Precoded V-BLAST for ISI MIMO channels

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***Abstract*— The V-BLAST (vertical Bell labs layered space-time) system is one of the MIMO systems designed to achieve a good multiplexing gain. In the recent literature, a V-BLAST precoder has been added in the transmitter which exploits channel information. This precoder forces each symbol stream to have identical MSE (mean square error). It can be viewed as an alternative to the bitloading method. In this paper, this precoded V-BLAST is extended to the case of ISI MIMO channels. Both the FIR and OFDM types of transceivers are derived.**[1]

I. INTRODUCTION

In a MIMO communication system, multiple transmission paths can be used to improve diversity and/or multiplexing gain. The V-BLAST (vertical Bell labs layered space-time) system suggested in [4] is one of the MIMO systems designed to achieve good multiplexing gain. In the V-BLAST transmitter, every antenna transmits its own independently coded symbol. In the V-BLAST receiver, a spatial domain decision feedback equalizer is used. One by one the symbols are decoded and fedback to cancel its interference with other symbols (Fig. 1). This process repeats until all of the symbols are decoded. The decoding order can be optimized by decoding the symbol with largest signal to noise ratio first. Due to this decision feedback structure, the V-BLAST system has a very good spectral efficiency in a scattering rich environment [4]. Recently, in [2] and [3], by exploiting channel information at the transmitter, an optimal linear precoder has been added in the V-BLAST transmitter. This linear precoder contains two parts. First, it performs optimal power loading. This part is the same as the SVD diversity techniques [9] which decompose the channel matrix using singular value decomposition (SVD) and use these decomposed unitary matrices as linear precoders. Then different power is allocated on each eigenmode of the channel matrix. The second part is a new idea. It is a linear precoder that makes the MSE of each symbol stream $s(k)$ identical for all k so that the equalized MSE becomes the geometrical mean of the original MSEs. It can be viewed as an alternative to the bitloading method.

In this paper, we extend the MSE-equalizing precoders and the V-BLAST from block MIMO channels to ISI MIMO channels. The paper is organized as follows. In Sections II and III, we review the V-BLAST system and the MSE-equalizing precoder. In Sections IV and V, we extend the system to the case of ISI MIMO channels and consider both the FIR and the MIMO-OFDM transceivers. Section VI shows the simulation results for different systems on ISI channels. For convenience, MATLAB index notations for matrices will be used throughout this paper. $\mathbf{H}(:, k:\text{end})$ denotes the kth though the last column of $\mathbf{H}$. Superscript * denotes transpose conjugation.

II. REVIEW OF THE V-BLAST SYSTEM

The V-BLAST system is a decision feedback equalizer in the spatial domain. Fig. 1 shows the V-BLAST scheme, where $\mathbf{s}$ is the $M \times 1$ transmitted signal, $\mathbf{H}$ is an $N \times M$ channel matrix, $\mathbf{v}$ is the channel noise and $\mathbf{x}^{(M)} = \mathbf{Hs} + \mathbf{v}$ is the $N \times 1$ received signal. We assume $\mathbf{s}$ and $\mathbf{v}$ are zero mean

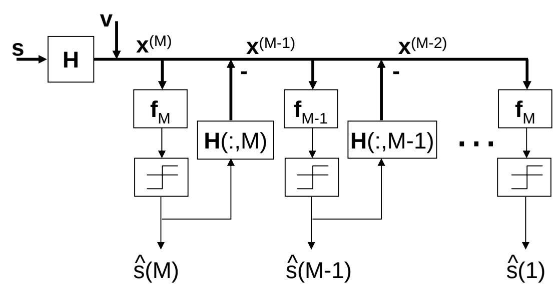

Fig. 1. The V-BLAST scheme.

and independent and $E[\mathbf{ss}^*] = \frac{1}{\alpha}\mathbf{I}_M$, $E[\mathbf{vv}^*] = \mathbf{I}_N$, where α is the noise to signal ratio. In Fig. 1, $\mathbf{f}_k$ is an $1 \times N$ vector such that $\mathbf{f}_k\mathbf{x}^{(k)}$ is the LMMSE (Linear Minimum Mean Square Estimator) of the kth element of $\mathbf{s}$, $s(k)$ based on the input $\mathbf{x}^{(k)}$ (i.e., $\mathbf{f}_k$ minimizes $E|\mathbf{f}_k\mathbf{x}^{(k)} - s(k)|^2$.) After this estimation, $\mathbf{f}_k\mathbf{x}^{(k)}$ is sent to the slicer and after the decision, it is fedback to cancel the interference caused by $s(k)$. That is $\mathbf{x}^{(k-1)} = \mathbf{x}^{(k)} - \mathbf{H}(:, k)\hat{s}(k)$, where $\hat{s}(k) = C\left[\mathbf{f}_k\mathbf{x}^{(k)}\right]$ and $C[\cdot]$ denotes the decision. As in many analyses of decision feedback systems, we assume there is no error propagation. That is, when considering the kth input $\mathbf{x}^{(k)}$, the previous decisions $\hat{s}(i)$, $i = M, M-1, \cdots, k+1$ are correct. Therefore,

$$\begin{aligned}\mathbf{x}^{(k)} &= \mathbf{x}^{(M)} - \sum_{i=k+1}^{M} \mathbf{H}(:, i)s(i) \\ &= \underbrace{\mathbf{H}(:, 1:k)}_{\text{Call this } \mathbf{A}_k.} \mathbf{s}(1:k) + \mathbf{v}.\end{aligned}$$

From the above equation, one can derive $\mathbf{f}_k$ and the corresponding MSE by the orthogonality principle and obtain

$$\mathbf{f}_k = \left((\mathbf{A}_k^*\mathbf{A}_k + \alpha\mathbf{I}_k)^{-1}\mathbf{A}_k^*\right)(k, :) \tag{1}$$

$$E\left|s(k) - \mathbf{f}_k\mathbf{x}^{(k)}\right|^2 = (\mathbf{A}_k^*\mathbf{A}_k + \alpha\mathbf{I}_k)^{-1}(k, k) \tag{2}$$

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A fast way to compute these is by using the QR algorithm [1]. We first compute the following QR decomposition:

$$\begin{pmatrix} \mathbf{H} \\ \sqrt{\alpha}\mathbf{I}_M \end{pmatrix} = \mathbf{QR}, \quad (3)$$

where $\mathbf{Q}$ is an $(N+M)\times M$ unitary matrix and $\mathbf{R}$ is $M\times M$ and upper triangular. Observing the first k columns of the equation above, we obtain

$$\underbrace{\begin{pmatrix} \mathbf{H}(:,1:k) \\ \sqrt{\alpha}\mathbf{I}_k \\ \mathbf{0}_{(M-k)\times M} \end{pmatrix}}_{\text{Call this } \mathbf{H}_k.} = \mathbf{Q}\begin{pmatrix} \mathbf{R}(1:k,1:k) \\ \mathbf{0}_{(M-k)\times k} \end{pmatrix}.$$

The solution of $\mathbf{f}_k$ in Eq. (1) can be rewritten as $\mathbf{f}_k = \left((\mathbf{H}_k^*\mathbf{H}_k)^{-1}\mathbf{H}_k^*\right)(k,1:N)$. Substituting the QR expression of $\mathbf{H}_k$ above, we obtain

$$\begin{aligned} \mathbf{f}_k &= \left(\mathbf{R}(1:k,1:k)^{-1}\mathbf{Q}(1:N,:)^*\right)(k,:) \\ &= R(k,k)^{-1}\mathbf{Q}(1:N,k)^*. \end{aligned} \quad (4)$$

The last equality comes from the fact that $\mathbf{R}(1:k,1:k)$ is an upper triangular matrix. One can obtain the corresponding MSE in Eq. (2) by a similar method and obtain

$$E\left|s(k)-\mathbf{f}_k\mathbf{x}^{(k)}\right|^2 = |R(k,k)|^{-2}. \quad (5)$$

Thus, by computing one QR decomposition in Eq. (3), all of the LMMSE vectors can be obtained by Eq. (4) and the MSE can be obtained by Eq. (5). The decoding order can be optimized by the following variation of the QR algorithm $\begin{pmatrix} \mathbf{H} \\ \sqrt{\alpha}\mathbf{I}_M \end{pmatrix} = \mathbf{QRE}$, where $\mathbf{E}$ is a permutation chosen so that $|R(k,k)|^2$ is decreasing. This reduces error propagation.

III. Review of the UCD system

The UCD (Uniform Channel Decomposition) system [2][3] is a V-BLAST system with an optimal precoder derived from channel information. The optimal precoder contains two parts. The first part performs optimal power loading. It uses a water-filling algorithm to allocate the transmitted power to each eigenmode of the channel matrix. The second part is a unitary matrix which equalizes the MSE of all the elements to be their geometrical mean. In this paper, we focus on the second part. Power loading can be obtained independently from this part.

Instead of computing the QR decomposition in Eq. (3), the following GMD (Geometrical Mean Decomposition) introduced in [5] is performed first.

$$\begin{pmatrix} \mathbf{H} \\ \sqrt{\alpha}\mathbf{I}_M \end{pmatrix} = \mathbf{Q}_1\mathbf{R}_1\mathbf{P}^*, \quad (6)$$

where $\mathbf{Q}_1$ and $\mathbf{P}$ are unitary matrices and $\mathbf{R}_1$ is $M\times M$ and upper triangular. Furthermore, $R_1(k,k)$ is a constant for all k and it can be expressed as

$$R_1(k,k) = \det(\mathbf{H}^*\mathbf{H}+\alpha\mathbf{I}_M)^{\frac{1}{2M}} = \left(\prod_{i=1}^{M} R(i,i)\right)^{\frac{1}{M}},$$

for all k, where $\mathbf{R}$ is defined in Eq. (3). By this decomposition, $R_1(k,k)$ equals the geometrical mean of $\{R(i,i)\}$, hence the name GMD. Now, let $\mathbf{P}$ be the precoder. Then the equivalent channel becomes $\mathbf{HP}$. Substituting the equivalent channel and using the GMD in Eq. (6), the QR decomposition corresponding to Eq. (3) can be obtained by

$$\begin{aligned} \begin{pmatrix} \mathbf{HP} \\ \sqrt{\alpha}\mathbf{I}_M \end{pmatrix} &= \begin{pmatrix} \mathbf{I}_N & \mathbf{0}_{N\times M} \\ \mathbf{0}_{M\times N} & \mathbf{P}^* \end{pmatrix}\begin{pmatrix} \mathbf{H} \\ \sqrt{\alpha}\mathbf{I}_M \end{pmatrix}\mathbf{P} \\ &= \underbrace{\begin{pmatrix} \mathbf{I}_N & \mathbf{0}_{N\times M} \\ \mathbf{0}_{M\times N} & \mathbf{P}^* \end{pmatrix}\mathbf{Q}_1}_{\text{unitary}}\mathbf{R}_1\underbrace{\mathbf{P}^*\mathbf{P}}_{\mathbf{I}_M}. \end{aligned}$$

Thus, by Eq. (5), the MSE of the kth element becomes

$$E\left|\mathbf{f}_k\mathbf{x}^{(k)}-s(k)\right|^2 = R_1(k,k)^{-2} = \det(\mathbf{H}^*\mathbf{H}+\alpha\mathbf{I}_M)^{-\frac{1}{M}},$$

for all k. It equals the geometrical mean of the MSEs of all elements in the original V-BLAST system. In [2], the authors point out that the UCD transceiver is capacity lossless. That is, the sum of the capacities of the individual channels equals the capacity of the MIMO channel.

Comparison of the MSE-equalizing precoder and bitloading. Bitloading is a technique that uses different sizes of constellation among parallel subchannels so that the BERs among all subchannels are approximately the same. The MSE-equalizing precoder also results in approximately the same BERs among all subchannels. Thus, the MSE-equalizing precoder can be viewed as an alternative to the bitloading algorithm. Moreover, under the same bit transmission rate, the MSE-equalizing precoder performs better in BER than the bitloading. This is because the bitloading is discrete. Practically the subchannel BERs can never be the same by performing bitloading. Some subchannels would have higher BER. This degrades the total BER performance. On the other hand, linear precoders are continuous, therefore the MSE can be exactly the same among all symbol streams.

IV. Precoded FIR V-BLAST for ISI MIMO channels

In this section, the V-BLAST and the UCD are generalized to the case of the ISI MIMO channel. The FIR MMSE DFE for the ISI MIMO channel and the corresponding precoder which results in identical MSE in every transmitted symbol stream are derived. The FIR MMSE DFE for the ISI MIMO channel was introduced in [6] and [7]. We briefly derive it again in a simpler way. The transmitted and received signal of the ISI-MIMO channel can be expressed as $\mathbf{x}_i = \sum_{n=0}^{L}\mathbf{h}_n\mathbf{s}_{i-n}+\mathbf{v}_i$, where $\mathbf{s}_i$ is the $M\times 1$ transmitted signal, $\mathbf{H}(z) = \mathbf{h}_0+\mathbf{h}_1z^{-1}+\cdots+\mathbf{h}_Lz^{-L}$ is the Lth order $N\times M$ ISI MIMO channel, $\mathbf{v}_i$ is the $N\times 1$ channel noise, and $\mathbf{x}_i$ is the $N\times 1$ received signal. At time i, a dth order FIR DFE decodes $\mathbf{s}_{i-\Delta}$ based on the observed received signals $\mathbf{x}_i, \mathbf{x}_{i-1}, \cdots, \mathbf{x}_{i-d}$ along with the previous decoded signals $\{\mathbf{s}_j\}_{j<i-\Delta}$, where Δ is the decision

delay. The observed received signals can be expressed by

$$\begin{pmatrix} \mathbf{x}_i \\ \mathbf{x}_{i-1} \\ \vdots \\ \mathbf{x}_{i-d} \end{pmatrix} = \mathbf{H}_T \begin{pmatrix} \mathbf{s}_i \\ \mathbf{s}_i - 1 \\ \vdots \\ \mathbf{s}_{i-d-L} \end{pmatrix} + \underbrace{\begin{pmatrix} \mathbf{v}_i \\ \mathbf{v}_{i-1} \\ \vdots \\ \mathbf{v}_{i-d} \end{pmatrix}}_{\text{Call this } \mathbf{v}_T},$$

where $\mathbf{H}_T$ is a block Toeplitz matrix with $\mathbf{H}_T(1:N,:) = \begin{pmatrix} \mathbf{h}_0 & \mathbf{h}_1 & \cdots & \mathbf{h}_L & \mathbf{0}_{N\times(d+1)M} \end{pmatrix}$ and $\mathbf{H}_T(:,1:M) = \begin{pmatrix} \mathbf{h}_0^T & \mathbf{0}_{M\times Nd} \end{pmatrix}^T$. One can move all the available information to the left hand side and obtain

$$\underbrace{\begin{pmatrix} \mathbf{x}_i \\ \mathbf{x}_{i-1} \\ \vdots \\ \mathbf{x}_{i-d} \end{pmatrix} - \mathbf{H}_{T2} \begin{pmatrix} \mathbf{s}_{i-\Delta-1} \\ \mathbf{s}_{i-\Delta-2} \\ \vdots \\ \mathbf{s}_{i-d-L} \end{pmatrix}}_{\text{Call this } \mathbf{x}_T.} = \mathbf{H}_{T1} \underbrace{\begin{pmatrix} \mathbf{s}_i \\ \mathbf{s}_{i-1} \\ \vdots \\ \mathbf{s}_{i-\Delta} \end{pmatrix}}_{\text{Call this } \mathbf{s}_T.} + \mathbf{v}_T$$

where $\mathbf{H}_{T1} = \mathbf{H}_T(:,1:(\Delta+1)M)$ and $\mathbf{H}_{T2} = \mathbf{H}_T(:,M(\Delta+1)+1:\text{end})$. Because $\mathbf{x}_T$ is known, it reduces to the block channel V-BLAST except only the last M elements of $\mathbf{s}_T$, namely $\mathbf{s}_{i-\Delta}$ need to be decoded, instead of decoding the whole vector. This can be done by the same scheme in Fig. 1 but keeping only the first M LMMSE estimators. The LMMSEs can be obtained from the following variation of the QR decomposition:

$$\begin{pmatrix} \mathbf{H}_{T1} \\ \sqrt{\alpha}\mathbf{I}_{(\Delta+1)M} \end{pmatrix} = \mathbf{Q} \begin{pmatrix} & \times \\ \mathbf{0}_{M\times\Delta M} & \mathbf{R} \end{pmatrix}, \quad (7)$$

where $\times$ denotes the irrelevant terms, $\mathbf{Q}$ is unitary and $\mathbf{R}$ is upper triangular. Because only the last M elements need to be decoded, only the last M rows need to be upper triangular. By Eq. (5), the MSE $E|\hat{s}_{i-\Delta}(k) - s_{i-\Delta}(k)|^2 = |R(k,k)|^{-2}$.

We now focus on the corresponding MSE-equalizing precoder. It can be obtained by the following GMD: $\mathbf{R} = \mathbf{Q}_1\mathbf{R}_1\mathbf{P}^*$, where $\mathbf{Q}_1$ and $\mathbf{P}$ are unitary, and $\mathbf{R}_1$ is an upper triangular matrix with $R_1(k,k) = (\prod_{i=1}^M R(i,i))^{\frac{1}{M}}$, for $k = 1,2,\cdots,M$. Let the unitary matrix $\mathbf{P}$ be the precoder. The equivalent channel becomes $\mathbf{H}(z)\mathbf{P}$ and the corresponding Toeplitz matrix becomes $\mathbf{H}_T\mathbf{P}_T$, where $\mathbf{P}_T$ is a block diagonal matrix equal to $\text{diag}\{\mathbf{P},\mathbf{P},\cdots,\mathbf{P}\}$. Substituting the equivalent channel and applying the above GMD, the QR decomposition which corresponds to Eq. (7) becomes

$$\begin{aligned} \begin{pmatrix} \mathbf{H}_{T1}\mathbf{P}_{T1} \\ \sqrt{\alpha}\mathbf{I}_{(\Delta+1)M} \end{pmatrix} &= \begin{pmatrix} \mathbf{I} & \\ & \mathbf{P}_{T1}^* \end{pmatrix} \begin{pmatrix} \mathbf{H}_{T1} \\ \sqrt{\alpha}\mathbf{I}_{(\Delta+1)M} \end{pmatrix} \mathbf{P}_{T1} \\ &= \begin{pmatrix} \mathbf{I} & \\ & \mathbf{P}_{T1}^* \end{pmatrix} \mathbf{Q} \begin{pmatrix} & \times \\ \mathbf{0} & \mathbf{Q}_1\mathbf{R}_1\mathbf{P}^* \end{pmatrix} \mathbf{P}_{T1} \\ &= \underbrace{\begin{pmatrix} \mathbf{I} & \\ & \mathbf{P}_{T1}^* \end{pmatrix} \mathbf{Q} \begin{pmatrix} \mathbf{I} & \\ & \mathbf{Q}_1 \end{pmatrix}}_{\text{unitary matrix}} \begin{pmatrix} & \times \\ \mathbf{0} & \mathbf{R}_1 \underbrace{\mathbf{P}^*\mathbf{P}}_{\mathbf{I}_M} \end{pmatrix} \end{aligned}$$

where $\mathbf{P}_{T1}$ is a block diagonal matrix equal to $\mathbf{P}_T(1:(\Delta+1)M, 1:(\Delta+1)M)$. Thus, by using the unitary precoder $\mathbf{P}$, the MSE becomes identical for $1 \le k \le M$:

$$E|\hat{s}_{i-\Delta}(k) - s_{i-\Delta}(k)|^2 = |R_1(k,k)|^{-2} = (\prod_{i=1}^{M} |R(i,i)|^{-2})^{\frac{1}{M}}$$

This is the geometrical mean of the original MSEs of the system without a precoder.

V. Precoded OFDM V-BLAST for ISI MIMO Channels

In the previous section, the ISI MIMO channel is equalized by the FIR DFE. A more computationally efficient way to equalize an ISI MIMO channel is by OFDM [8]. In this section, we derive the MMSE DFE and the MSE-equalizing precoder for the MIMO OFDM system. The transmitted and received signal of the MIMO OFDM system can be expressed by $\mathbf{x}_k = \mathbf{H}_k\mathbf{s}_k + \mathbf{v}_k, k = 1,2,\cdots,K$, where $\mathbf{s}_k$ is the $M\times 1$ transmitted signal on the kth carrier, $\mathbf{H}_k$ is the $N\times M$ channel of the kth carrier, $\mathbf{v}_k$ is the $N\times 1$ channel noise, $\mathbf{x}_k$ is the $N\times 1$ received signal on the kth carrier, and K is the total number of carriers. A simple way to obtain the precoded V-BLAST is by deriving it on every carrier independently. Thus, the MSEs are identical within every carrier. However, the MSEs are different between carriers. To obtain the transceivers with identical MSE among all carriers and antennas, the following GMD needs to be computed:

$$\begin{pmatrix} \mathbf{H}_d \\ \sqrt{\alpha}\mathbf{I}_{MK} \end{pmatrix} = \mathbf{QRP}^*, \quad (8)$$

where $\mathbf{H}_d = \text{diag}\{\mathbf{H}_1,\mathbf{H}_2,\cdots,\mathbf{H}_K\}$, $\mathbf{Q}$ and $\mathbf{P}$ are unitary, and $\mathbf{R}$ is an upper triangular matrix with $R(j,j) = (\prod_{i=1}^K \det(\mathbf{H}_i^*\mathbf{H}_i + \alpha\mathbf{I}_M))^{\frac{1}{KM}}$. By the same argument established in Section III, when $\mathbf{P}$ is the precoder, the LMMSEs and the MSE can be obtained by

$$\mathbf{f}_j = R(j,j)^{-1}\mathbf{Q}(1:NK,j)^*, j = 1,2,\cdots,MK,$$
$$E|\hat{s}_k(n) - s_k(n)|^2 = (\prod_{i=1}^{K} \det(\mathbf{H}_i^*\mathbf{H}_i + \alpha\mathbf{I}_M))^{-\frac{1}{KM}},$$

for all k and n.

The computation for the GMD in Eq. (8) can be reduced. Instead of directly computing it, we first permutate the rows in the following way:

$$\begin{pmatrix} \mathbf{H}_d \\ \sqrt{\alpha}\mathbf{I}_{MK} \end{pmatrix} = \mathbf{\Pi}\mathbf{H}_{\alpha d},$$

where $\mathbf{H}_{\alpha d} = \text{diag}\{\mathbf{H}_{\alpha 1},\mathbf{H}_{\alpha 2},\cdots,\mathbf{H}_{\alpha K}\}, \mathbf{H}_{\alpha k} = \begin{pmatrix} \mathbf{H}_k \\ \sqrt{\alpha}\mathbf{I}_M \end{pmatrix}$ and $\mathbf{\Pi}$ is a permutation matrix. Then the matrix becomes block diagonal. It will greatly reduce the complexity of the GMD because an SVD needs to be computed in GMD [5], and the SVD of the block diagonal matrix is much easier to compute.

The BER performance can be further improved by using the power loading precoder among all antennas and carriers. It can be obtained independently from the MSE-equalizing precoder established here.

VI. Simulation Results and Conclusions

In this section, the BER performances of the ISI MIMO systems are compared. The ISI MIMO channel matrix $\mathbf{H}(z)$ used in the simulations is 2×2 with order $L = 3$. The real and imaginary part of each coefficient are generated by i.i.d. Gaussian distribution with zero mean and unit variance. The channel noise is white Gaussian with unit variance as we described in the beginning of Section II. This setting is the same as the simulations in [6]. The simulation is done by averaging among many different channel and noise realizations.

The following ten systems are compared:

1) **FIR-DFE.** This system uses the *FIR mmse DFE* h described in Section IV with FIR order $d = 2$ and decision delay $\Delta = 1$. No precoder or bitloading is used.
2) **FIR-DFE-BL.** This system is the *FIR-DFE with Bit-Loading* . Different sizes of QAM are used according to the MSEs. The greedy algorithm is used to perform bitloading.
3) **FIR-DFE-PC.** This system is the *FIR-DFE along with the mse-equalizing PreCoder* described in Section IV.
4) **OFDM-LE.** In this system, the ISI-MIMO channel is first converted to K parallel block MIMO subchannels by MIMO-OFDM, where $K = 16$. Then each block MIMO subchannel is equalized by a *Linear mmse Equalizer*. No precoding is used.
5) **OFDM-LE-BL.** This system is the *OFDM-LE with BitLoading*. Different sizes of QAM are used according to the MSEs.
6) **OFDM-DFE.** In this system, the ISI MIMO channel is converted to K block MIMO subchannels by MIMO-OFDM, where $K = 16$. Then each block MIMO subchannel is decoded by a V-BLAST (*DFE*) receiver independently.
7) **OFDM-DFE-BL.** This system is the *OFDM-DFE with BitLoading*. Different sizes of QAM are used according to the MSEs among all antennas and carriers.
8) **OFDM-DFE-PC.** The ISI MIMO channel is first converted to K parallel block MIMO channels by MIMO-OFDM, where $K = 16$. For each block MIMO channel, an MSE-equalizing *PreCoder* and a V-BLAST receiver are used. Note that in this case, the MSEs are identical only within each carrier.
9) **OFDM-DFE-PC-BL.** Because the OFDM-DFE-PC has different MSEs between different carriers, one can use different constellation sizes on different carriers. This system is the *OFDM-DFE-PC with BitLoading*.
10) **OFDM-DFE-TPC.** In this system, the ISI MIMO channel is converted to K block MIMO subchannels by MIMO-OFDM, where $K = 16$. The DFE and the *PreCoder* equalizing the MSEs among all antennas and carriers established in Section V is applied.

The BER performance of these ten systems are compared in Fig. 2. In this comparison, the OFDM-DFE-TPC system

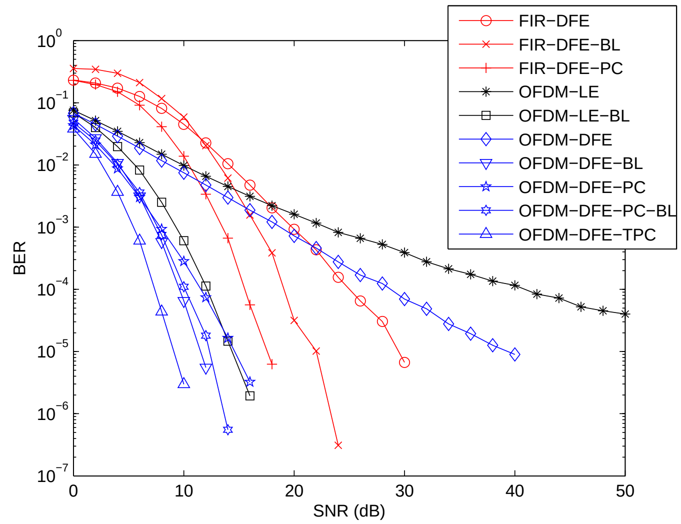

Fig. 2. BER comparison of the ten systems.

has the best performance. The OFDM-DFE-BL system has a slightly better performance than the OFDM-DFE-PC-BL system. For the FIR systems, the FIR-DFE-PC system has the best performance. The worst three systems are those without precoding or bitloading. This reveals that the MSE-equalizing precoder and bitloading are crucial to the BER performance. One can also observe that the precoder based method has a better performance than the bitloading based method as we discussed in Section III. For the equalizers, DFE is better than LE. This is because it performs signal cancellations which reduces the ISI. When we compare the OFDM and FIR systems, the FIR system suffers from endless error propagation but the OFDM systems only suffer from error propagation in one OFDM symbol duration. This makes the performance of the OFDM DFE systems better.

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