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Re:	IEEE Call for Comments on IEEE 802.16m Amendment Working Document (AWD)
Abstract	This document proposes that each MU-MIMO user may be informed about some scheduling parameters of other co-scheduled MU-MIMO users, such as their beamforming vectors, power levels and modulations. It also proposes text for the DL A-MAP IE to be included in 802.16m amendment
Purpose	To be discussed and adopted by TGm for the 802.16m amendment
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MU-MIMO: Demodulation at the UE

NEC

April 28, 2009

1 Introduction

In this report, our focus is on the MU-MIMO downlink where each scheduled user (UE) is served via codebook based precoding. We note that if precoded demodulation reference signals (DRS) for each scheduled user is employed, the precoding vectors and power levels corresponding to the co-scheduled users need not be explicitly conveyed to a user in the MU-MIMO mode. Instead, this information can be conveyed by informing each such scheduled user about the positions of the precoded DRS corresponding to some or all of the other co-scheduled users. The user can then estimate columns of its equivalent channel matrix (i.e., its observed channel matrix multiplied by a (scaled) precoding matrix) corresponding to some or all of the other co-scheduled users.

2 Demodulation at the UE

For convenience in exposition, we assume that each scheduled user is served only a single stream. Consider the received signal model on any subcarrier at a user terminal k, which is equipped with N receive antennas and where the BS has M transmit antennas,

$$\boldsymbol{y}_k = \boldsymbol{H}_k \boldsymbol{x} + \boldsymbol{v}_k, \tag{1}$$

where $\boldsymbol{H}_k \in \mathbb{C}^{N\times M}$ is the channel matrix and $\boldsymbol{v}_k \sim \mathcal{CN}(\boldsymbol{0}, \boldsymbol{I})$ is the additive noise. The signal vector $\boldsymbol{x}$ transmitted by the BS can be expanded as

$$\boldsymbol{x} = \sum_{j\in\mathcal{S}} \sqrt{p_j}\boldsymbol{g}_j s_j \tag{2}$$

where $\mathcal{S}$ is the set of users that are scheduled on the same resource unit, $\boldsymbol{g}_k$ is the precoding vector and s_k is the (unit average energy) data symbol corresponding to user $k \in \mathcal{S}$. The precoding vectors are selected from a codebook $\mathcal{C}$ of unit-norm vectors. The sum power constraint is given by $E[\boldsymbol{x}^\dagger\boldsymbol{x}] = \sum_{j=1}^{|\mathcal{S}|} p_j \le \rho$. Each scheduled user is informed about the MCS assigned to it by the BS. Further, using the precoder index or the precoded DRS assigned to it, the user can obtain an estimate of $\sqrt{p_k}\boldsymbol{H}_k\boldsymbol{g}_k$. In the sequel, we will refer to $\boldsymbol{H}_k[\sqrt{p_j}\boldsymbol{g}_j]_{j\in\mathcal{S}}$ as the equivalent channel matrix for user k.

The demodulators that are usually considered for an MU-MIMO UE are the linear combiner based demodulators, which in particular are the MRC and MMSE based demodulators. We review these demodulators in Appendix 5.1 and 5.2. Note that in order to employ the MMSE demodulator, the UE must have an estimate of the columns of its equivalent channel matrix corresponding to the other co-scheduled UEs. In the following section we consider another demodulator.

2.1 Non-linear Modulation-aware Demodulators

In order to further improve the performance, the user of interest (user-1) may employ non-linear demodulators. Suppose that the user knows the columns of its equivalent channel matrix corresponding to the other co-scheduled UEs. Moreover, the user also knows the modulations used to serve some or all of the other co-scheduled users. In particular, let $\mathcal{J} \subseteq \mathcal{S}$ denote the set of users whose corresponding modulations are known to user 1 and clearly $1 \in \mathcal{J}$. Then, user 1 can first design a filter to suppress the interference from the signals intended for the other co-scheduled users not in $\mathcal{J}$ to obtain the model

$$\boldsymbol{z} \triangleq \left(\boldsymbol{I} + \tilde{\rho}\boldsymbol{H}_1 \sum_{k\notin\mathcal{J}} \boldsymbol{g}_k\boldsymbol{g}_k^\dagger\boldsymbol{H}_1^\dagger\right)^{-1/2} \boldsymbol{y}_1 = \left(\boldsymbol{I} + \tilde{\rho}\boldsymbol{H}_1 \sum_{k\notin\mathcal{J}} \boldsymbol{g}_k\boldsymbol{g}_k^\dagger\boldsymbol{H}_1^\dagger\right)^{-1/2} \boldsymbol{H}_1\sqrt{\tilde{\rho}}\sum_{k\in\mathcal{J}} \boldsymbol{g}_k s_k + \boldsymbol{\eta}, \tag{3}$$

where $\tilde{\rho} = \frac{\rho}{|\mathcal{S}|}$ and $E[\boldsymbol{\eta}\boldsymbol{\eta}^\dagger] = \boldsymbol{I}$. Letting $\boldsymbol{B} = \left(\boldsymbol{I} + \tilde{\rho}\boldsymbol{H}_1 \sum_{k\notin\mathcal{J}} \boldsymbol{g}_k\boldsymbol{g}_k^\dagger\boldsymbol{H}_1^\dagger\right)^{-1/2} \boldsymbol{H}_1\boldsymbol{G}_{\mathcal{J}}$ with $\boldsymbol{G}_{\mathcal{J}} = [\sqrt{\tilde{\rho}}\boldsymbol{g}_k]_{k\in\mathcal{J}}$ denoting the $N \times |\mathcal{J}|$ matrix formed by the (scaled) precoding vectors employed to serve users in $\mathcal{J}$ and $\boldsymbol{s}_{\mathcal{J}}$ denoting the $|\mathcal{J}|\times 1$ vector formed by the corresponding symbols transmitted to those users, the model in (3) can be re-written as

$$\boldsymbol{z} = \boldsymbol{B}\boldsymbol{s}_{\mathcal{J}} + \boldsymbol{\eta}. \tag{4}$$

Now we can employ several suitable non-linear demodulators, such as the soft-output sphere decoder [2] and its recent lower-complexity variants over the model in (4). *The key point to be noted is that the log-likelihood ratios (LLRs) need to be generated only for the coded bits in the symbol s_1. Note that in order to generate these LLRs, we do not need any information about the coding rates employed to serve any user in $\mathcal{J}$. This is particularly useful since there are only three distinct modulations that can be used to serve each scheduled user. Also, no attempt is made to decode the codeword of any other user (apart from user 1) in $\mathcal{J}$.* A particular non-linear demodulator, referred to as the partial-MLD, is provided in Appendix-5.3.

3 Simulation results

The block error rate (BLER) performance is investigated via link-level simulations. We assume that the BS has four transmit antennas whereas the user of interest (UE-1) has two receive antennas. UE-1 reports its preferred PMI (per resource block)[1] to the BS. On a fixed set of 16 RBs, the BS then serves UE-1 using its (per RB) reported precoding vector and a

[1]The UE uses the SU-MIMO rank-1 rule to determine its preferred PMI.

Parameters	Assumption
Scenario	FDD Downlink, Linear precoding based MU-MIMO
Modulation and Coding rates	turbo codes; QPSK, 16QAM; 1/2 and 2/3
System bandwidth/FFT size/Data tones	10.0MHz/1024/600
Quantization codebook	3-bit and 4 bit
Number of antennas at BS	4
Number of antennas at UE	2
Channel estimation	Ideal
Channel model	SCM: urban macro
Linear transmit precoding	Non-unitary with and without Zero-forcing

Table 1: Simulation Parameters

fixed MCS. It also randomly picks a modulation and another precoding vector per-RB (that is near-orthogonal to the one reported by UE-1) meant for the co-scheduled user. The BLER performance obtained using the match filter (MF) based demodulator, the LMMSE based demodulator and the partial-MLD based demodulator is shown in the figures 1(a) and 1(b). Other specific simulation parameters are listed in Table 1.

In Fig. 1(a), we compare the performance obtained using the linear MF (or MRC), the linear MMSE and the non-linear partial-MLD demodulators, respectively, for QPSK modulation. The (rank-1) precoding matrix index (PMI) feedback is reported for each resource block. It is seen that the partial-MLD demodulator has much better performance than both MF and LMMSE demodulators. For example, at a BLER of 10^{-2}, the partial-MLD has 1dB and 1.5dB gains, respectively, over the LMMSE for R= 1/2 and R= 2/3.

In Fig. 1(b), we compare the performance obtained using the linear MMSE and the non-linear partial-MLD demodulators, respectively, for 16 QAM modulation. The PMI feedback is again reported for each resource block. It is seen that the partial-MLD demodulator yields an even larger gain over the LMMSE demodulator. For instance, the partial-MLD has 2dB and 3.0dB gains, respectively, over the LMMSE for R= 1/2 and R= 2/3.

In figures 2(a) and 2(b), we consider the scenario in figures 1(a) and 1(b), respectively, except that UE-1 is always scheduled on its best M sub-bands with $M = 4$ and 4 RBs per sub-band. Also, the UE is served by the BS using its (common) reported PMI on the M

sub-bands. The BS also randomly picks a modulation and another precoding vector (that is near-orthogonal to the one reported by UE-1) meant for the co-scheduled user. Note that while best-M scheduling reduces the gains, substantial gains are still obtained over the LMMSE, particularly for 16 QAM and/or higher coding rates.

Figures 3(a) and 3(b) present the 4-bit codebook versions of Figures 2(a) and 2(b), respectively. Note that while 4-bit codebook reduces the gains, substantial gains are still obtained over the LMMSE for 16 QAM and/or higher coding rates.

Finally, Fig. 4 presents the results with the 4-bit codebook, best-M scheduling and common PMI on all M sub-bands but where the BS employs (naive) zero-forcing beamforming (ZFBF) [1]. As seen from the plot, ZFBF does not provide additional residual interference mitigation and substantial gains are still obtained over the LMMSE for 16 QAM and/or higher coding rates.

4 Conclusions

It is evident that the linear combiner which requires no information about other co-scheduled users results in the least amount of (feedforward) signaling overhead. Such combiners rely on a large extent on the precoder employed by the transmitter (BS) to mitigate or remove the interference seen by them. If the BS has perfect knowledge of the channel matrix [2] of each user on each subcarrier, it can employ a zero-forcing based precoder to ensure that each scheduled user sees no interference from transmission to other co-scheduled users. In such a scenario the MRC combiner is optimal.

However, in practical FDD systems with limited feedback, providing such perfect channel knowledge about each user to the BS is not possible, particularly in the wideband OFDMA based downlink, where each user's channel response matrix varies across subcarriers. Consequently, the users may need to employ more sophisticated demodulators to combat residual

[2] As suggested in [1], each user can use a linear combiner to convert its channel matrix into an effective channel vector and report the latter to the BS.

interference and help the system realize the benefit of multi-user MIMO. These sophisticated demodulators need the knowledge about some scheduling parameters such as the modulations and the precoder indices or the positions of the precoded DRS corresponding to some or all of the other co-scheduled users. The signaling should be designed in an appropriate manner so that an MU-MIMO user can deduce such parameters of other co-scheduled MU-MIMO users.

5 Appendix

5.1 Linear Combiner: Match Filter Linear Combiner

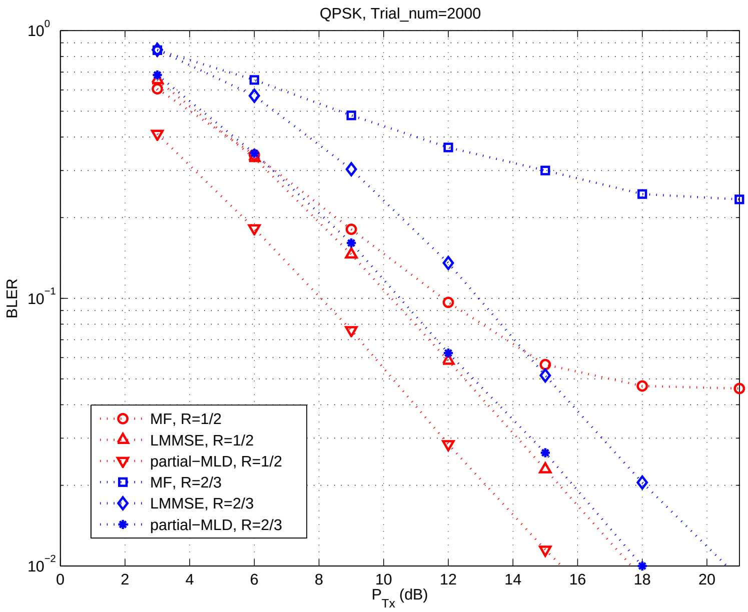
A popular linear combiner is the match filter (a.k.a. maximum ratio combiner (MRC)) which is optimal in the absence of interference from the signals intended for the other co-scheduled users. In order to employ this linear combiner, the user of interest (say user 1) does not need to know any information about the co-scheduled users and it can demodulate its own data by completely neglecting the co-channel interference resulting from the transmission intended for the other co-scheduled users. In particular, user 1 employs the unit norm combiner given by,

$$\boldsymbol{u}_1 = \frac{\boldsymbol{H}_1\boldsymbol{g}_1}{||\boldsymbol{H}_1\boldsymbol{g}_1||}. \tag{5}$$

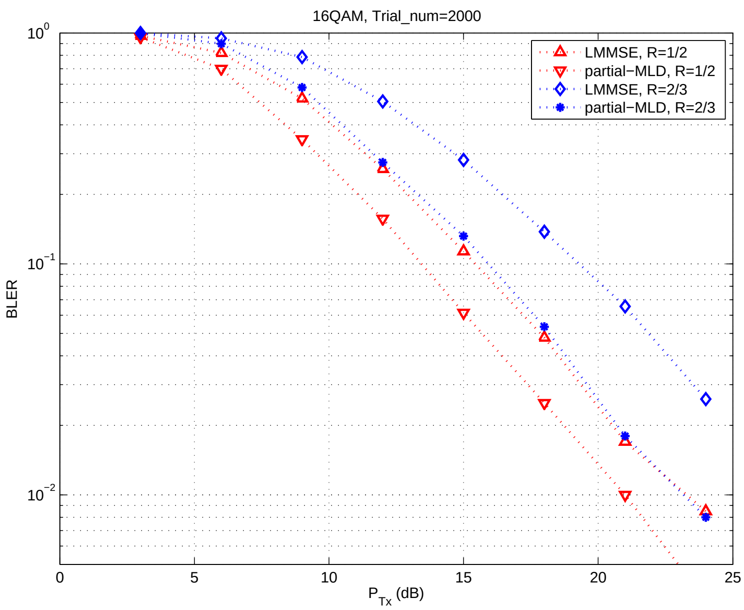
The received signal at user 1 post-combining can now be written as

$$\boldsymbol{u}_1^\dagger\boldsymbol{y}_1 = \boldsymbol{u}_1^\dagger\sqrt{p}_1\boldsymbol{H}_1\boldsymbol{g}_1 s_1 + \sum_{k\in\mathcal{S},k\neq 1}\boldsymbol{u}_1^\dagger\sqrt{p}_k\boldsymbol{H}_1\boldsymbol{g}_k s_k + \tilde{v}_1 \tag{6}$$

and note that due to our normalization $E[|\tilde{v}_1|^2] = 1$. In order to compute the log-likelihood ratios (LLRs) for the coded bits corresponding to the symbol s_1, the user can completely neglect the interference term $\sum_{k\in\mathcal{S},k\neq 1}\boldsymbol{u}_1^\dagger\sqrt{p}_k\boldsymbol{H}_1\boldsymbol{g}_k s_k$ and assume the SNR to be $p_1||\boldsymbol{H}_1\boldsymbol{g}_1||^2$. Alternatively, an estimate of the true SINR can instead determined. However, the accuracy of such an estimate depends on the number of sample observations available.

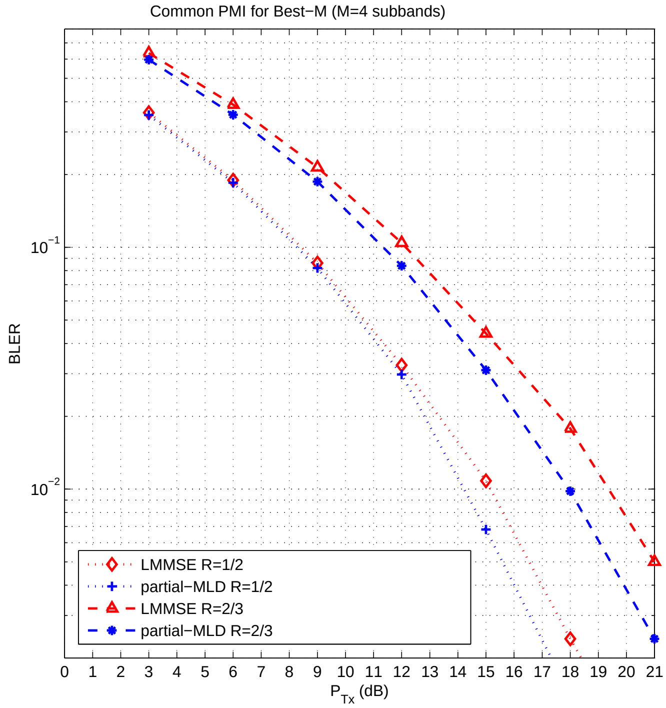


(a) Block error rate performance for QPSK

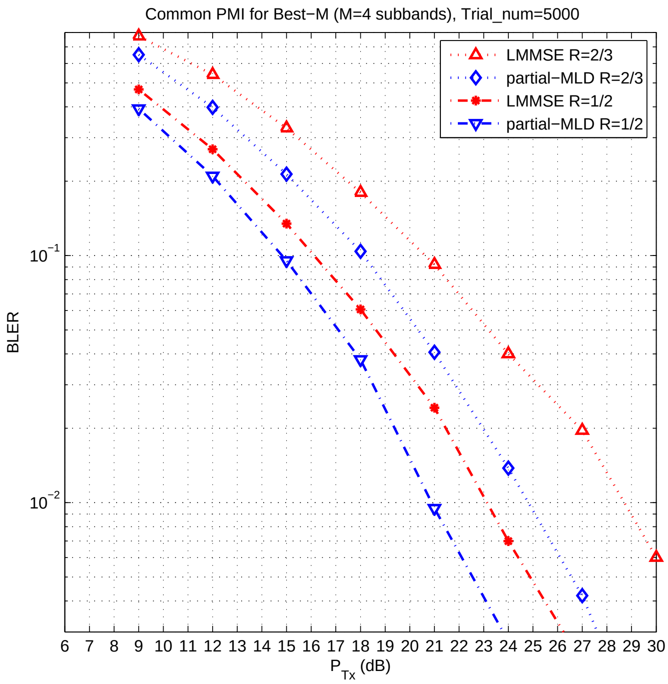


(b) Block error rate performance for 16 QAM

Figure 1: BLER performance without best-M but with independent PMI per RB.

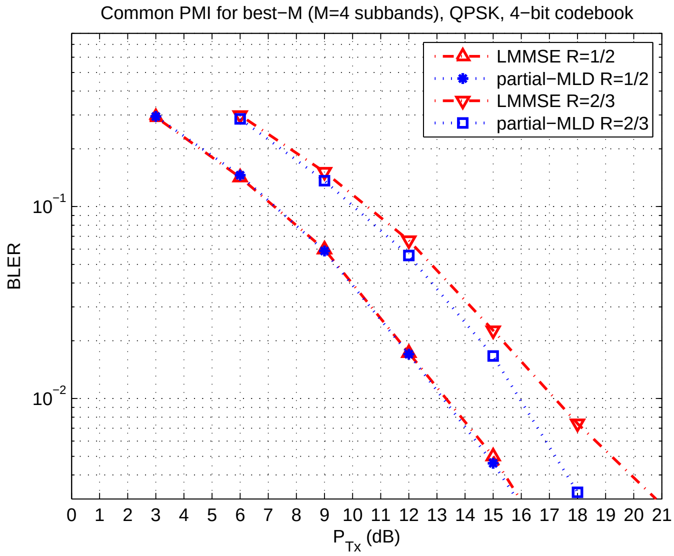


(a) Block error rate performance for QPSK

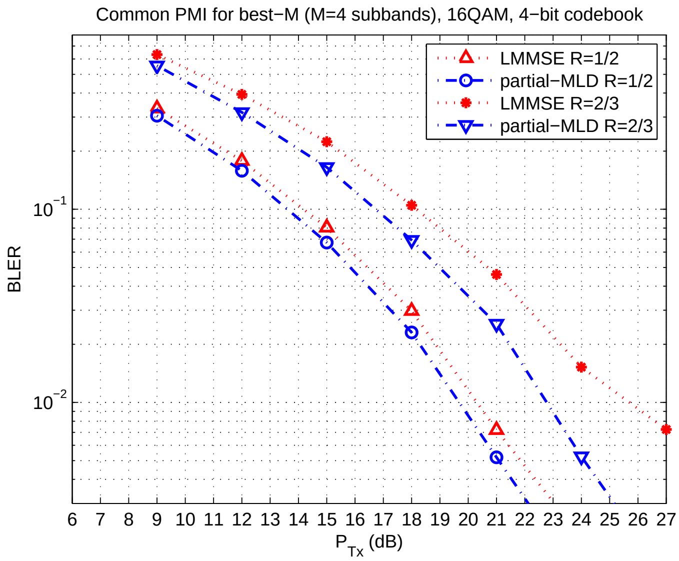


(b) Block error rate performance for 16 QAM

Figure 2: BLER performance with best-M but common PMI for best-M sub-bands.

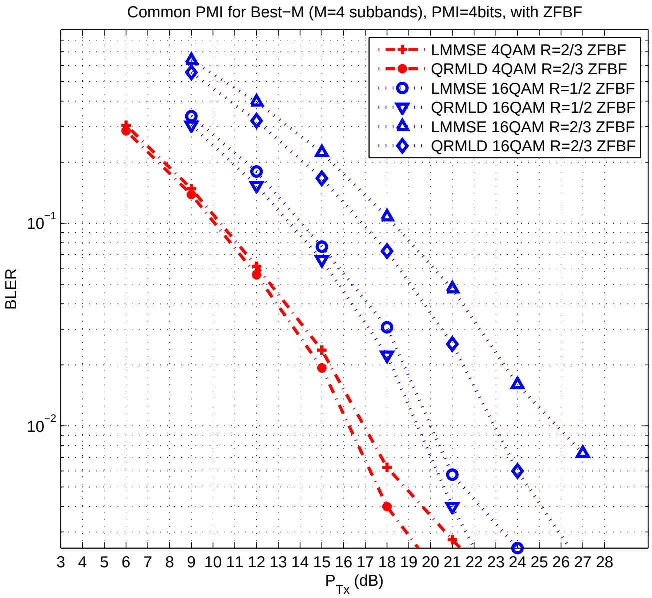


(a) Block error rate performance for QPSK



(b) Block error rate performance for 16 QAM

Figure 3: BLER performance with best-M but common PMI for best-M sub-bands and a 4-bit codebook.



(a) Block error rate performance

Figure 4: BLER performance with best-M, common PMI for best-M sub-bands, 4-bit codebook and ZFBF.

The MRC based demodulator can result in substantial performance degradation when the interference power is not negligible compared to that of the additive noise. This is particularly true at high SNR. In order design a better linear combiner for such scenarios, a quantization error minimization approach is employed in [1]. In particular, assuming $N < M$, the linear combiner used by the UE is now given by

$$\boldsymbol{u}_1 = \frac{(\boldsymbol{g}_1^\dagger \boldsymbol{H}_1^+)^\dagger}{||\boldsymbol{g}_1^\dagger \boldsymbol{H}_1^+||} \tag{7}$$

where $\boldsymbol{H}_1^+ = \boldsymbol{H}_1^\dagger(\boldsymbol{H}_1\boldsymbol{H}_1^\dagger)^{-1}$ denotes the pseudo-inverse. In general, the user can use any linear combiner $\boldsymbol{u}_1$ (which could have been pre-computed based only on the channel estimate $\boldsymbol{H}_1$). Upon using such a combiner, the resulting model is given by (6) and the LLRs can either be computed by completely neglecting the interference and using the SNR $p_1||\boldsymbol{u}_1^\dagger \boldsymbol{H}_1 \boldsymbol{g}_1||^2$ or by using an estimated SINR.

5.2 Linear Combiner: MMSE Linear Combiner

In some scenarios, each scheduled user can estimate the columns of its equivalent channel matrix corresponding to the other co-scheduled UEs. In this case, assuming that the transmit power is equally split among all scheduled users, user 1 can determine the optimal minimum mean squared error (MMSE) linear combiner as

$$\boldsymbol{u}_1 = \left(\boldsymbol{I} + \tilde{\rho} \sum_{k \in \mathcal{S}, k \neq 1} \boldsymbol{H}_1 \boldsymbol{g}_k \boldsymbol{g}_k^\dagger \boldsymbol{H}_1^\dagger \right)^{-1} \boldsymbol{H}_1 \boldsymbol{g}_1, \tag{8}$$

where $\tilde{\rho} = \frac{\rho}{|\mathcal{S}|}$. The resulting SINR is determined to be

$$\tilde{\rho} \boldsymbol{g}_1^\dagger \boldsymbol{H}_1^\dagger \left(\boldsymbol{I} + \tilde{\rho} \sum_{k \in \mathcal{S}, k \neq 1} \boldsymbol{H}_1 \boldsymbol{g}_k \boldsymbol{g}_k^\dagger \boldsymbol{H}_1^\dagger \right)^{-1} \boldsymbol{H}_1 \boldsymbol{g}_1. \tag{9}$$

5.3 Partial-MLD

We next illustrate one non-linear demodulator, referred to as the partial-MLD. Suppose that $\mathcal{J} = \{1, 2\}$ and that user-1 has two receive antennas so that $\boldsymbol{B} = [\boldsymbol{b}_1, \boldsymbol{b}_2]$ is a 2×2 matrix. Denote $\widetilde{\boldsymbol{b}}_i \triangleq \boldsymbol{b}_i / \|\boldsymbol{b}_i\|, i = 1, 2$, where $\|\boldsymbol{b}_i\| = \sqrt{\boldsymbol{b}_i^\dagger \boldsymbol{b}_i}$. Let $\boldsymbol{B} = \|\boldsymbol{b}_2\| \boldsymbol{U} \boldsymbol{L}$ be the modified QR decomposition of $\boldsymbol{B}$ with $\boldsymbol{U} = [\boldsymbol{u}_1, \boldsymbol{u}_2]$ being a semi-unitary matrix such that $\boldsymbol{U}^\dagger \boldsymbol{U} = \boldsymbol{I}$ and $\boldsymbol{L}$ being lower triangular with positive diagonal elements. In particular, we have

$$\boldsymbol{u}_1 = \frac{\widetilde{\boldsymbol{b}}_1 - \rho \widetilde{\boldsymbol{b}}_2}{\sqrt{1 - |\rho|^2}}, \quad \boldsymbol{u}_2 = \widetilde{\boldsymbol{b}}_2, \quad \text{with} \ \ \rho \triangleq \widetilde{\boldsymbol{b}}_2^\dagger \widetilde{\boldsymbol{b}}_1, \tag{10}$$

$$\text{and} \qquad \boldsymbol{L} = \begin{bmatrix} l_{11} & 0 \\ l_{21} & 1 \end{bmatrix}, \quad \text{with} \ \ l_{11} = \frac{\|\boldsymbol{b}_1\|}{\|\boldsymbol{b}_2\|} \sqrt{1 - |\rho|^2}, \quad l_{21} = \frac{\|\boldsymbol{b}_1\|}{\|\boldsymbol{b}_2\|} \rho. \tag{11}$$

We then obtain

$$\boldsymbol{w} \triangleq \boldsymbol{U}^\dagger \boldsymbol{z} / \|\boldsymbol{b}_2\| = \boldsymbol{L} \boldsymbol{s}_{\mathcal{J}} + \widehat{\boldsymbol{n}}, \tag{12}$$

where $\boldsymbol{w} = [w_1, w_2]^T, \boldsymbol{s}_{\mathcal{J}} = [s_1, s_2]^T$ and $E[\widehat{\boldsymbol{n}}\widehat{\boldsymbol{n}}^{\dagger}] = \|\boldsymbol{b}_2\|^{-2}\boldsymbol{I}$. Let s_i^{R} and s_i^{I} denote the real and imaginary parts of $s_i, i = 1, 2$, respectively. For the j-th possibility of symbol s_1, denoted as $s_{1,j}$, we define the metric

$$Q(s_{1,j}) \triangleq |w_1 - l_{11}s_{1,j}|^2 + \min_{s_2 \in \mathcal{S}_2} |w_2 - l_{21}s_{1,j} - s_2|^2, \tag{13}$$

where $\mathcal{S}_2$ denotes the constellation (modulation-type) of user-2. Suppose the modulation $\mathcal{S}_1$ assigned to user-1 is M-QAM, Using (13) we determine the M metrics $\{Q(s_{1,j})\}$.

The QAM symbol s_1, corresponds to $\log_2 M$ bits, i.e., is represented by a $\log_2 M$ length bit-vector. The M metrics $\{Q(s_{1,j})\}$ defined above are sufficient to determine the max-log LLR for each bit associated with s_1. To see this, consider the $\log_2 M$ bits associated with s_1. Then letting $\lambda_{1,\ell}$ denote the max-log LLR of the ℓ-th bit $b_{1,\ell}$ associated with s_1 and assuming equal *a priori* bit probabilities, we have

$$\lambda_{1,\ell} = \|\boldsymbol{b}_2\|^2 \min_{s_{1,j} \in \mathcal{S}_1 : b_{1,\ell}=0} Q(s_{1,j}) - \|\boldsymbol{b}_2\|^2 \min_{s_{1,j} \in \mathcal{S}_1 : b_{1,\ell}=1} Q(s_{1,j}), \quad 1 \leq \ell \leq \log_2 M. \tag{14}$$

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[1] N.Jindal, "Antenna Combining for the MIMO Downlink Channel,", *IEEE Trans. Wireless Communications*, Vol. 7, No. 10, pp. 3834-3844, Oct. 2008.

[2] B. M. Hochwald and S. ten Brink, "Achieving near-capacity on a multiple-antenna channel," *IEEE Trans. Commun.*, vol. 51, pp. 389-399, Mar. 2003.