A New Diagonally Layered Spatial Multiplexing Scheme with Limited Feedback

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***Abstract*—Multiple antenna wireless technology has the potential to enable high data rate wireless applications. To realize the spatial diversity offered by a MIMO system, either the receiver has to perform complex signal processing (such as ML decoding) or the transmitter has to preprocess the signals which requires channel knowledge at the transmitter, often through a feedback link. In this paper, we propose a new space-time signaling scheme for an $N_r \times N_t$ MIMO system that improves the diversity gain of the weak layers by encoding information along multiple dimensions and interleaving the co-ordinates of the symbols over all the layers. We prove analytically that the proposed scheme achieves full spatial diversity of N_tN_r. We also present simulation results that confirm our analytical results and show the superior performance of our method in comparison with existing methods. When compared with other diagonally layered schemes, the proposed scheme requires a very low feedback of $log_2 N_t!$ bits and attains full diversity at the cost of a slight increase in coding and decoding complexity.**

Notations. Vectors are denoted by lowercase letters in boldface and matrices by capital letters in boldface.$(\cdot)^T$ and $(\cdot)^H$ denote transpose and conjugate transpose, respectively. E is the expectation operator and $\|\cdot\|$ stands for Euclidean norm. $tr(.)$ denotes trace of a matrix.

I. Introduction

Multiple-input multiple-output (MIMO) wireless systems, employing multiple transmit and receive antennas, promise significant improvement in the capacity compared to conventional single-input single-output (SISO) systems. It was shown that the capacity of MIMO wireless systems increases linearly with the minimum of the number of antennas at the transmitter and receiver in rich scattering environments [1], [2]. The increase in data rate can be achieved through spatial multiplexing (SM) as shown by Bell labs layered space time (BLAST) scheme [3]. Initially, MIMO systems were developed assuming channel state information (CSI) only at the receiver. When perfect CSI is available also at the transmitter, channel-dependent preprocessing or precoding the data streams can further improve performance by adapting the transmitted signal to the instantaneous channel realization.

II. System model

The discrete-time input-output relation of a MIMO system with N_t transmit antennas and N_r receive antennas is given by

$$\mathbf{y}(k) = \mathbf{H}\mathbf{x}(k) + \mathbf{n}(k) \tag{1}$$

where, $\mathbf{x}(k) = [x_1 x_2 \ldots x_{N_t}]^T \in \mathbb{C}^{N_t \times 1}$ is the transmit signal vector at time instant k and $trE[\mathbf{x}(k)\mathbf{x}(k)^H] = 1$. The received signal vector at time instant k is given by $\mathbf{y}(k) = [y_1 y_2 \ldots y_{N_r}]^T \in \mathbb{C}^{N_r \times 1}$. $\mathbf{n}(k)$ denotes the channel noise with i.i.d entries distributed according to $\mathcal{CN}(0, \sigma_n^2)$.

$\mathbf{H} \in \mathcal{CN}(N_r, N_t)$ is the channel matrix with entries h_{ij} denoting the channel gain from the j^{th} transmit antenna to the i^{th} receive antenna. h_{ij} are assumed to be i.i.d. complex Gaussian random variables with zero mean and unit variance, i.e. $h_{ij} \in \mathcal{CN}(0, 1)$.

Also, perfect channel knowledge is assumed at the receiver.

III. Achieving the available spatial diversity in MIMO systems

An $N_r \times N_t$ MIMO channel provides a maximum diversity gain of N_tN_r.

In a frequency flat narrow band MIMO system, when there is channel knowledge only at the receiver, the optimal receiver performs ML decoding achieving N_r^{th} order diversity. But the complexity scales as m^{N_t}, where m denotes the cardinality of the complex signal set $\mathcal{A}$. With increase in N_t and/or m, the complexity of ML receiver becomes prohibitive.

Successive interference cancelation (SIC) receiver [4] is a sub-optimal low complexity approach that performs better than other low-complexity receivers. It decodes the symbols sequentially by using *successive nulling and cancellation* technique[1]. $\mathbf{H}$ is decomposed as $\mathbf{H} = \mathbf{QR}$ using standard QR decomposition, where $\mathbf{Q} \in \mathbb{C}^{N_r \times N_r}$ is unitary matrix and $\mathbf{R} \in \mathbb{C}^{N_r \times N_t}$ is upper triangular matrix with diagonal elements $r_{kk} \in \mathbb{R}$. Left-multiplying $\mathbf{y}$ with $\mathbf{Q}^H$ yields, $\mathbf{z} = \mathbf{Rx} + \mathbf{w}$ where, $\mathbf{z} = \mathbf{Q}^H\mathbf{y}$ and $\mathbf{w} = \mathbf{Q}^H\mathbf{n}$. The estimates $\hat{x}_{N_t}, \ldots, \hat{x}_1$ are obtained by (successive) back-substitution. Ignoring error propagation, we see that the SIC receiver decomposes the MIMO channel into N_t parallel SISO channels

$$z_k = r_{kk}x_k + w_k, \quad k = 1, 2, \ldots, N_t \tag{2}$$

The symbol error probability of the k^{th} layer or k^{th} sub-stream depends on r_{kk} and it can be shown that it has a diversity gain of $N_r - k + 1$ [5]. The overall diversity gain is limited by the last layer having a diversity gain of $N_r - N_t + 1$. For example, when $N_r = N_t$, the system diversity gain gets reduced to unity.

If channel knowledge is available *also* at the transmitter, channel-dependent *precoding* of data streams can improve the performance of the MIMO system. Next, we discuss one such precoding scheme with partial CSI that improves the performance of the SIC receiver.

[1]Note that SIC requires $N_r \geq N_t$

A. SIC Receiver based on Greedy QR Decomposition (GQR-SIC)

To improve the diversity performance of SIC receiver, [9] has proposed a diversity optimal SIC receiver , based on greedy QR decomposition of $\mathbf{H}$. In greedy QR decomposition, a permutation matrix $\mathbf{\Pi}$ is found such that the QR decomposition of $\mathbf{H\Pi}$ yields an $\mathbf{R}$ matrix with the following property: r_{kk}, the k^{th} diagonal element of $\mathbf{R}$ satisfies

$$\lim_{\epsilon\to 0^+}\frac{\log \Pr(r_{kk}^2<\epsilon)}{\log\epsilon}=(N_t-k+1)(N_r-k+1)$$

As shown in [9], it implies that the k^{th} layer of GQR-SIC receiver has a diversity gain of $(N_t-k+1)(N_r-k+1)$.

To implement GQR-SIC, the receiver has to find the permutation matrix $\mathbf{\Pi}$ from its knowledge of $\mathbf{H}$ and has to feed it back to the transmitter. This results in a feedback of $\log_2(N_t!)$ bits. At the transmitter, the symbol vector $\mathbf{x}$ is pre-multiplied by $\mathbf{\Pi}$. GQR-SIC does not impose the restriction that $N_r \geq N_t$ as the permutation matrix selects only N_r transmit antennas out of N_t. It was proved in [9] that GQR-SIC is *diversity gain optimal* among all ordered SIC receivers.

The overall diversity gain of the GQR-SIC receiver is $g_d^{\text{GQR}}=|N_r-N_t|+1$. In square MIMO systems, i.e., when $N_t=N_r$, g_d^{GQR} gets reduced to unity.

We now propose a new diagonally layered spatial multiplexing scheme that employs *co-ordinate interleaving* over a multidimensional QAM constellation to improve the overall diversity gain. The constellation used while transmitting the symbols would still be over two dimensions. We refer to the proposed scheme as *multi-dimensional co-ordinate interleaved spatial multiplexing* (MD CISM).

IV. Multi-Dimensional Co-ordinate Interleaved Spatial Multiplexing (MD CISM)

Co-ordinate interleaving with input symbols from *rotated* multi-dimensional QAM constellations has been first proposed in [7] for improving diversity gains over single-antenna Rayleigh fading channels. This technique exploits the *co-ordinate* (or *component*) level diversity by transmitting different co-ordinates of the input symbols over independently fading channels. This is also known as signal space diversity. In [8],based on co-ordinate interleaving, a dialogonally layered CISM scheme has been introduced for a 2×2 MIMO-OFDM system, which achieves full diversity with a feedback of 1 bit per tone. When this scheme is extended for a MIMO-OFDM system with higher number of transmit and receive antennas, the diversity gain of the overall system is limited by the diversity gain of the $floor(N_t/2)+1$ layer.

We now present a novel spatial multiplexing scheme that uses co-ordinate interleaving over multi-dimensional QAM constellation to achieve full diversity. The constellation used while transmitting the symbols would still be over two dimensions. We specifically take the case of a 4×4 MIMO system to explain the transceiver structure. The proposed scheme can be easily extended for any number of transmit and receive antennas, provided $N_r \geq N_t$.

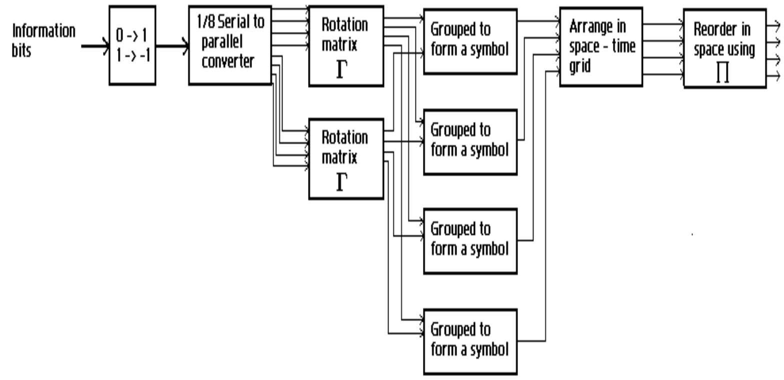

Fig. 1. Transmitter scheme for MD CISM

In the proposed scheme, we precode (or pre-multiply) the transmitted symbol vector with the permutation matrix $\mathbf{\Pi}$ fed back to the transmitter. As the precoding can be absorbed into $\mathbf{H}$ to get a new effective channel $\mathbf{H\Pi}$, in the following discussion of MD CISM, we do not explicitly show pre-multiplying the symbol vector with $\mathbf{\Pi}$ for simplifying the presentation.

A. MD CISM Transmission

Figure 1 shows the transmitter schematically.

The input bit stream is passed through a $1/8$ serial-to-parallel converter after mapping 0 to 1 and 1 is mapped to -1. 4 consecutive bits from the output of the serial to parallel converter are grouped to form a 4-dimensional QAM symbol. It is then rotated using a rotation matrix $\mathbf{\Gamma}$ obtained from [7]. Hence the resulting symbols are from a constellation $e^{\jmath\theta}\mathcal{A}$, where $\mathcal{A}$ is a standard (un-rotated) 4-dimensional QAM signal set and θ is the angle of rotation. Let us denote the two symbols thus obtained as $[x_1,x_2,x_3,x_4]$ and $[x_5,x_6,x_7,x_8]$. Co-ordinates of two consecutive symbols are then paired up as shown in Figure 1 to form symbols over 2 dimensions. The resultant output is then arranged in a space-time grid as shown in Figure 2. We denote the resultant symbol vector at time t as $\bar{\mathbf{x}}(\mathbf{t})=[\bar{x}_1(t),\ \bar{x}_2(t),\ \bar{x}_3(t),\ \bar{x}_4(t)]^T$, each co-ordinate of $\bar{\mathbf{x}}(\mathbf{t})$ corresponding to each antenna. Before transmission, the symbols are permuted along space using a permutation matrix $\mathbf{\Pi}$ obtained using greedy QR decomposition.

The symbols are thus, co-ordinate interleaved along each diagonal. As in a D-BLAST system, initially zeroes are filled till the first diagonal.

While terminating the transmission, there is no need to transmit zeroes for the strongest, i.e. the first layer. Hence, a total of 3 zeros are transmitted, 1 zero corresponding to the third layer for the last time instant and 2 zeroes, corresponding to the second layer for the last and last but one time instants.

As we know, V-BLAST for a 4×4 MIMO system transmits $4t_f$ symbols in t_f symbol durations resulting in a spectral efficiency of 4 symbols per channel use while MD CISM

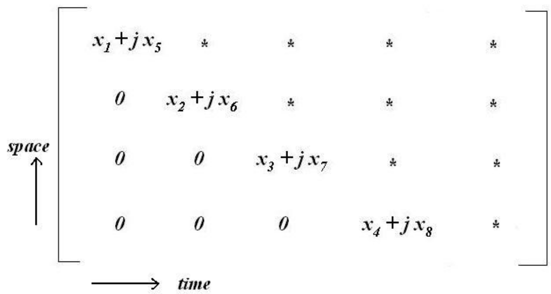

Fig. 2. Space-time grid for MD CISM

transmits $(4t_f-3-6)/t_f$ symbols per channel use. The loss in spectral efficiency becomes negligible as t_f increases.

B. MD CISM Receiver

We decode the transmitted symbols through GQR-SIC receiver. At the first time instant, i.e., at $t=1$, we receive $\mathbf{y}(1)=\mathbf{H\Pi}\bar{\mathbf{x}}(1)+\mathbf{n}(1)$. We decompose $\mathbf{H\Pi}$ as $\mathbf{H\Pi}=\mathbf{QR}$, and left-multiply $\mathbf{y}$ with $\mathbf{Q}^H$. With a slight abuse of notataion, we express the resulting input-output relation as $\mathbf{y}(1)=\mathbf{R}\bar{\mathbf{x}}(1)+\mathbf{w}(1)$.This equation can be broken down as,

$$y_1(1)=r_{11}\bar{x}_1(1)+w_1(1) \tag{3}$$

From the received symbol vector at time $t=2$, $\mathbf{y}(2)=[y_1(2),\ y_2(2),\ y_3(2),\ y_4(2)]$, we obtain,

$$y_1(2)=r_{11}\bar{x}_1(2)+r_{12}\tilde{x}_2(2)+w_1(2) \tag{4}$$
$$y_2(2)=r_{22}\bar{x}_2(2)+w_2(2) \tag{5}$$

Similarly, when $t=3$,

$$y_1(3)=r_{11}\bar{x}_1(3)+r_{12}\bar{x}_2(3)+r_{13}\bar{x}_3(3)+w_1(3) \tag{6}$$
$$y_2(3)=r_{22}\bar{x}_2(3)+r_{23}\bar{x}_3(3)+w_2(3) \tag{7}$$
$$y_2(3)=r_{33}\bar{x}_3(3)+w_2(3) \tag{8}$$

A similar expression can be obtained for the case when $t=4$.

$y_1(1),y_2(2),y_3(3)$ and $y_4(4)$ are free of inter antenna interference. From the real and imaginary part of these received symbols, we obtain z_1 and z_2, given by

$$\mathbf{z_1}=\Re[y_1(1),y_2(2),y_3(3),y_4(4)] \tag{9}$$
$$\mathbf{z_2}=\Im[y_1(1),y_2(2),y_3(3),y_4(4)] \tag{10}$$

This can be re-written as

$$\mathbf{z_1}=[r_{11}x_1,\ r_{22}x_2,\ r_{33}x_3,\ r_{44}x_4]+ \tag{11}$$
$$\Re[w_1(1),\ w_2(2),\ w_3(3),\ w_4(4)] \tag{12}$$
$$\mathbf{z_2}=[r_{11}x_5,\ r_{22}x_6,\ r_{33}x_7,\ r_{44}x_8]+ \tag{13}$$
$$\Im[w_1(1),\ w_2(2),\ w_3(3),\ w_4(4)] \tag{14}$$
$$\tag{15}$$

Now, $\hat{\mathbf{x}}_\mathbf{1}=[\hat{x_1},\ \hat{x_2},\ \hat{x_3},\ \hat{x_4}]$ can be obtained from z_1 through single-symbol ML decoding:

$$\hat{\mathbf{x}}_\mathbf{1}=\arg\min_{x_1\in\mathcal{A}}\left\|\mathbf{z_1}-[r_{11}x_1,\ r_{22}x_2,\ r_{33}x_3,\ r_{44}x_4]\right\|^2 \tag{16}$$

Similarly, $\hat{\mathbf{x_2}}=[\hat{x_5},\ \hat{x_6},\ \hat{x_7},\ \hat{x_8}]$ can be obtained from $\mathbf{z_2}$. Next, we interleave the co-ordinates of $\hat{\mathbf{x_1}}$ and $\hat{\mathbf{x_2}}$ to cancel out the interference caused by these symbols in $y_1(2),y_2(3)$ and $y_3(4)$ using (4), (7). Again, single- symbol ML decoding is performed and successive interference cancelation done. Note that because of the way decoding is done, the symbols going through the weak layers enjoy the diversity benefit of the stronger layers.

V. Diversity analysis of the proposed scheme

In the following, we compute the diversity gain of all the data streams transmitted in MD CISM. To be precise, we obtain the SNR exponent of $P_{s,k}$, SEP of k^{th} data stream (i.e., k^{th} layer), $k=1,2,\ldots,N_t$. We assume that the same signal set $e^{\jmath\theta}\mathcal{A}$, where $\mathcal{A}$ is a standard (un-rotated) M-dimensional QAM signal set and θ is the angle of rotation, is employed for all the data streams. The rotation angle θ is chosen such that no two signal points in the rotated constellation have the same ordinate [6]. With slight abuse of notation, we denote $e^{\jmath\theta}\mathcal{A}$ by $\mathcal{A}$.

We assume uniform power allocation to all the data streams, i.e., $p_1=p_2=\ldots=p_{N_t}=P/N_t$. Assume that an arbitrary symbol x^l in the constellation is transmitted. Let $\Pr\left\{x^l\rightarrow x^i\right\}$ be the pairwise error probability (PEP), i.e. the probability that the received symbol is closer to x^i than x^l. At high SNRs, we can write

$$\Pr\left\{x^l\rightarrow x^{\eta(l)}\right\}\leq\Pr\{\text{error}|x^l\text{ transmitted}\}\leq\sum_{x^i\in\mathcal{A}(l)}\Pr\left\{x^l\rightarrow x^i\right\} \tag{17}$$

where $x^{\eta(l)}$ denotes the nearest neighbor to symbol x^l and $\mathcal{A}(l)\subset\mathcal{A}$ is the set of nearest neighbors of x^l.

$$\Pr\{x^l\rightarrow x^i|r_{11},r_{22},\ldots,r_{N_tN_t}\}=Q\left(\frac{|u^l-u^i|}{2}\right) \tag{18}$$

where $Q(\cdot)$ is the Gaussian Q-function and,

$$u^l=\sqrt{P/N_t}[r_{11}x_1^l\ r_{22}x_2^l\ldots\ r_{N_tN_t}x_{N_t}^l]$$
$$u^i=\sqrt{P/N_t}[r_{11}x_1^i\ r_{22}x_2^i\ r_{33}x_3^i\ldots\ r_{N_tN_t}x_{N_t}^i]$$

x_k^l and x_k^i denote the k^{th} ordinate of the symbols x^l and x^i respectively.

$$|u^l-u^i|= \tag{19}$$
$$\sqrt{P/N_t(r_{11}^2(x_1^l-x_1^i)^2+r_{22}^2(x_2^l-x_2^i)^2+\ldots r_{N_tN_t}^2(x_{N_t}^l-x_{N_t}^i)^2)} \tag{20}$$

As the constellation is rotated such that no two symbols lie on the same co-ordinates along any dimension, $x_k^l-x_k^i\neq 0, k=1,2,\ldots,N_t$ for any $x^l,\ x^i\in\mathcal{A}$.

$$\Rightarrow|u^l-u^i|=\sqrt{(P(r_{11}^2c_1+r_{22}^2c_2+\ldots+r_{N_tN_t}^2c_{N_t}))} \tag{21}$$

where, $c_1, c_2, \ldots, c_{N_t}$ are constants that wont affect the diversity gain. As P is the average SNR at each of the receive antenna,

$$|u^l - u^i| = \sqrt{\mathsf{SNR}(r_{11}^2 c_1 + r_{22}^2 c_2 + \ldots + r_{N_t N_t}^2 c_{N_t})} \quad (22)$$

To evaluate (18), we need $f(r_{11}, r_{22}, \ldots, r_{N_t N_t})$, the joint pdf of $r_{11}, r_{22}, \ldots, r_{N_t N_t}$. With greedy QR decomposition, since it is difficult of obtain the pdfs $f(r_{ii})$ and joint pdfs involving r_{ii},we use the bounds derived in [9].

$$\frac{\sum_{j=1}^{N_t} \lambda_j^2}{N_t - j + 1} \leq r_{11}^2 \leq \lambda_1^2 \prod_{j=1}^{i-1} N_t - j + 1, \quad i = 1, 2, \ldots, N_t \quad (23)$$

Here, $\lambda_1 \geq \lambda_2 \ldots \lambda_{N_t} > 0$ are the non-zero singular values of the channel matrix $\mathbf{H}$. By substituting the upper bounds on r_{ii} into (21),

$$|u^l - u^i| \leq \quad (24)$$

$$\sqrt{\mathsf{SNR}(\lambda_1^2 c_1 + \lambda_2^2 N_t c_2 + \ldots + \lambda_2^2 (N_t)(N_t - 1)\ldots(2)(1) c_{N_t})} \quad (25)$$

As $\lambda_1^2 \geq \lambda_2^2 \geq \ldots \geq \lambda_N^2 > 0$, we have the following upper bound on $|u^l - u^i|$

$$|u^l - u^i| \leq \quad (26)$$

$$\sqrt{\mathsf{SNR}\lambda_1^2 (c_1 + N_t c_2 + \ldots + N_t(N_t - 1)\ldots(2)(1) c_{N_t})} \quad (27)$$

Similarly, we can obtain a lower bound on $|u^l - u^i|$ as

$$|u^l - u^i| > \sqrt{\mathsf{SNR}\lambda_1^2 \left(\frac{c_1}{N_t} + \frac{c_2}{N_t - 1} + \ldots + \frac{c_{N_t}}{1} \right)} \quad (28)$$

Now, by substituting these bounds on $|u^l - u^i|$ into (18),

$$E_{\lambda_1^2}\left[Q\left(\sqrt{\mathsf{SNR}\lambda_1^2 c_1'} \right) \right] \leq \Pr\left\{ x^l \to x^i \right\} < E_{\lambda_1^2}\left[Q\left(\sqrt{\mathsf{SNR}\lambda_1^2 c_2'} \right) \right] \quad (29)$$

where c_1' and c_2' are appropriately defined constants. The near zero behavior of $f(\lambda_k^2)$, marginal pdf of k^{th} largest eigen value of Wishart matrix $\mathbf{W}$ has been characterized in [10] and was shown that, as $\lambda_k^2 \to 0^+$,

$$f(\lambda_k) = a_k (\lambda_k^2)^{d_k} + o\left(\lambda_k^{2d_k} \right) \qquad k = 1, \ldots, n \quad (30)$$

where $d_k = (m - k + 1)(n - k + 1) - 1,\ m = \max\{N_t, N_r\}$ and $n = \min\{N_t, N_r\}$. Wang *et. al.* [11] has shown that the diversity gain depends on depends on the behavior of the distribution of channel gain near zero and using proposition 1 from [11], along with (30), it can be shown that

$$\Pr\left\{ x^l \to x^i \right\} = C_{li} \mathsf{SNR}^{-(d_1 + 1)} \quad (31)$$

where C_{li} is the coding gain. Observe that the SNR exponent of the PEP do not depend on the particular pair of symbols being considered. It implies that,

$$\Pr\{\text{error}|x^l \text{ transmitted}\} = C\mathsf{SNR}^{-(d_1+1)} \quad (32)$$

and results in,

$$P_{s,1} = \sum_{x^l \in \mathcal{A}} \Pr\{\text{error}|x^l \text{ transmitted}\} \Pr\{x^l \text{ transmitted}\} = K_1 \mathsf{SNR}^{-(d_1+1)} \quad (33)$$

where K_1 is a constant that determines the coding gain (or, equivalently, the array gain) of the SEP. SEP of the k^{th} data stream can be calculated in a similar way to show that $P_{s,k} = K_k \mathsf{SNR}^{-(d_1+1)}$. As $d_1 + 1 = N_r \times N_t$, the average SEP of MD CISM is then given by

$$P_s = K\mathsf{SNR}^{-N_t N_r} \quad (34)$$

where K is represents the coding gain. Thus we obtain the following result:

All the data streams in MD CISM achieve maximum diversity gain of $N_t Nr$ in a $N_r \times N_t$ MIMO channel.

VI. Simulation results

In this section, we evaluate the BER performance of the proposed scheme for a 4×4 MIMO system through simulations and compare it with that of ML detection and ordered MMSE-VBLAST which require no feedback, UCD and GMD with full feedback .

Rotated 4-dimensional QAM symbols are generated as explained but grouped together eventually encoding it over two dimensions. $\mathbf{H}$ is assumed to be constant over a block length of T symbols and varying independently from block to block. The real and imaginary parts of the channel coefficients are i.i.d Gaussian random variables with zero mean and unit variance. SNR is defined as $E(\mathbf{x}^{\mathbf{H}}\mathbf{x})/\sigma_n^2$, where σ_n^2 is the variance of the noise and $\mathbf{x}$ is the transmitted symbol vector. The simulation is performed by averaging over many different channel and noise realizations.

Figure 2 shows that the slope of the BER curve of MMSE receiver corresponds to first order diversity. ML and GMD achieve a diversity order of 4. UCD, with full CSIT, has the best performance among all the schemes but it requires full feedback of 16 complex numbers. MD CISM has the same diversity order of 16 as UCD but the feedback required is only 5 bits as compared to $16 * 2 * N$ bits required by UCD, where $2 * N$ is the number of bits used to represent the real and imaginary parts of one channel coefficient. The curve generated in Figure 2 is generated by using $N = 8$, i.e. a feedback of 256 bits.

VII. Conclusions

In this paper, we have proposed a novel spatial multiplexing scheme for a flat fading MIMO system by encoding information over N_t dimensions, interleaving their co-ordinates over all the layers and transmitting them diagonally along the space-time grid. Further, the channel is decomposed using greedy

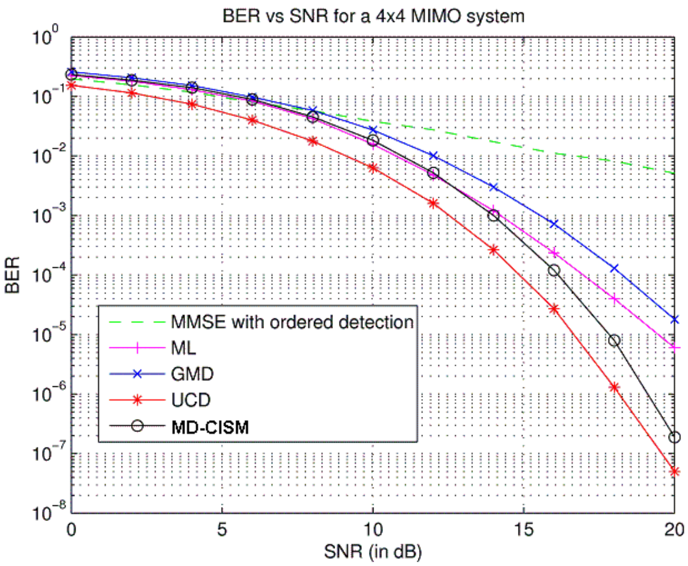

Fig. 3. Comparison of BER vs SNR performance of various transceiver schemes for a 4×4 MIMO system

QR decomposition, hence, only the permutation matrix needs to be fed back to the transmitter which amounts to a feedback of $log_2 N_t!$ bits. We have proved analytically that the proposed scheme achieves full diversity of $N_t N_r$ and supported it by Monte Carlo simulations. When compared to other closed loop diversity schemes like GMD and UCD, the proposed scheme achieves superior performance with very less feedback.

This scheme can be easily extended to a frequency selective MIMO channel by using it in combination with OFDM.

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