high frequency selectivity in the channel, $\lambda_s(k)$s are independent for different k, and identically distributed for the same s. Let's denote $\mu_s(k) = \lambda_s^2(k)$. The joint pdf of $\mu_s(k)$, for given k, is given in [15]. The joint pdf is in exponential form, and for the diversity analysis when the integration in (13) is taken, the upper limit of the integration (infinity) tends to zero. Therefore, an approximation of marginal pdfs of each $\mu_s(k)$ around zero can be used. Approximation to the marginal pdfs of each $\mu_s(k)$ is given in [16], [4] as

$$f(\mu_s(k)) \approx \kappa \mu_s(k)^{(N-s+1)(M-s+1)-1} \quad (14)$$

where κ is a scaling constant. Let's denote α_s as the number of times the s^{th} channel is used within d_{free} bits under consideration such that $\sum_{s=1}^{S} \alpha_s = d_{free}$. Note that, criterion 4 guarantees $\alpha_s \geq 1, \forall s$. The expectation in (13) can be taken using the marginal pdfs of (14)

$$\begin{aligned} P(\underline{c} \to \underline{\hat{c}}) \leq& \frac{1}{2}\prod_{s=1}^{S}\left[\int \exp\left(-\frac{d_{min}^2 \mu_s(k)}{4N_0}\right) f(\mu_s(k)) d_{\mu_s(k)}\right]^{\alpha_s} \\ =& \frac{1}{2}\prod_{s=1}^{S} \kappa^{\alpha_s}\left(\frac{d_{min}^2}{4N} SNR\right)^{-\alpha_s (N-s+1)(M-s+1)}. \end{aligned} \quad (15)$$

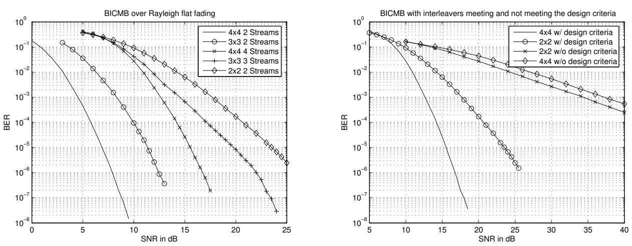
As can be seen from (15), BICMB-OFDM provides a diversity order of $\sum_{s=1}^{S} = \alpha_s (N-s+1)(M-s+1)$ for a spatial multiplexing of S. Note that, if the interleaver design criterion 4 is not met, then the maximum diversity order reduces to $(N-S+1)(M-S+1)d_{free}$ for spatial multiplexing of S. It is known that the maximum diversity order of MIMO systems over L-tap frequency selective channels is NML [17], [18]. As will be shown in Section IV-B, BICMB-OFDM achieves full diversity order of NML when $NML \leq \sum_{s=1}^{S} = \alpha_s(N-s+1)(M-s+1)$ for spatial multiplexing of S.

IV. Simulation Results

In the simulations below, the industry standard 64-state 1/2-rate (133,171) $d_{free} = 10$ convolutional code is used. For BICMB, coded bits are separated into different streams of data and a random interleaver is used to interleave the bits in each substream. BICMB-OFDM deploys the interleaver given in [19]. The interleavers guarantee the design criteria of Sections II, and III. The coded bits are mapped onto symbols using 16 QAM with Gray labeling. Each packet has 1000 bytes of information bits, and the channel is changed independently from packet to packet. Each OFDM symbol has 64 subcarriers, and has 4 μs duration, of which 0.8 μs is CP. All the comparisons below are carried at 10^{-5} bit error rate (BER).

A. BICMB

As can be seen from Figure 1 (a), while transmitting at spatial multiplexing of 2, 3, and 4, BICMB achieves full diversity at high SNR for the 2×2, 3×3, and 4×4 cases. Even though the 4×4 system transmits twice the data rate of 2x2 system, the performance of 4×4 system is significantly better than the the 2×2 system. This is due to the fact that the 4×4 system achieves a diversity order of 16 where the 2×2 system has a diversity order of 4. Consequently, BICMB provides both advantages of MIMO systems: It provides full diversity and full spatial multiplexing.

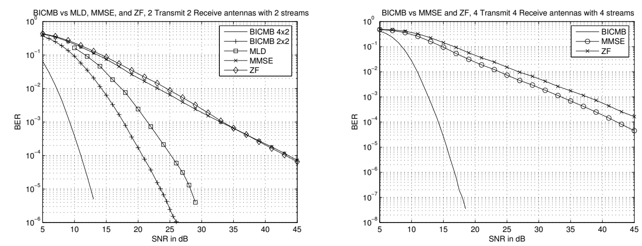


(a) BICMB Results (b) Interleaver Effect

Fig. 1. BICMB over Rayleigh flat fading channel.

Figure 1 (b) illustrates the importance of the interleaver design. We simulated a random interleaver such that consecutive coded bits are transmitted over the same subchannel. In other words, on a trellis path, consecutive bits of length $1/S$th of the coded packet size are transmitted over the same subchannel. Consequently, an error on the trellis occurs over the paths that are spanned by the worst channel, and the diversity order of coded multiple beamforming approaches to that of uncoded multiple beamforming.

Figures 2 (a) and (b) show the simulation results of BICMB when compared to maximum likelihood decoder (MLD), minimum mean squared error receiver (MMSE), and zero forcing receiver (ZF) for spatial multiplexing of 2 and spatial multiplexing of 4 case, respectively. While MLD achieves a high diversity order with substantial complexity, ZF achieves a diversity order of $M - N + 1$ [20], [21]. MLD is known as the optimal receiver for a spatial multiplexing system. Using BICM at the transmitter with an interleaver spreading the consecutive bits over the transmit antennas and deploying MLD at the receiver end can be considered as the Vertical Encoding (VE) in [20]. Such a system is capable of providing a high diversity order. However, its substantially high complexity makes it almost impossible to implement. Therefore, sub-optimal (therefore poorer performance) but easy-to-implement receivers are designed such as MMSE, ZF, successive cancellation (SUC) and ordered SUC [20]. As illustrated for the 2x2 case, BICMB outperforms MLD by 4.5dB, while the performance gain compared to MMSE and ZF is more than 25dB. It is possible that the base station (or the access point) has more antennas than the receiver. BICMB with 4 transmit and 2 receive antennas with spatial multiplexing of 2 outperforms MLD by 15.5dB. When spatial multiplexing is 4, BICMB outperforms MMSE and ZF by more than 30dB.



(a) 2×2 (b) 4×4

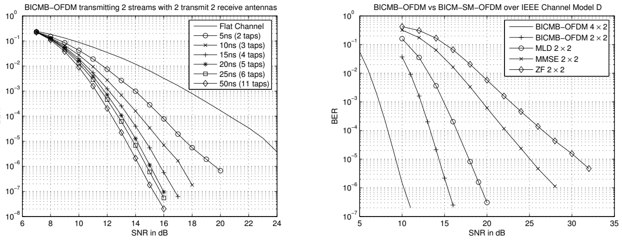
Fig. 2. BICMB vs MLD, ZF and MMSE.

B. BICMB-OFDM

Figure 3 (a) illustrates the results for BICMB-OFDM for different rms delay spread values, when 2 streams of data are transmitted at the same time. The maximum delay spread of the channel is assumed to be ten times the rms delay spread. The channel is modeled as in Section III-A, where each tap is assumed to have equal power. The spectrum of (133,171) shows that there are 11 codewords with an Hamming distance of d_{free} from all-zero codeword. When compared to all-zero codeword, the codeword [11100101000101011100000000...] has the worst performance for BICMB-OFDM. On this codeword $\alpha_1 = 3$, and $\alpha_2 = 7$. Consequently, when $S = 2$, BICMB-OFDM achieves a maximum diversity order of $3NM + 7(N-1)(M-1)$ (19 for 2×2 system). Note that, up to 15 ns rms delay spread, BICMB-OFDM achieves the maximum diversity with full spatial multiplexing of 2. A 2×2 system over 20 ns channel provides a maximum achievable diversity order of 20. Therefore, BICMB-OFDM achieves a diversity order of 19 for rms delay spreads of 20 ns, 25 ns, and 50 ns.

Figure 3 (b) illustrates the simulation results for BICMB-OFDM, BICM-OFDM with spatial multiplexing (BICM-SM-OFDM) using MLD, MMSE, and ZF. The spatial multiplexing is set as two. The simulations are carried over the IEEE channel model D [22], [23], [24]. As can be seen, BICMB-OFDM outperforms significantly high complexity, but best spatial multiplexing receiver, MLD, by more than 3.5 dB. The decoding complexity of BICMB-OFDM is substantially lower in complexity than MLD. BICMB-OFDM outperforms easy-to-implement MMSE and ZF receivers by more than 10 dB and more than 15 dB, respectively. BICMB-OFDM with 4 transmit and 2 receive antennas with spatial multiplexing of 2 outperforms MLD by 9 dB.

At this point, we would like to make the following important point: In all the simulations presented in this section, it is



(a) Equal Power Taps (b) IEEE Channel Model D

Fig. 3. (a) BICMB-OFDM over equal power frequency selective channels, and (b) BICMB-OFDM vs MLD, MMSE, and ZF transmitting 2 streams over IEEE Channel Model D.

assumed that the beamforming vectors are perfectly known at the transmitter. This may not be the case for a practical system, since it may require a high-speed feedback channel depending on the application. However, as shown in the figures, BICMB based systems provide substantial gain when compared to current practical systems. This substantial gain with perfect feedback leaves room for significant gain with limited feedback BICMB and BICMB-OFDM systems. Our goal in this paper has been to provide the performance analysis of BICMB and BICMB-OFDM with the given interleaver design criteria using perfect CSI assumption at the transmitter. The performance of BICMB based systems with limited feedback is left as future work.

V. Conclusion

In this paper, we analyzed bit interleaved coded multiple beamforming (BICMB). BICMB utilizes the channel state information at the transmitter and the receiver. By doing so, BICMB achieves full spatial multiplexing of $\min(N, M)$, while maintaining full spatial diversity of NM for a $N \times M$ system. We presented interleaver design guidelines to guarantee full diversity at full spatial multiplexing. We combined BICMB with OFDM in order to combat ISI caused by the frequency selective channels. The resulting system, named as BICMB-OFDM, achieves full spatial multiplexing of $\min(N, M)$, while maintaining full spatial and frequency diversity of NML for a $N \times M$ system over L-tap frequency selective channels when an appropriate convolutional code is used.

Simulation results also showed that BICMB and BICMB-OFDM outperforms the optimal high complexity MLD, and easy-to-implement MMSE and ZF receivers substantially. The BICMB and BICMB-OFDM systems analyzed in this paper use perfect channel information at the transmitter end. This may not be the case for a practical system. However, the performance gains compared to more practical systems are encouraging to investigate limited feedback problem for BICMB based systems.

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