An Effective Link Error Prediction Technique for MIMO-OFDM Systems with ML Receiver

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***Abstract*—In this paper, we propose an accurate link performance abstraction technique for multiple-input multiple-output (MIMO) orthogonal frequency-division multiplexing systems with maximum likelihood (ML) receiver. The performance of ML detection (MLD) is estimated by using capacity bounds of two simple linear receivers. To this end, we give a simple parametrization to compute the desired per-stream signal-to-noise ratio (SNR) values, which can be applied for both vertically and horizontally coded MIMO systems. Based on the per-stream SNR estimates, the block error rate performance for each encoding block is finally obtained using the received-bit information rate metrics. From extensive simulations, we verify that the proposed method is accurate in the MIMO-MLD link evaluation with very low computational complexity.**

I. Introduction

Next generation wireless cellular systems are expected to support extremely high speed packet data services, providing users with rapid access to high quality multimedia applications. Recently, a capacity gain offered by multiple-input multiple-output (MIMO) techniques have been extensively studied for the last decade [1]. Also, the orthogonal frequency-division multiplexing (OFDM) enables reliable high bit rate transmission over frequency selective fading channels without requiring complicated equalizers [2]. Accordingly, the MIMO-OFDM air-interface allows us to satisfy the high speed service without increasing bandwidth or transmit powers. Combined with bit-inverleaved coded modulation (BICM) [3], the MIMO-OFDM system offers high spectral efficiency and good diversity gains against multipath fading channels [4] [5].

In order to approach the capacity limit of MIMO-OFDM systems, an adaptive modulation and coding (AMC) technique is indispensable in typical cellular systems [6]. When adopting the AMC, accurate link performance estimation is essential to decide adequate modulation levels and/or channel code rates. This step is also called physical layer (PHY) abstraction, and a variety of effective signal-to-noise ratio (SNR) mapping (ESM) methods have been proposed in single-antenna OFDM systems [7]–[9]. However, it is widely recognized that accurate link prediction becomes challenging for MIMO-OFDM systems especially with maximum likelihood (ML) receiver. Although it is adopted in most future standards such as 3GPP long term evolution (LTE) and IEEE 802.16m, MIMO ML detection (MLD) still needs more development in the area of link adaptation.

For this matter, a new approach to the PHY abstraction for the MIMO-MLD was proposed in [10] for the horizontal coding structure where each data stream is encoded by individual encoders. The same idea has been applied to the vertical encoding case [11] where all data streams are simultaneously encoded by a single channel encoder. In this paper, we propose an improved link error prediction technique for MIMO-OFDM systems with ML receiver. By utilizing capacity upper and lower bounds, we develop a simple yet accurate streamwise SNR representation process for both vertical and horizontal encoding MIMO-MLD systems. Based on the derived SNR estimates, the corresponding block error rate (BLER) performance for each encoding block is obtained through the ESM using the received-bit information rate (RBIR) metrics [9]. From simulation results, we confirm that the proposed method is quite accurate in the MIMO-MLD link performance evaluation while preserving very low computational complexity.

The remainder of this paper is organized as follows: In Section II, we describe the system model of adaptive MIMO-OFDM systems. Section III reviews general PHY abstraction methodologies and in Section IV, we explain the proposed link error prediction technique for the MIMO-MLD. Section V verifies the accuracy of the proposed method from extensive simulation results, and we finish the paper with conclusions in Section VI.

Throughout this paper, we use the following notations. Boldface upper-case letters and boldface lower-case letters indicate matrices and column vectors, respectively. Also, $(\cdot)^T$ and $(\cdot)^H$ are denoted as transpose and Hermitian transpose, respectively. The notation $[\mathbf{A}]_{kk}$ indicates the k-th diagonal entry of a matrix $\mathbf{A}$ and an $N \times N$ identity matrix is defined by $\mathbf{I}_N$. The expectation operation is represented as $\mathbb{E}[\cdot]$.

II. System Model

We consider an adaptive MIMO-OFDM system with N_t transmit and N_r receive antennas. In each time slot, a receiver chooses a proper modulation and coding set (MCS) according to the current channel condition and feeds back its index to the AMC controller at the transmitter. The MCS levels consist of the quadrature amplitude modulation (QAM) level and the channel code rate, which are predetermined depending on system requirements. Then, at the transmitter, based on the BICM structure, the information bits are encoded either vertically or horizontally, bit-wise interleaved and mapped to symbol constellations after serial-to-parallel conversion. After the mapping, the data symbols are modulated by the N_c-size inverse fast Fourier transform (IFFT).

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Assuming proper cyclic prefix (CP), the N_r dimensional received signal vector at the k-th subcarrier after the FFT demodulation is given by

$$\mathbf{y}_k = \mathbf{H}_k \mathbf{x}_k + \mathbf{n}_k \tag{1}$$

where $\mathbf{x}_k \in \mathbb{C}^{N_t \times 1}$ is the transmitted symbol vector, $\mathbf{n}_k \in \mathbb{C}^{N_r \times 1}$ denotes the additive white Gaussian noise (AWGN) vector with zero mean and the covariance matrix $\sigma_n^2 \mathbf{I}_{N_r}$, and $\mathbf{H}_k = [\mathbf{h}_1 \cdots \mathbf{h}_{N_t}] \in \mathbb{C}^{N_r \times N_t}$ equals the MIMO Rayleigh fading channel matrix whose entries have an independent and identically distributed (i.i.d.) complex Gaussian distribution with $\mathcal{CN}(0,1)$. In the absence of the channel state information at the transmitter, $\mathbf{x}_k$ satisfies $\mathbb{E}[\mathbf{x}_k \mathbf{x}_k^H] = \frac{P}{N_t}\mathbf{I}_{N_t}$ with uniform power allocation across the transmit antennas, where P is the total transmit power. We assume that all subchannel matrices $\mathbf{H}_1, \cdots, \mathbf{H}_{N_c}$ are perfectly known at the receiver.

III. Review of Effective SNR Mapping

In this section, we review the principle of the physical layer link performance abstraction by focusing on existing ESM methods.

A. Single-input single-output case

For single-input single-output (SISO) OFDM systems, most ESM methods calculate the effective SNR γ_{eff} according to a mapping function $\mathcal{F}$ as

$$\gamma_{\text{eff}} = \mathcal{F}^{-1}\left(\frac{1}{N_c}\sum_{k=1}^{N_c}\mathcal{F}(\gamma_k)\right) \tag{2}$$

where γ_k is the received SNR of the k-th subcarrier determined as $\gamma_k = |h_k|^2 P/\sigma_n^2$. Here, h_k represents the gain of the k-th subchannel. Once γ_{eff} is computed from (2), we can estimate the BLER of the link for each MCS by looking up the AWGN reference curves. We may choose γ_{eff} as the simple average $\gamma_{\text{eff}} = \frac{1}{N_c}\sum_{k=1}^{N_c}\gamma_k$ with $\mathcal{F}(x) = x$. However, this linear estimation cannot capture different diversity gains coming from large deviations of multipath channel delay profiles.

In order to accurately model the link performance (2), exponential ESM (EESM) has been proposed based on the Chernoff bound of the coded symbol error rate performance [7], which computes γ_{eff} in (2) with the exponential function $\mathcal{F}(x) = \exp(-x/\beta)$, where β is a predetermined system parameter. Another approach for the ESM is to define the mapping $\mathcal{F}$ based on the mutual information function such as the RBIR [9] and the mean mutual information per bit (MMIB) [8] [12] metrics. For SISO systems, these mutual information based ESM methods are preferable because they do not require any system parameter and work well in hybrid automatic repeat request (HARQ) schemes [9]. All these methods are shown to have pretty good accuracy for single antenna systems.

B. MIMO case with ML receiver

The BLER estimation for MIMO systems with linear receiver or successive interference cancellation (SIC) receiver is straightforward, since the link quality of (2) for each stream can be easily defined by per-stream SNR values. However, it becomes challenging with the MIMO-MLD because the link quality of each individual data stream is difficult to estimate due to non-linear joint processing. When the MMIB method is applied for MIMO-MLD, quite a complicated process is required for optimizing its system parameters [12] and it does not work well with the horizontal encoding. Also, the RBIR procedure suggested in [13] is found to be invalid for the MIMO-MLD either [14].

IV. Improved PSSR Techniques

In this section, we illustrate the proposed PSSR technique for the MIMO-MLD link prediction. We start with defining upper and lower bounds of the MIMO-MLD capacity. Then, we explain how to set up and optimize the parameters to estimate the per-stream SNR values. For now, we drop the subcarrier index k for notational simplicity.

The purpose of the PSSR method is to obtain accurate per-stream SNR values with the MLD, which are denoted by $\gamma_{\text{ML},1}, \cdots, \gamma_{\text{ML},N_t}$, for a given channel matrix $\mathbf{H}$. First, we define the capacity C_{ML} as

$$C_{\text{ML}} = \sum_{n=1}^{N_t} C_{\text{ML},n} \triangleq \sum_{n=1}^{N_t}\log_2(1+\gamma_{\text{ML},n}) \tag{3}$$

where $C_{\text{ML},n}$ indicates the n-th stream capacity. Here, C_{ML} is the achievable rate when the MLD is performed at the receiver, which should be distinguished from the MIMO open-loop capacity $C_{\text{open}} = \log_2\left|\mathbf{I}_{N_r} + \frac{P}{N_t\sigma_n^2}\mathbf{H}\mathbf{H}^H\right|$ [1].

Because the true values of $\{\gamma_{\text{ML},n}\}_{n=1}^{N_t}$ are unknown, two simple bounds for (3) can be adopted to estimate $\{\gamma_{\text{ML},n}\}_{n=1}^{N_t}$. First, an upper bound of C_{ML} is derived by assuming that interference among data symbols is perfectly removed at the receiver, called as perfect interference cancellation (PIC). As a lower bound, the linear minimum mean-square error (MMSE) receiver $\mathbf{G} = (\mathbf{H}^H\mathbf{H} + \frac{N_t\sigma_n^2}{P}\mathbf{I}_{N_t})^{-1}\mathbf{H}^H$ is considered. The received SNR of the PIC and the signal-to-interference-plus-noise ratio (SINR) of the MMSE receiver for the n-th stream are given, respectively, as

$$\gamma_{\text{PIC},n} = \frac{\|\mathbf{h}_n\|^2 P}{N_t\sigma_n^2} \tag{4}$$

$$\gamma_{\text{MMSE},n} = \frac{1}{\left[\left(\mathbf{I}_{N_t} + \frac{P}{N_t\sigma_n^2}\mathbf{H}^H\mathbf{H}\right)^{-1}\right]_{nn}} - 1, \tag{5}$$

and the corresponding instantaneous capacities are expressed by $C_{\text{PIC}} = \sum_{n=1}^{N_t} C_{\text{PIC},n} = \sum_{n=1}^{N_t}\log_2(1+\gamma_{\text{PIC},n})$ and $C_{\text{MMSE}} = \sum_{n=1}^{N_t} C_{\text{MMSE},n} = \sum_{n=1}^{N_t}\log_2(1+\gamma_{\text{MMSE},n})$, respectively.

Depending on these bounds, it follows that $\gamma_{\text{ML},n}$ is bounded as $\gamma_{\text{MMSE},n} \le \gamma_{\text{ML},n} \le \gamma_{\text{PIC},n}$, which leads to the same relation among the capacities $C_{\text{ML}}, C_{\text{PIC}}$ and C_{MMSE} as

$$C_{\text{MMSE}} \le C_{\text{ML}} \le C_{\text{open}} \le C_{\text{PIC}} \tag{6}$$

where the equalities hold when all columns of $\mathbf{H}$ are orthogonal. Note that for a given $\mathbf{H}$, $C_{\text{MMSE}}, C_{\text{open}}$ and C_{PIC} are known values. In what follows, we explain how to determine the per-stream SNRs $\gamma_{\text{ML},1}, \cdots, \gamma_{\text{ML},N_t}$ by using the relation (6) for both vertical and horizontal encoding.

TABLE I
THE OPTIMAL VALUES OF b FOR DIFFERENT MCS LEVELS

Modulation \ Code rate	1/2	2/3	3/4	5/6	7/8
4-QAM	0.6	0.8	0.9	1.0	1.0
16-QAM	0.4	0.6	0.7	0.8	0.8
64-QAM	0.2	0.4	0.5	0.6	0.6

A. Vertical encoding

For vertically coded systems, we introduce two parameters a and b to denote the ratios of capacity gaps from (6), respectively, as

$$a = \frac{C_{\text{open}} - C_{\text{MMSE}}}{C_{\text{PIC}} - C_{\text{MMSE}}}, \quad 0 \le a \le 1 \tag{7}$$

$$b = \frac{C_{\text{ML}} - C_{\text{MMSE}}}{C_{\text{open}} - C_{\text{MMSE}}}, \quad 0 \le b \le 1 \tag{8}$$

where the ranges of a and b are from $C_{\text{open}} \le C_{\text{PIC}}$ and $C_{\text{ML}} \le C_{\text{open}}$, respectively. In order to obtain $\gamma_{\text{ML},n}$, we also define β_n as the ratio of capacity gaps for the n-th data stream as

$$\beta_n = \frac{C_{\text{ML},n} - C_{\text{MMSE},n}}{C_{\text{PIC},n} - C_{\text{MMSE},n}} \quad \text{for} \quad n = 1, \cdots, N_t. \tag{9}$$

Naturally, β_n are different for each individual stream. However, these different values of $\beta_1, \cdots, \beta_{N_t}$ cannot be quantitatively evaluated since $C_{\text{ML},n}$ is unknown.

Accordingly, to simplify the derivation, we assume that all substreams have the same ratio of (9) for any $\mathbf{H}$ as

$$\beta_n = \frac{C_{\text{ML}} - C_{\text{MMSE}}}{C_{\text{PIC}} - C_{\text{MMSE}}} = a \cdot b. \tag{10}$$

For vertical encoding systems where a single MCS is employed across streams, this approximation is true in an average sense over all channel realizations[1]. Moreover, simulations show that $\beta_1, \cdots, \beta_{N_t}$ are not very distinct, even if $\mathbf{H}$ is ill-conditioned.

By rearranging (9) and (10), we have $C_{\text{ML},n} = a \cdot b\, C_{\text{PIC},n} + (1 - a \cdot b) C_{\text{MMSE},n}$. Consequently, the SNR of the n-th stream for the MIMO-MLD is finally represented as

$$\gamma_{\text{ML},n} = (1 + \gamma_{\text{PIC},n})^{a \cdot b} (1 + \gamma_{\text{MMSE},n})^{1 - a \cdot b} - 1. \tag{11}$$

Note that in (11), $\gamma_{\text{PIC},n}$, $\gamma_{\text{MMSE},n}$ and a are directly obtained from (4), (5) and (7) as functions of $\mathbf{H}$, respectively. Therefore, only one single parameter b needs to be optimized.

In Table I, the optimal values of b with 15 different MCS levels are listed, which are obtained by the link level simulation (LLS). For each MCS, the optimal b can be found by fitting the BLER curves to the AWGN reference curve. In addition, since b has a small range, $0 \le b \le 1$, the search is quite simple. The LLS environments for Table I are illustrated in Section V.

At this point, the advantages of our parametrization should be emphasized. First, it turns out that the parameter b is independent of N_t and N_r. In other words, Table I is valid for any MIMO antenna configurations. Also, we find that b is insensitive to the channel delay profile, which is a very desirable condition for a good PHY abstraction method. These two features will be verified in the simulation section. In addition, Table I shows that b monotonically increases with the code rate and decreases with the modulation level. This property ensures that by applying simple linear interpolation, b for any other new MCS level can be readily found without additional simulation efforts.

[1]This is because all columns of $\mathbf{H}$ are statistically equivalent (i.i.d. random vectors). However, this assumption may be broken if the modulation level in each stream are not identical, which can occur in horizontal encoding systems. The description for the horizontal encoding is provided in the next subsection.

B. Horizontal encoding

Unlike the vertical encoding case, different modulation levels can be allocated to each data stream for the horizontally coded system. If one or more different modulation levels are applied in other substreams, each value of our predictions $\gamma_{\text{ML},1}, \cdots, \gamma_{\text{ML},N_t}$ made in (11) may not be valid. This is because the MLD performance of each layer is determined by the minimum Euclidean distance of the joint symbol constellation. For example, let us consider a case where 4-QAM and 16-QAM are applied for the first and the second data stream for $N_t = 2$, respectively. Compared to the case where 4-QAM is employed for both streams, the detection accuracy of the first stream gets degraded due to smaller minimum distance of 16-QAM constellation at the second stream. On the contrary, the second stream exhibits higher accuracy since 4-QAM at the first stream improves the overall minimum distance compared to the case where 16-QAM is applied for both streams. To reflect these changes from the identical MCS case, proper compensations should be made to our SNR expression (11).

For this purpose, we adopt another system parameter Δb_n per each stream to modify b. Certainly, a is not related to this change of the MLD performance. By replacing b with $b + \Delta b_n$, we rewrite (10) as

$$\overline{\beta}_n = \frac{C_{\text{ML},n} - C_{\text{MMSE},n}}{C_{\text{PIC},n} - C_{\text{MMSE},n}} = a\,(b + \Delta b_n).$$

Then, in the same way as the vertical encoding case in (11), we can estimate the per-stream SNR of the MIMO-MLD $\gamma_{\text{ML},n}$ as

$$\gamma_{\text{ML},n} \approx (1 + \gamma_{\text{PIC},n})^{a(b+\Delta b_n)} (1 + \gamma_{\text{MMSE},n})^{1 - a(b+\Delta b_n)} - 1. \tag{12}$$

It is found that the values of Δb_n are independent of the code rate since the detection performance is not affected by channel coding. Also, if all modulation levels are identical, we have $\Delta b_1 = \cdots = \Delta b_{N_t} = 0$. As a result, we only have total of $3, 7$ and 12 combinations of different modulation levels for $N_t = 2, 3$ and 4, respectively, assuming 4, 16 and 64-QAM. Table II shows the optimal values of Δb_n found by simulations for $N_t = 2$ and 3 cases. Note that Δb_n is also obtained offline and does not include numerical approximations depending on the SNR and the channel condition number as in [13].

C. Received bit information rate link error prediction

So far, we have investigated how to estimate the SNR values $\{\gamma_{\text{ML},n}\}_{n=1}^{N_t}$ of the MIMO-MLD for each subcarrier from the proposed improved PSSR technique. Now, the next step is to predict the link performance of each layer based

TABLE II
TUNING PARAMETERS Δb_n FOR NON-IDENTICAL MODULATION LEVELS WITH $N_t = 2$ AND 3

	Modulation levels	$\{\Delta b_n\}_{n=1}^{N_t}$
$N_t = 2$	$(4,16),(4,64),(16,64)$	$(-0.3,0.7),(-0.4,1.2),(-0.2,0.6)$
$N_t = 3$	$(4,4,16),(4,4,64),(4,16,16),(4,16,64),$ $(4,64,64),(16,16,64),(16,64,64)$	$(-0.3,-0.3,0.6),(-0.3,-0.3,1.5),(-0.6,0.3,0.3),(-0.8,0.3,1.2),$ $(-0.4,0.8,0.8),(-0.2,-0.2,0.6),(-0.2,0.4,0.4)$

on $\{\gamma_{\mathrm{ML},n}\}_{n=1}^{N_t}$. As previously noted, there are many well-designed link quality models which can be applied for single-antenna systems. Among them, we choose the RBIR approach [9] since it is simple and has several advantages over the EESM.

For the horizontal encoding, the RBIR metric for the n-th stream is defined by

$$\mathrm{RBIR}_n = \sum_{k=1}^{N_c} \frac{\mathrm{SI}_{n,k}(\gamma_{\mathrm{ML},n,k})}{\log_2(m_{n,k})}, \quad n = 1, \cdots, N_t \tag{13}$$

where $m_{n,k}$ denotes the modulation level of the n-th stream at the k-th subcarrier and $\mathrm{SI}_{n,k}(\cdot)$ is the symbol mutual information (SI) derived as [9]

$$\mathrm{SI}_{n,k}(\gamma) = \frac{1}{m_{n,k}} \sum_{j=1}^{m_{n,k}} \int_l f_{LLR_j(\gamma)}(l) \log_2 \frac{m_{n,k}}{1+e^{-l}} dl. \tag{14}$$

Here, $f_{LLR_j(\gamma)}$ equals the probability density function of symbol level log-likelihood ratio (LLR) of the j-th constellation point with SNR equal to γ [9]. To avoid repeated calculations, (14) is computed for a wide range of SNRs and saved in a lookup table [13, Table 25]. Using (13) and the SNR-to-SI mapping table, N_t RBIR values are obtained from $\{\gamma_{\mathrm{ML},n,k}\}$ and then inversely mapped after normalization to get the effective SNRs $\gamma_{\mathrm{eff},1}, \cdots, \gamma_{\mathrm{eff},N_t}$. Finally the BLER of each data stream can be estimated by looking up the AWGN reference curves.

In contrast, the link performance of the vertical encoding is represented by a single BLER estimate. This can be resolved by averaging the RBIR values in the mutual information domain as

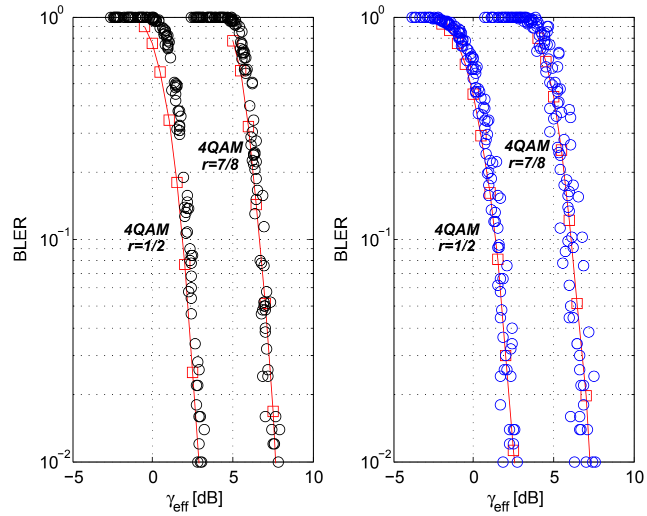
$$\mathrm{RBIR}_{\mathrm{ave}} = \frac{1}{N_t} \sum_{n=1}^{N_t} \sum_{k=1}^{N_c} \frac{\mathrm{SI}_{n,k}(\gamma_{\mathrm{ML},n,k})}{\log_2(m_{n,k})}. \tag{15}$$

The remaining steps to get a single effective SNR γ_{eff} and the corresponding BLER expectation are the same as the horizontal case.

V. SIMULATION RESULTS

In this section, we provide the LLS results to demonstrate the BLER prediction accuracy of the proposed scheme based on the improved PSSR method. We consider an OFDM system with $N_c = 64$ and the CP length of 16 samples, and a coding block is set to one OFDM symbol. We assume block Rayleigh fading channels with a 5-tap exponentially decaying delay profile, if not specified otherwise. For the channel code, the rate-compatible punctured convolutional (RCPC) codes with polynomials $(133, 171)$ in octal is employed [15].

In Figures 1 and 2, we compare the BLER estimates from the proposed scheme (circle) with the AWGN reference curves (solid line) for both vertical and horizontal encoding systems. For each MCS, 20 independent channel realizations are simulated and each channel is averaged over $3,000$ noise samples. Also, we utilize Table I as the optimal values of the parameter b. Figure 1 shows the fitting results for 4×4 MIMO systems over various modulation levels and code rates. We emphasize that for both the vertical and the horizontal encoding, almost all the estimated γ_{eff} values in the graphs are within 1 dB from the reference curves. Note that this level of accuracy is quite acceptable in typical AMC schemes with moderate numbers of the MCSs.

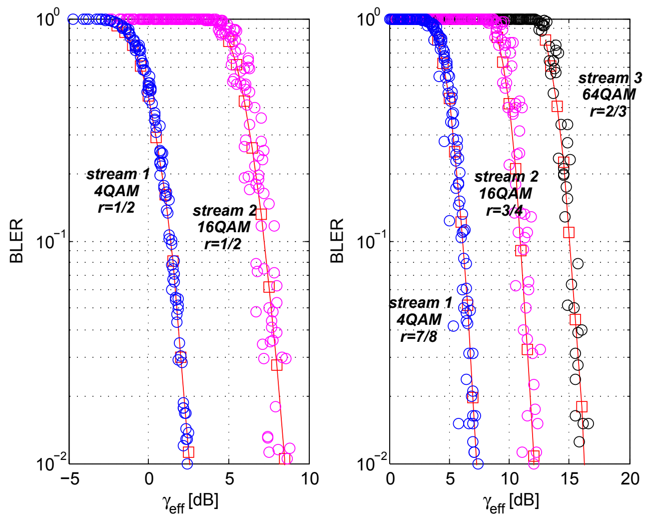


(a) vertical encoding (b) horizontal encoding

Fig. 1. Link prediction accuracy for 4×4 vertical and horizontal encoding systems with identical MCSs

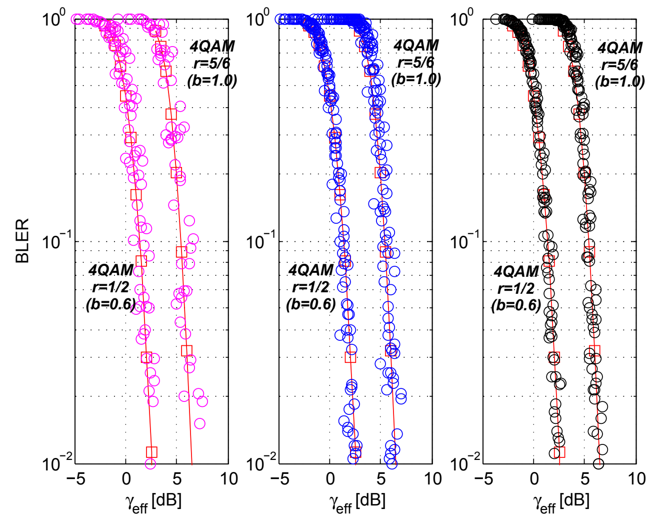
In Figure 2, the horizontal encoding with different MCS levels over streams are plotted in 2×2 and 3×3 systems. In these simulations, the offset parameters $(\Delta b_1, \Delta b_2) = (-0.3, 0.7)$ and $(\Delta b_1, \Delta b_2, \Delta b_3) = (-0.8, 0.3, 1.2)$ are applied to all streams for $N_t = 2$ and 3, respectively, according to Table II. Similar to Figure 1, we find that high prediction accuracy is achieved at all data streams. From the figures, we can conclude that the proposed link error prediction method is very suitable for the MIMO ML systems.

Next, we verify the property of the system parameter b discussed in Section IV-A. In Figures 3 and 4, we show the performance of the PHY abstraction for different numbers of the receive antennas and the channel taps, respectively. As representative examples, we plot the results for $N_t = 2$ horizontal encoding systems with 4-QAM and 16-QAM modulations. Here, we observe that for each MCS, a single fixed value of b in Table I is well fitted for all cases of $N_r = 2, 3$ and 4



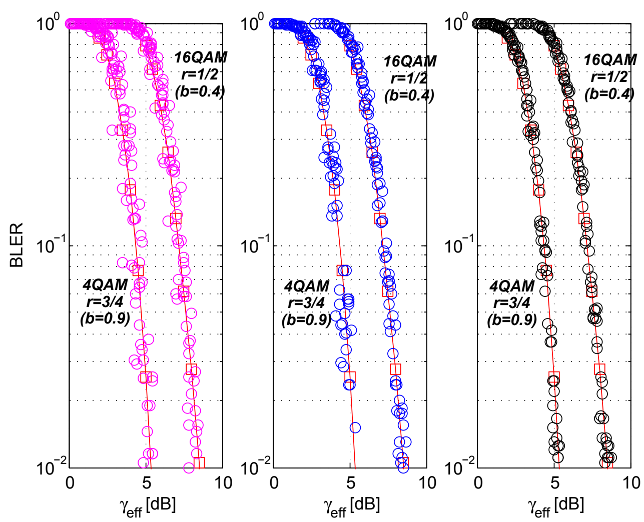
(a) 2×2 systems (b) 3×3 systems

Fig. 2. Link prediction accuracy for 2×2 and 3×3 horizontal encoding systems with non-identical MCSs



(a) 1-tap (b) 5-tap (c) 10-tap

Fig. 4. Link prediction accuracy in terms of the number of channel taps for 2×2 systems with horizontal encoding



(a) $N_r = 2$ (b) $N_r = 3$ (c) $N_r = 4$

Fig. 3. Link prediction accuracy in terms of the number of receive antennas for horizontal encoding with $N_t = 2$

and different channel delay profiles with the number of taps $1, 5$ and 10. Although not shown here, we have confirmed that Table II also does not depend on the number of channel taps. This robustness is certainly a great merit for implementation of practical AMC schemes.

VI. Conclusions

In this paper, we have proposed an accurate link abstraction technique for MIMO-OFDM systems with ML receiver. By combining our improved PSSR method with the existing SISO RBIR mapping procedure, an exact link error prediction has been made with very low computational complexity. We have confirmed from simulations that the proposed method is quite effective in both vertical and horizontal MIMO-MLD systems with various number of antennas and different amount of frequency selectivity.

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