

A Simple SNR Representation Method for AMC Schemes of MIMO Systems with ML Detector

Jihoon Kim, Kyoung-Jae Lee, Chang Kyung Sung, and Inkyu Lee

Abstract—Adaptive modulation and coding (AMC) is a powerful technique to enhance the link performance by adjusting the transmission power, channel coding rates and modulation levels according to channel state information. In order to efficiently utilize the AMC scheme, an accurate signal-to-noise ratio (SNR) value is normally required for determining the AMC level. In this paper, we propose a simple method to represent the SNR values for maximum likelihood (ML) detector in multi-input multi-output (MIMO) systems. By analyzing the relation between the upper bound and the lower bound of the ML detector performance, we introduce an efficient way to determine the SNR for the ML receiver. Based on the proposed SNR representation, an AMC scheme for single antenna systems can be extended to MIMO systems with ML detector. From computer simulations, we confirm that the proposed SNR representation allows us to achieve almost the same system throughput as the optimum AMC systems in frequency selective channels with reduced complexity.

Index Terms—Multi-input multi-output (MIMO), maximum likelihood detector (MLD), adaptive modulation and coding (AMC).

I. INTRODUCTION

MOST modern wireless communication systems have been designed to support high speed packet data service over wideband channels. Such high speed data transmission becomes feasible only when sophisticated techniques are applied to combat frequency selectivity in wireless wideband channels. In frequency selective fading channels, intersymbol interference (ISI) distorts the transmitted signal severely. By adopting orthogonal frequency division multiplexing (OFDM), wideband channels with frequency selective fading are transformed into multiple flat fading channels [1]. Moreover, the OFDM system combined with bit-interleaved coded modulation (BICM) [2] [15] has been applied to a wide range of wireless standards such as the IEEE 802.16e mobile WiMAX system [3].

Adaptive modulation and coding (AMC) is a powerful technique to enhance the link performance for packet transmission systems [4]. The basic idea of the adaptive transmission is to adjust the transmission power level, channel coding rates

and/or modulation levels according to the current channel state information (CSI) such as signal-to-noise ratio (SNR) or signal-to-interference plus noise ratio (SINR). Recently, the AMC scheme is extended to multi-input multi-output (MIMO) systems for very high system throughput such as the IEEE 802.11n standard [5]. In order to utilize an AMC scheme effectively, the accurate estimation for the link performance such as bit error rate (BER) or frame error rate (FER) is required.

The exponential effective SNR mapping (EESM) method proposed in [6] achieves good link error prediction in a single-input single-output (SISO) channel, which maps the instantaneous channel state sets into a single effective SINR value. However, one of disadvantages of the EESM method is that a normalization parameter should be computed for every set of modulation and code rate. Moreover, for the case of multi-input multi-output (MIMO) systems with maximum likelihood detector (MLD), referred to as MIMO-MLD, the EESM based prediction becomes much more complicated because of joint detection in MLD. Many methods to overcome this issue have been studied to extend the EESM to MIMO-MLD [7] [8] [9]. Another method of the link performance prediction for MIMO-MLD was proposed in [10] based on the mean mutual information per bit (MMIB). However, these methods based on EESM and MMIB need several pre-determined parameters which should be optimized according to system configurations.

In this paper, we propose a simple yet effective method to represent the SNR for spatial subchannels of the MIMO-MLD. Our approach is based on the block-AMC (BL-AMC) method in [11] which does not require any pre-determined parameters to predict the link performance. In order to solve this problem, we first define an upper bound and lower bound of the MLD performance and analyze the relation between these bounds. Then, a new SNR representation for the MIMO-MLD is derived by combining those bounds, and is applied to the AMC scheme. Unlike previous approaches based on EESM and MMIB, our proposed scheme requires only a single parameter to optimize the AMC performance. From simulation results, we show that the proposed SNR representation achieves the almost identical system throughput as the optimal AMC with reduced complexity.

This paper is organized as follows: In Section II, we show a basic MIMO-OFDM system combined with the AMC scheme. Section III proposes a simple method which represents the SNR for the MIMO-MLD to utilize the AMC system. In Section IV, simulation results are provided for the MIMO-MLD with AMC. Finally we conclude this paper in Section V.

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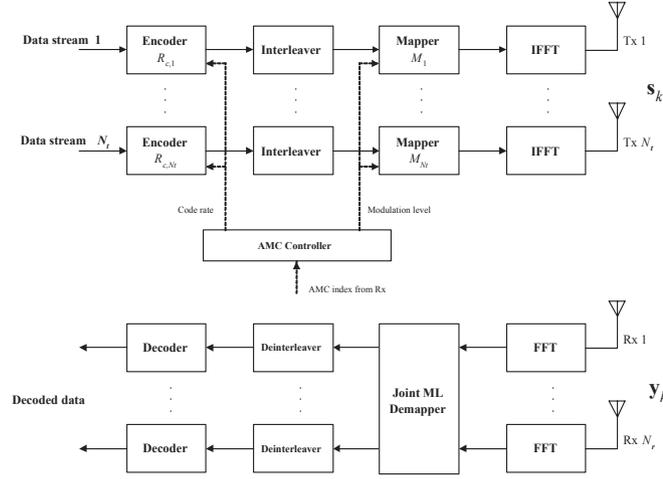


Fig. 1. MIMO-OFDM systems with ML detector.

II. SYSTEM MODEL FOR MIMO-OFDM WITH AMC

Fig. 1 illustrates the structure of MIMO-OFDM systems with N_t transmit and N_r receive antennas. At the transmitter side where the BICM structure is adopted, each data stream is encoded by convolutional codes and mapped by QAM constellation independently. Notice that there are two types of channel encoding in MIMO system: one is *vertical encoding* where all data streams are simultaneously encoded by a single channel encoder and the encoder output is then divided into several substreams. The other is *horizontal encoding* where each stream is encoded by individual encoders. In this paper, the horizontal encoding case is considered where each encoder employs the rate compatible punctured convolutional codes (RCPC) [12]. In each stream, the code rate and the modulation level are adjusted according to the feedback information. Let us denote $R_{c,i}$ and M_i as the code rate and the modulation level for the i th stream ($i = 1, \dots, N_t$), respectively. Then the AMC scheme at the receiver determines $R_{c,i}$ and M_i for each stream independently according to the current CSI, and the receiver reports the AMC indices corresponding to $R_{c,i}$ and M_i for each stream to the transmit side through the feedback channel. For OFDM operations, the mapper outputs of each stream are passed through the N_c -size inverse Fast Fourier Transform (IFFT) process and transmitted through each antenna.

At the receiver side, the received signal at the k th subcarrier after the FFT demodulation can be expressed as

$$\mathbf{y}_k = \mathbf{H}_k \mathbf{s}_k + \mathbf{w}_k \quad \text{for } k = 1, \dots, N_c \quad (1)$$

where \mathbf{y}_k represents the received signal vector of length N_r , \mathbf{s}_k denotes the transmit symbol vector with the assumption $E[\mathbf{s}_k \mathbf{s}_k^\dagger] = \sigma_s^2 \mathbf{I}_{N_t}$ and \mathbf{w}_k is an additive noise vector whose elements are circularly symmetric complex Gaussian with $E[\mathbf{w}_k \mathbf{w}_k^\dagger] = \sigma_w^2 \mathbf{I}_{N_r}$. Here we denote \mathbf{I}_K as a $K \times K$ identity matrix and $(\cdot)^\dagger$ as the conjugate transpose. In this paper, we focus on the link adaptation of open loop systems by controlling the channel coding rate and the modulation level based on the limited feedback channel. For this reason, we use the AMC index as the feedback information from the receiver. Although the link performance can be increased by applying

the precoding or the power loading in the transmit side with much larger amount of the feedback information, we do not consider such a closed-loop case in this paper. In (1), \mathbf{H}_k is the $N_r \times N_t$ channel matrix whose (j, i) th element stands for the channel coefficient between the i th transmit and the j th receive antenna at the k th subcarrier.

The expected throughput for the MIMO-MLD system, denoted by R_T , is defined as

$$R_T = \sum_{i=1}^{N_t} (1 - P_{F,i}) R_{c,i} \log_2 M_i \quad (2)$$

where $P_{F,i}$ represents the estimated FER of the i th stream. In equation (2), the accurate estimation of $P_{F,i}$ is required to predict the system throughput at the receiver for a given AMC set. In this paper, we consider the BL-AMC method proposed in [11] for estimating the FER. The BL-AMC method predicts the link error performance in SISO systems without any pre-determined parameters. For the MIMO-MLD system based on the BL-AMC, the estimation of SNRs for each stream is required. In the next section, we introduce a simple method to represent the SNR of each stream for the MIMO-MLD. After the SNR value is obtained, the BL-AMC is utilized to compute $P_{F,i}$ in (2). Then, at the receiver side, the expected throughput for all possible combinations of an AMC set is computed and the AMC level for each stream is determined which generates the largest expected throughput.

III. PROPOSED SNR REPRESENTATION METHODS FOR MIMO-MLD

In this section, we propose a simple SNR representation method for MIMO-MLD. From now on, we omit the subcarrier index k for simplicity. We start with evaluating the upper and lower bound of the MIMO-MLD performance.

First we consider the upper bound case. As is well known, the MLD outperforms any other detection structures under the system model (1). If there is no interference among symbols in the joint detection process, the detection upper bound performance is achieved. In this case, the ML detection metric reduces to

$$\begin{aligned} \arg \min_{\mathbf{s}} \|\mathbf{y} - \mathbf{H}\mathbf{s}\|^2 &= \arg \min_{\mathbf{s}} \left\{ \sum_{i=1}^{N_t} \|\mathbf{y} - \mathbf{h}_i s_i\|^2 \right\} \\ &= \bigcup_i \arg \min_{s_i} \|\mathbf{y} - \mathbf{h}_i s_i\|^2 \end{aligned}$$

where s_i indicates the i th element of the transmit symbol vector \mathbf{s} and \mathbf{h}_i denotes the i th column vector of \mathbf{H} . In other words, the orthogonal channel condition serves as an upper bound of the MIMO-MLD performance, and each symbol s_i can be detected from the maximum ratio combining (MRC) solution [13]. Therefore, in this case, the SNR of the i th stream, denoted by $\rho_{\text{MRC},i}$, becomes

$$\rho_{\text{MRC},i} = \frac{\|\mathbf{h}_i\|^2 \sigma_s^2}{\sigma_w^2} = \alpha \|\mathbf{h}_i\|^2 \quad \text{for } i = 1, \dots, N_t$$

where $\alpha = \sigma_s^2 / \sigma_w^2$. Note that $\rho_{\text{MRC},i}$ is always greater than or equal to the SNR in the MIMO-MLD since there exists a

performance loss in actual detection due to nonorthogonality in \mathbf{H} .

Next, we consider the lower bound. It is possible to mitigate the interference among symbols using a minimum mean square error (MMSE) filter \mathbf{G} as

$$\mathbf{G} = [\mathbf{g}_1 \quad \mathbf{g}_2 \quad \cdots \quad \mathbf{g}_{N_t}] = \mathbf{H} \left(\mathbf{H}^\dagger \mathbf{H} + \frac{1}{\alpha} \mathbf{I}_{N_t} \right)^{-1}$$

where \mathbf{g}_i indicates the i th column vector of \mathbf{G} . Applying \mathbf{G}^\dagger to the received symbol vector \mathbf{y} , the filter output for the i th stream, denoted by r_i , can be expressed as

$$r_i = \mathbf{g}_i^\dagger \mathbf{y} = \mathbf{g}_i^\dagger \mathbf{h}_i s_i + \sum_{j \neq i} \mathbf{g}_i^\dagger \mathbf{h}_j s_j + \mathbf{g}_i^\dagger \mathbf{w}.$$

Thus, the SINR of the i th stream for the MMSE receiver, defined by $\rho_{\text{MMSE},i}$, can be calculated as [13]

$$\begin{aligned} \rho_{\text{MMSE},i} &= \frac{|\mathbf{g}_i^\dagger \mathbf{h}_i|^2 \sigma_s^2}{\sum_{j \neq i} |\mathbf{g}_i^\dagger \mathbf{h}_j|^2 \sigma_s^2 + \|\mathbf{g}_i^\dagger\|^2 \sigma_w^2} \\ &= \frac{\alpha |\mathbf{g}_i^\dagger \mathbf{h}_i|^2}{\alpha \sum_{j \neq i} |\mathbf{g}_i^\dagger \mathbf{h}_j|^2 + \|\mathbf{g}_i^\dagger\|^2}. \end{aligned} \quad (3)$$

It is well known that the performance of an MMSE receiver is suboptimal compared to an ML receiver. Therefore, denoting $\rho_{\text{MLD},i}$ as the SNR of the i th stream for the MIMO-MLD, it follows

$$\rho_{\text{MMSE},i} \leq \rho_{\text{MLD},i} \leq \rho_{\text{MRC},i}$$

where equality holds when all columns of \mathbf{H} are orthogonal. From the above inequality, we will express $\rho_{\text{MLD},i}$ in terms of the upper bound $\rho_{\text{MRC},i}$ and the lower bound $\rho_{\text{MMSE},i}$.

In order to determine the relation among these SNRs (or SINRs), we define capacities of the i th stream for the MRC, the MMSE receiver and the MIMO-MLD as $C_{\text{MRC},i} = \log_2(1 + \rho_{\text{MRC},i})$, $C_{\text{MMSE},i} = \log_2(1 + \rho_{\text{MMSE},i})$ and $C_{\text{MLD},i} = \log_2(1 + \rho_{\text{MLD},i})$, respectively. Using these capacity expressions, we define the ratio of capacity gaps β_i as

$$\begin{aligned} \beta_i &\triangleq \frac{C_{\text{MRC},i} - C_{\text{MLD},i}}{C_{\text{MLD},i} - C_{\text{MMSE},i}} \\ &= \frac{\log_2(1 + \rho_{\text{MRC},i}) - \log_2(1 + \rho_{\text{MLD},i})}{\log_2(1 + \rho_{\text{MLD},i}) - \log_2(1 + \rho_{\text{MMSE},i})}. \end{aligned} \quad (4)$$

Here we notice that the ratio of capacity gaps is undetermined when the channel matrix \mathbf{H} becomes orthogonal.

Unfortunately, it is impossible to find β_i directly since $\rho_{\text{MLD},i}$ is unknown. Therefore, we assume that all substreams have the same ratio ($\beta_i = \beta$). Then, this ratio can be expressed by sum capacities as

$$\beta = \frac{C_{\text{MRC}} - C_{\text{MLD}}}{C_{\text{MLD}} - C_{\text{MMSE}}} \quad (5)$$

where $C_{\text{MRC}} = \sum_{i=1}^{N_t} C_{\text{MRC},i}$, $C_{\text{MLD}} = \sum_{i=1}^{N_t} C_{\text{MLD},i}$ and $C_{\text{MMSE}} = \sum_{i=1}^{N_t} C_{\text{MMSE},i}$. Additionally, in equation (5), we may replace C_{MLD} with other known quantity in order to determine the value of β . In this paper, we use the open loop capacity C_{open} , which is equal to $\log_2 \det(\mathbf{I}_{N_r} + \alpha \mathbf{H} \mathbf{H}^\dagger)$. In this case, the relation between C_{MLD} and C_{open} can be expressed as $C_{\text{MLD}} = \gamma C_{\text{open}}$, where $\gamma \leq 1$ indicates a

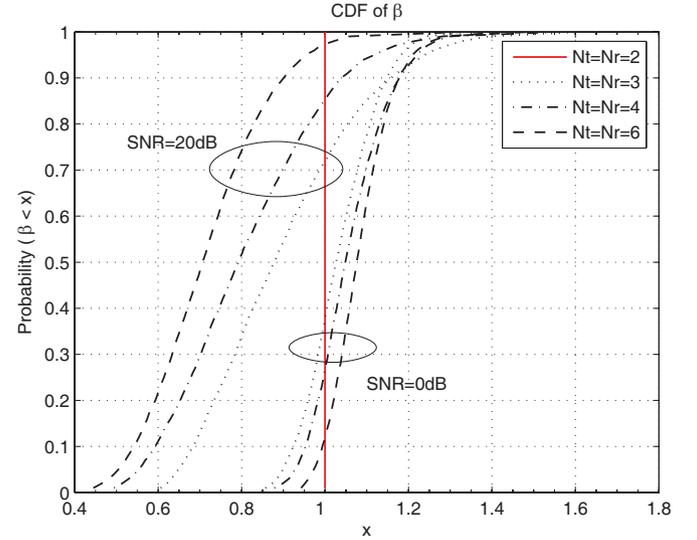


Fig. 2. The CDF of β .

parameter to determine the system optimization related to MLD with respect to the theoretical capacity. Note that, in our formulation, the capacity expression is adopted as a function to derive the parameter β regardless of its information theoretic meaning.

From the definition of β , this ratio is dependent on \mathbf{H} . Under the condition that each element in \mathbf{H} is independently Rayleigh distributed, the cumulative distribution function (CDF) of β in (5) for various configurations is depicted in Fig. 2. The following lemma states the property of β :

Lemma 1: β is always 1 regardless of the received SNR for 2×2 MIMO systems with $\gamma = 1$.

Proof: See Appendix.

From equations (4) and (5), we can obtain $(1 + \rho_{\text{MLD},i})^{1+\beta} = (1 + \rho_{\text{MRC},i}) \cdot (1 + \rho_{\text{MMSE},i})^\beta$. At high SNR, this expression reduces to

$$(\rho_{\text{MLD},i})^{1+\beta} = \rho_{\text{MRC},i} \cdot (\rho_{\text{MMSE},i})^\beta$$

or in a dB scale as

$$\rho_{\text{MLD},i} \text{ (dB)} = (1 - \eta) \cdot \rho_{\text{MRC},i} \text{ (dB)} + \eta \cdot \rho_{\text{MMSE},i} \text{ (dB)} \quad (6)$$

where $\eta = \frac{\beta}{1+\beta}$.

From lemma 1, we can conclude that the optimum η for 2×2 systems is equal to 0.5. However, in practical systems, this value may not maximize the system throughput, since employed channel codes, the constellation, the AMC set and the receiver structure may not be optimum in terms of achieving the channel capacity. In other words, η which maximizes the system throughput may not be 0.5 in practical systems. Therefore, the optimum η can simply be determined through simulations. Also, we can see from Fig. 2 that the optimum η depends on the SNR values for systems with $N_r = N_t > 2$. Thus, systems with more than two antennas should employ η as a function of SNR. However, it will be shown in the following simulation section that a good performance gain is achieved for the case with $N_t > 2$ even

TABLE I
AMC TABLE

AMC index	Spectral efficiency	Code rate	Modulation
1	1 bps/Hz	1/2	4-QAM
2	2 bps/Hz	1/2	16-QAM
3	3 bps/Hz	3/4	16-QAM
4	4 bps/Hz	2/3	64-QAM
5	5 bps/Hz	5/6	64-QAM

when a fixed value of η is used for all SNR ranges. Also, it should be noted that η is the only parameter to optimize in our proposed scheme regardless of the number of AMC levels.

IV. SIMULATION RESULTS

In this section, we present simulation results for the proposed scheme. We consider an OFDM system with $N_c = 64$ subcarriers and the cyclic prefix length is set to 16 samples. In our simulation, we consider the 5 tap exponentially decaying channel whose power delay profile equals [1.000 0.6065 0.3679 0.2231 0.1353]. Also we assume that each channel tap is independently generated with Rayleigh distribution with no spatial correlation.

The AMC set listed in Table I is adopted in our simulations. Each data stream is encoded by the RCPC, where the constraint length of the mother code is 7 and the puncturing patterns in [12] are used to generate various code rates. When evaluating the link performance of AMC schemes, we adopt the “goodput” [14] to measure the system throughput by counting information bits in decoded frames with correct cyclic redundancy check (CRC) in the automatic repeat request (ARQ) mechanism. In our simulation, the “goodput” is measured as

$$\text{goodput} = \frac{\sum N_b}{N_{\text{Tx}} \cdot N_s}$$

where N_b , N_{Tx} and N_s indicate the number of information bits in a frame with correct CRC, the number of transmission and the number of symbols per frame, respectively.

In Fig. 3, we compare the goodput of 2×2 MIMO-OFDM systems with ML detector and the MMSE linear receiver. The results of ideal AMC systems are obtained by exhaustive search where the receiver computes the goodput for all possible AMC set combinations (in our case, $5^2=25$) and then, selects the AMC level which yields the highest goodput for each channel realization. When the number of antennas and the AMC set size increase, the exhaustive search size becomes prohibitive. In practice, it is impossible to employ such a scheme due to extremely high computational complexity. In contrast, with the BL-AMC scheme using our SNR representation, we can easily obtain the expected goodput for each AMC set. Through simulations, we find that $\eta = 0.7$ achieves the maximum link performance for 2×2 systems. It should be noted that the optimum η which maximizes the throughput may change when different AMC sets or channel codes are employed. From this figure, we confirm that the MIMO-MLD shows about a 1.5 dB gain over the system with MMSE receivers. Also we can see that the AMC systems based on the proposed SNR representation yield the almost identical throughput performance compared to the optimum

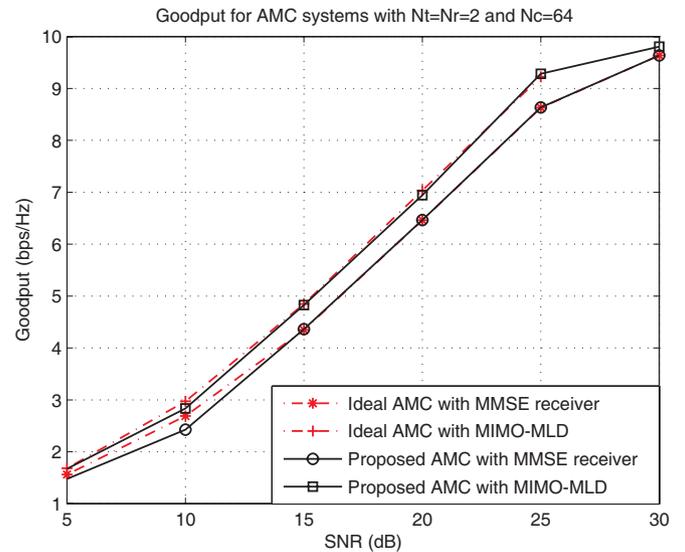


Fig. 3. Goodput comparison of 2×2 AMC systems in frequency selective fading channels.

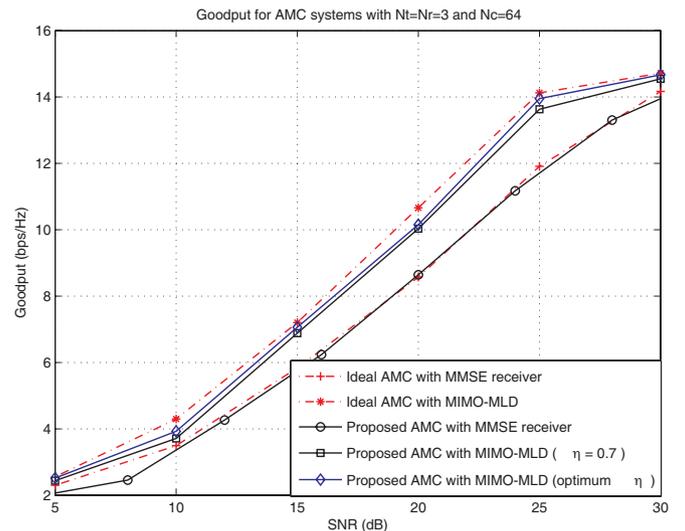


Fig. 4. Goodput comparison of 3×3 AMC systems in frequency selective fading channels.

MIMO system obtained by exhaustive search. Moreover, this result is an evidence that our assumption of $\beta_i = \beta$ in (5) is a valid approximation for the proposed AMC system. One reason of a small difference in the goodput at low SNR can be attributed to a fact that (6) is derived under the high SNR assumption.

Fig. 4 shows the goodput for MIMO-OFDM systems with different AMC schemes with $N_t = N_r = 3$. It is clear that the performance gap between the MIMO-MLD and the MMSE linear receiver becomes larger as the number of the transmit and receive antennas increases, compared with Fig. 3. In this simulation, we also use $\eta = 0.7$ for the AMC system with the MIMO-MLD, which may not be optimum since β depends on the SNR for 3×3 systems unlike 2×2 systems as shown in Fig. 2. For comparison purpose, we plot the goodput with the optimum η for each SNR value. Nevertheless, we can see that the goodput of the proposed AMC system using the fixed η achieves about 95% of that of the ideal AMC case

with reduced complexity. From these simulation results, we conclude that the proposed SNR representation for MIMO-MLD is quite effective in achieving the optimum throughput with reduced complexity.

V. CONCLUSION

In this paper, we have proposed a simple SNR representation method for AMC systems with the MIMO-MLD based on the BL-AMC. By analyzing the upper bound and lower bound of the ML receiver performance based on the channel capacity, we obtain the SNR expression for each stream in the MIMO-MLD. Using the proposed SNR representation, the system throughput is maximized by employing the BL-AMC scheme. The proposed method achieves a significant reduction in the computational complexity in determining the AMC level. Through simulations in OFDM channels, we confirm that the proposed method generates the goodput performance very close to the optimum case with reduced complexity. The proposed method requires the optimization of a single parameter regardless of the AMC size. We expect that the performance gain of the proposed scheme increases with the number of antennas.

VI. ACKNOWLEDGMENT

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APPENDIX PROOF OF LEMMA 1

We denote a 2×2 MIMO channel matrix as

$$\mathbf{H} = \begin{bmatrix} h_1 & h_2 \\ h_3 & h_4 \end{bmatrix}.$$

Then C_{MRC} can be expressed as

$$\begin{aligned} C_{\text{MRC}} &= C_{\text{MRC},1} + C_{\text{MRC},2} \\ &= \log_2 \left(1 + \alpha (|h_1|^2 + |h_2|^2 + |h_3|^2 + |h_4|^2) \right. \\ &\quad \left. + \alpha^2 (|h_1|^2|h_2|^2 + |h_2|^2|h_3|^2 + |h_3|^2|h_4|^2 + |h_4|^2|h_1|^2) \right) \\ &= \log_2 \prod_{i=1}^2 (1 + \alpha \|\mathbf{h}_i\|^2) \end{aligned} \quad (7)$$

where $C_{\text{MRC},i} = \log_2 (1 + \rho_{\text{MRC},i})$.

Also, the open loop capacity can be obtained by

$$\begin{aligned} C_{\text{open}} &= \log_2 \det (\mathbf{I}_2 + \alpha \mathbf{H} \mathbf{H}^\dagger) \\ &= \log_2 \left(1 + \alpha (|h_1|^2 + |h_2|^2 + |h_3|^2 + |h_4|^2) \right. \\ &\quad \left. + \alpha^2 (|h_1|^2|h_4|^2 + |h_2|^2|h_3|^2) \right. \\ &\quad \left. - \alpha^2 (h_1 h_2^* h_3^* h_4 + h_1^* h_2 h_3 h_4^*) \right). \end{aligned} \quad (8)$$

For the MMSE case, the receive filter \mathbf{G} is calculated as

$$\mathbf{G} = \frac{\alpha}{\det(\alpha \mathbf{H}^\dagger \mathbf{H} + \mathbf{I}_2)} \begin{bmatrix} g_1 & g_2 \\ g_3 & g_4 \end{bmatrix} \quad (9)$$

where $g_1 = h_1 + \alpha h_1 |h_4|^2 - \alpha h_2 h_3 h_4^*$, $g_2 = h_2 + \alpha h_2 |h_3|^2 - \alpha h_1 h_3^* h_4$, $g_3 = h_3 + \alpha h_3 |h_2|^2 - \alpha h_1 h_2^* h_4$ and $g_4 = h_4 + \alpha h_4 |h_1|^2 - \alpha h_1^* h_2 h_3$.

Applying the MMSE filter \mathbf{G}^\dagger to the received signal \mathbf{y} , the filter output \mathbf{r} becomes

$$\mathbf{r} = [r_1 \ r_2]^T = \mathbf{G}^\dagger \mathbf{y} = \mathbf{D} \mathbf{s} + \tilde{\mathbf{w}}$$

where $\mathbf{D} = \mathbf{G}^\dagger \mathbf{H}$, $\tilde{\mathbf{w}} = \mathbf{G}^\dagger \mathbf{w}$ and $(\cdot)^T$ stands for the transpose operation. Here, we can express \mathbf{D} as

$$\mathbf{D} = \begin{bmatrix} \mathbf{g}_1^\dagger \mathbf{h}_1 & \mathbf{g}_1^\dagger \mathbf{h}_2 \\ \mathbf{g}_2^\dagger \mathbf{h}_1 & \mathbf{g}_2^\dagger \mathbf{h}_2 \end{bmatrix} = \frac{\alpha}{\det(\alpha \mathbf{H}^\dagger \mathbf{H} + \mathbf{I}_2)} \begin{bmatrix} d_1 & d_2 \\ d_3 & d_4 \end{bmatrix} \quad (10)$$

where $d_1 = |h_1|^2 + |h_3|^2 + \alpha (|h_2|^2|h_3|^2 + |h_1|^2|h_4|^2) - \alpha (h_1 h_2^* h_3^* h_4 + h_1^* h_2 h_3 h_4^*)$, $d_2 = h_1^* h_2 + h_3^* h_4$, $d_3 = h_1 h_2^* + h_3 h_4^*$ and $d_4 = |h_2|^2 + |h_4|^2 + \alpha (|h_2|^2|h_3|^2 + |h_1|^2|h_4|^2) - \alpha (h_1 h_2^* h_3^* h_4 + h_1^* h_2 h_3 h_4^*)$.

Substituting (9) and (10) into (3), we can obtain the SINR of the first and second stream as

$$\begin{aligned} \rho_{\text{MMSE},1} &= \frac{\alpha |d_1|^2}{\alpha |d_2|^2 + (|g_1|^2 + |g_3|^2)} \\ \rho_{\text{MMSE},2} &= \frac{\alpha |d_4|^2}{\alpha |d_3|^2 + (|g_2|^2 + |g_4|^2)}. \end{aligned}$$

After some manipulations with (7) and (8), it can be shown that $2C_{\text{open}} = C_{\text{MRC}} + C_{\text{MMSE}}$ where $C_{\text{MMSE}} = \log_2 (1 + \rho_{\text{MMSE},1}) (1 + \rho_{\text{MMSE},2})$. This result proves that β for 2×2 systems is always 1 for an arbitrary channel matrix \mathbf{H} . ■

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