Near Optimal Combining Scheme for MIMO-OFDM HARQ with Bit Rearrangement

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Abstract

In this paper, we study the combining scheme for type I hybrid Automatic Repeat-reQuest (HARQ) in MIMO-OFDM system. A new pre-combining scheme based on hybrid QRMC detector is proposed. The proposed scheme is applied to HARQ with various bit rearrangement strategies. The significant gain has been obtained for both slow fading and fast fading channel comparing with the system using near optimal QRD-M detector as a bitwise post-combining scheme at the cost of increased storage requirement. The performance of various bit rearrangement strategies using the proposed combining method is evaluated and compared. It is shown that proper bit-rearrangement can greatly improve the system performance even in slow fading channel. The new combining scheme has flexible structure and can be applied to any bit rearrangement strategies.

I. Introduction

The rapid growing demand for higher data rate, lower delay latency for quality of service (QoS) guarantee is driving recent development of new communication technologies for broadband wireless communication. The orthogonal frequency division multiplexing (OFDM) digital modulation scheme has been identified as one of the promising technologies for 4th generation wireless broadband systems. Meanwhile, the multiple-input multiple-output (MIMO) technology is one of most significant advancement in the past decade and is known to increase the spectral efficiency of a communication system. The combination of the two, MIMO-OFDM, has been selected as the key physical layer technology for a couple of modern standards system. High quality transmission in a mobile radio channel is another important issue. In severe channel condition, the error rate performance of MIMO-OFDM system is degraded rapidly.

To ensure reliable transmission of MIMO-OFDM system, hybrid automatic repeat request (HARQ) schemes are required. In HARQ schemes, error packets are reused and smartly combined to improve the efficiency of classical ARQ scheme. HARQ is easily implemented and provides a good tradeoff between throughput and reliability. Therefore, it is widely used in modern wireless communication systems such as IEEE 802.16e [1] and 3GPP long term evolution (LTE) [2] standards.

There are two main classes of HARQ: HARQ with chase combining (HARQ-CC) [3] and HARQ with incremental redundancy (HARQ-IR) [4]. Here we focus on HARQ-CC. In original HARQ-CC, the same packet is transmitted during each retransmission until the packet is successfully decoded or the maximal number of retransmission is reached. To further improve the performance of HARQ, several enhanced schemes are proposed [5]–[7]. Essentially, those enhanced schemes reconfigure the bits or symbols in each retransmission to provide additional diversity.

The combining scheme has a great impact on the achievable diversity order and consequently the overall receiver performance. The combining scheme design is closely related with the detector design of MIMO system. In a conventional MIMO system, the optimal *maximal a posterior* (MAP) detector has exponential complexity. Therefore low-complexity MIMO detectors have been developed in recent years. The discussion of various detectors can been found in [8]. Extension of conventional MIMO detectors to the system with HARQ faces some new problems which will be discussed in the paper in detail. In this paper, we study combining scheme for bit rearrangement based HARQ-CC in a MIMO-OFDM system where the bit arrangements are adapted in every retransmission. In particular, a novel near optimal pre-combining scheme is proposed. Similar work is presented in [9], where a pre-combing scheme using a parallel sphere decoding with a special candidate enumeration unit has been proposed to the HARQ system with constellation re-mapping at increased complexity. However, the combining scheme in [9] can only be applied to the system with bit rearrangement in the same symbol. Our proposed combining scheme can not only achieve near optimal performance but also can be applied to any bit rearrangement HARQ system without additional complexity.

The remainder of the paper is organized as follows. In Section II, we introduce the system

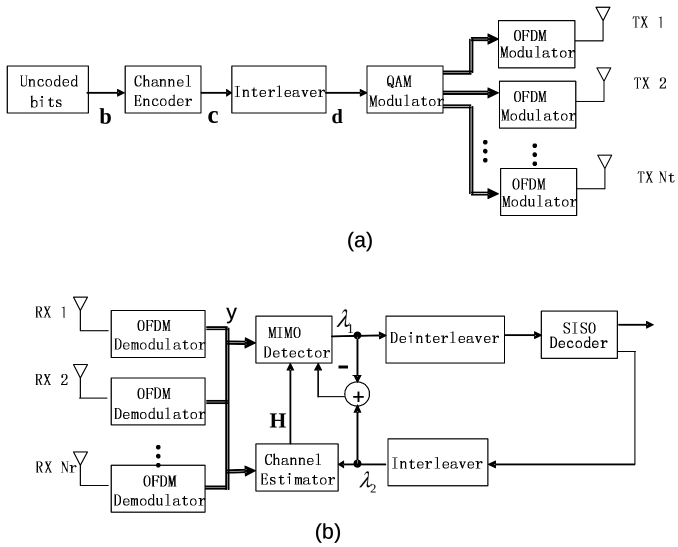


Fig. 1. MIMO-OFDM system model. (a) Transmitter model and (b) Receiver models.

setup. Several bit rearrangement strategies for MIMO-OFDM system is presented in Section III. Section IV includes a detailed description of the proposed combining scheme based hybrid QRMC detector. Simulation results are presented in Section VI. Finally, the conclusions are given in Section VII.

II. System description

In this section, we briefly describe a MIMO-OFDM spatial multiplexing downlink system model. A MIMO-OFDM system with N_t transmitter antennas and N_r receive antennas is assumed. The transmitter structure considered in this work is illustrated in Fig. 1(a). The uncoded binary sequence $\{u_i\}$ are first encoded by channel encoder with a code rate R and are interleaved. After interleaving, the coded bits $\{u_i\}$ are mapped to QAM symbols using mapping function ψ. The modulated symbols $\{d_j\}$ stream is then partitioned to N_t parallel spatial streams. Each spatial stream is broken into K parallel substreams and allocated to K subcarriers over T consective OFDM symbols. The pilot data are also inserted to the pilot carriers at the same time. The inverse fast Fourier transform (IFFT) is performed by the OFDM modulator and cyclically prefixed (CP) guard interval of length N_g are appended. The CP must be set longer than the delay spread of the channel to avoid inter-symbol interference.

When no ICI exists, After FFT, the received samples at the k-th subcarrier of the n-th symbol can be represented by

$$Y_{rx}^{(r)}[n,k] = \sum_{tx=1}^{N_t} d_{tx}^{(r)}[n,k] H_{rx,tx}^{(r)}[n,k] + W_{rx}^{(r)}[n,k] \tag{1}$$

.

where $d_{tx}^{(r)}[n,k]$ is the data transmitted on the k-th subcarrier of the n-th OFDM symbol at the r-th retransmissionm from the antenna tx, $H_{rx,tx}^{(r)}[n,k]$ is corresponding channel frequency response between antenna tx and rx, $W_{rx}^{(r)}[n,k]$ is the corresponding additive white Gaussian noise at antenna Rx. Note r is the set to zero for the initial transmission.

In the system without HARQ, conventional MIMO detection is performed independently on the subcarriers. Hereafter, we will drop the index of subcarrier in most cases. Only where the index of subcarrier needs special attention will it be mentioned explicitly. The soft output MIMO detector generates likelihood ratio of each bit and sent them through deinterleaver to soft input soft output (SISO) channel decoding for decoding. If turbo receiver is adopted, the soft out of channel decoding is feedback through the interleaver for next iteration of detection and decoding. In case the decoding fails, the error packet is discarded.

In HARQ system, the error packets are reused for later processing, thus better utilized the useful information in error packet. Depending on how the data is retansmitted, HARQ is classified into two main types: HARQ-CC and HARQ-IR. For HARQ-CC, the same packet of data are retransmitted during each retransmission. The receiver will perform a combining with buffered data from previous transmission. After decoding, a cyclic redundancy check (CRC) decoder is used to detect the decoding errors. If no error is found, an ACK signal is transmitted through ARQ feedback channel to request the new packet transmission. On the other hand, if any error is detected, an NACK signal is sent to notify the transmitter for retransmission. The above process is repeated until the packet is correctly decoded or the maximum allowed retransmission number reaches. HARQ-IR works in a similar way to HARQ-CC except that HARQ retransmit different symbol sequences which come from the different code scheme of the same data. HARQ-IR can benefit from a coding gain and thus in general obtains better performance than HARQ-CC with more implementation complexity. In this paper, we will focus on the HARQ-CC scheme.

III. Bit rearrangement strategies for retransmissions

Conventional HARQ-CC transmit the same packet between retransmissions. Recent work shows additional diversity can be obtained by rearranging the bits/symbols for retransmissions with minimal increase in system complexity. In this section, we study the bit rearrangement strategies for MIMO-OFDM system. Before we look at the specific bit rearrangement strategies, we first present a general bit rearrangement model.

In principle, bit rearrangement based HARQ-CC protocols adapt the bit arrangement during retransmission to extract additional retransmission diversity. It can be modeled it as following:

$$\mathbf{b}^{(r)} = \phi^{(r)}(\mathbf{b}^{(0)}) \tag{2}$$

where $\mathbf{b}^{(r)}$ denotes the a set of *correlated bits* transmitted on the r-th retransmission, with $r = 0$ denotes the initial transmission and ϕ denote a *bit rearrangement function* (BRF) defined over the bit sequence. The BRF could be a permutation, inverse of some bits or other mapping schemes. The correlated bits are a minimal set of bits correlated at the receiver due to in the same MIMO transmission and/or in the same BRF during retransmission. We assume the same constellation mapping function is used in every retransmissions after BRF is applied.

To achieve the best performance, the BRF should be optimized. However, finding optimal BRF is a non-trivial task when the size of a correlated bits is large. A more practical way is to find sub-optimal scheme, for example, limiting the correlated bits to be the bits in a symbol or the bits in one MIMO transmission based on some diversity criteria. We will follow this approach in the paper.

The first work on bit rearrangement is presented in [5] where BRF is performed over each modulation symbol. Observing the variations in bit reliabilities caused by the multilevel signal constellation, Wengerter *et al.* propose to use different bit swapping and inversion (BSI) function in every retransmission to average the bit reliabilities. The constellation re-mapping diversity is obtained through retransmission. Table I shows the BSI function of a 16QAM symbol for different retransmissions where i_k, q_k are the k-th in-phase and quadrature bit and $\overline{i_k}$ denotes the inversion of i_k.

This work was extended to MIMO system by introducing spacial diversity into retransmission. One scheme is antenna permutation [10], where the mapping between transmitted signals

TABLE I

BIT REARRANGEMENT FOR A 16QAM SYMBOL

Retrans	**Bit sequence**
0	$i_1 q_1 i_2 q_2$
1	$i_2 q_2 \overline{i_1 q_1}$
2	$i_2 q_2 i_1 q_1$
3	$i_1 q_1 \overline{i_2 q_2}$

and transmit antennas are changed upon retransmission. The antenna permutation can be also considered as a symbol-wise shifting scheme between antennas (SSA). Another scheme is bit shifting between antennas (BSA) [7]. In this scheme, bit-wise exchanges between antennas is adopted to maximize spatial diversity. Table II shows the SSA and BBA for the system with 2 transmitters and using 16QAM during retransmissions where d_k denotes the k-th transmitting symbol and $i_{k,j}, q_{k,j}$ denote the j-th in-phase and quadrature bit of the k-th symbol.

TABLE II

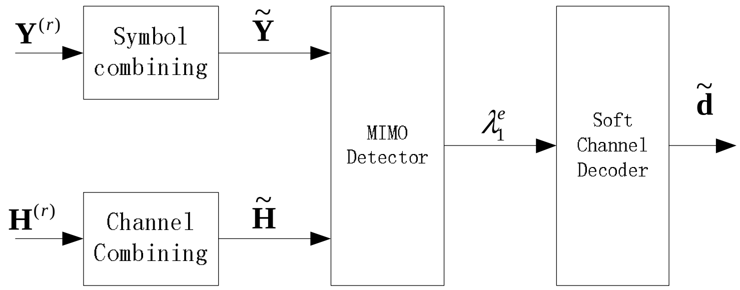
SSA AND BSA FOR 16QAM SYMBOLS WITH 2 TRANSMITTERS

	SSA		**BSA**	
Retrans	Ant 1	Ant 2	Ant 1	Ant 2
0	d_1	d_2	$i_{1,1} i_{1,2} q_{1,1} q_{1,2}$	$i_{2,1} i_{2,2} q_{2,1} q_{2,2}$
1	d_2	d_1	$i_{1,1} i_{2,2} q_{1,1} q_{2,2}$	$i_{2,1} i_{1,2} q_{2,1} q_{1,2}$

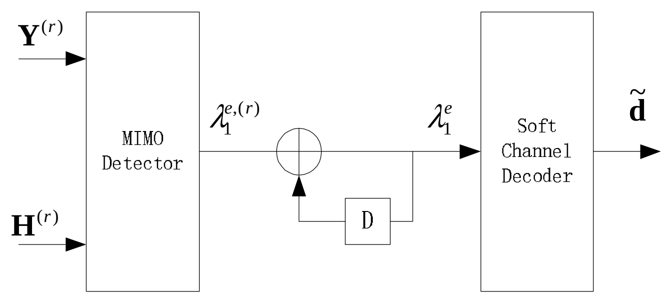
For OFDM system, exploiting multiple subcarriers gives an additional degree of flexibility for retransmission. Different subcarriers can be to allocated in every retransmission to utilize frequency diversity. Analog to MIMO case, there are bit-wise and symbol-wise shift between subcarriers (BSC and SSC) by changing the shift from spacial dimension to frequency dimension.

IV. HARQ COMBINING SCHEMES

The performance of HARQ system is not only decided by the retransmission strategies but also the combining schemes used in the receiver. In this section, we describe two combining schemes assuming the use of MIMO HARQ-CC systems. To simplify the presentation, in this



(a) Pre-combining MTMO-HARQ receiver



(a) Post-combining MTMO-HARQ receiver

Fig. 2. Receiver architectures for MIMO systems with HARQ

section, we only show how to calculates LLR based on ML decoding. In the following section, we will show how to reduce the complexity using the proposed combining scheme.

In a MIMO HARQ-CC system, multiple transmissions of a single transmit vectors leads to multiple received signal vectors at the receiver. The HARQ cumulatively combines the information from multiple recived signal vectors either *before* or *after* the MIMO detection. These two alternatives, depicted in Fig. 2 (a) and (b), are termed *pre-combining* and *post-combining* schemes.

A. Pre-combining scheme

For pre-combining scheme, the received signal vectors along with channel state information (CSI) are combined by concatenating them and form an equivalent big MIMO detector. Consider a transmission of bit sequence $\mathbf{b}$ with length p, a *cumulative* optimal *maximal a posterior* (MAP) detector is given by

$$\begin{aligned}\lambda_{1,k}^{e} &= \ln \frac{P(b_k=+1|\tilde{\mathbf{Y}},\tilde{\mathbf{H}})}{P(b_k=-1|\tilde{\mathbf{Y}},\tilde{\mathbf{H}})} = \ln \frac{\sum\limits_{\mathbf{b}_{-k}} P(b_k=+1,\mathbf{b}_{-k}|\tilde{\mathbf{Y}},\tilde{\mathbf{H}})}{\sum\limits_{\mathbf{b}_{-k}} P(b_k=-1,\mathbf{b}_{-k}|\tilde{\mathbf{Y}},\tilde{\mathbf{H}})} \\ &\approx \ln \frac{\max\limits_{\mathbf{b}_{-k}} P(b_k=0,\mathbf{b}_{-k}|\tilde{\mathbf{Y}},\tilde{\mathbf{H}})}{\max\limits_{\mathbf{b}_{-k}} P(b_k=1,\mathbf{b}_{-k}|\tilde{\mathbf{Y}},\tilde{\mathbf{H}})} \\ &= \max_{\mathbf{b}_{-k,0}} \left\{ \frac{-|\tilde{\mathbf{Y}}-\tilde{\mathbf{H}}\tilde{\mathbf{D}}|^2}{2\sigma^2} + \ln P(\mathbf{b}) \right\} - \max_{\mathbf{b}_{-k,1}} \left\{ \frac{-|\tilde{\mathbf{Y}}-\tilde{\mathbf{H}}\tilde{\mathbf{D}}|^2}{2\sigma^2} + \ln P(\mathbf{b}) \right\}\end{aligned} \tag{3}$$

where $\lambda_{1,k}^{e}$ the extrinsic LLR of a particular bit b_k, $\mathbf{b}_{-k} = (b_1, \cdots, b_{k-1}, b_{k+1}, \cdots, b_p)$, $\mathbf{b}_{-k,a} = (b_1, \cdots b_{k-1}, b_k = a, b_{k+1}, \cdots, b_p)$; $b_i \in \{0, 1\}$,

$$\tilde{\mathbf{Y}} = \begin{bmatrix} \mathbf{Y}^{(0)} \\ \vdots \\ \mathbf{Y}^{(R)} \end{bmatrix}, \tilde{\mathbf{H}} = \mathrm{Diag} \begin{bmatrix} \mathbf{H}^{(0)} \\ \vdots \\ \mathbf{H}^{(R)} \end{bmatrix}, \tilde{\mathbf{D}} = \begin{bmatrix} \mathbf{d}^{(0)} \\ \vdots \\ \mathbf{d}^{(R)} \end{bmatrix}$$

$\mathbf{Y}^{(r) \in C^{rx \times 1}}, \mathbf{H}^{(r)} \in C^{rx \times tx}, \mathbf{d}^{(r)} \in C^{tx \times 1}$ denotes received vectors, channel matrix and transmitted vectors at the r-th retransmission

The second line of (4) is a max-log approximation version of MAP detection (max-log MAP). Hereinafter, we only consider the max-log MAP detection. It is not difficult to find that $p = N_t M_c$ for MIMO system, where M_c is the number of bits for a constellation.. The complexity of cumulative MAP detector is increased exponentially with p.

B. Post-combining scheme

Instead of performing combining before MIMO detection, Post-combining performs combining after MIMO detection. As shown in Fig. 2 (b) , post-combining schemes calculates the LLR values individually in each retransmission as normal system without HARQ and then combines those soft values to get the final LLRS. Specifically, the computation of LLRs is shown as following:

$$\lambda_{1,k}^{e} = \sum\nolimits_{r=1}^{R} \ln \frac{P(b_k=+1|\mathbf{Y}_r,\mathbf{H}_r)}{P(b_k=-1|\mathbf{Y}_r,\mathbf{H}_r)} = \sum\nolimits_{r=1}^{R} \ln \frac{\sum\limits_{\mathbf{b}_{-k}} P(b_k=+1,\mathbf{b}_{-k}|\mathbf{Y}_r,\mathbf{H}_r)}{\sum\limits_{\mathbf{b}_{-k}} P(b_k=-1,\mathbf{b}_{-k}|\mathbf{Y}_r,\mathbf{H}_r)} \tag{4}$$

Pre-combining and post-combining schemes have their own advantages. Pre-combining performs combining before demodulation, no information is lost. Therefore, it has the best performance. However, it needs to store received vector and channel matrix in previous transmissions. Post-combining scheme, on the other hand, has sub-optimal performance but requires fewer storage. In addition, pre-combining scheme is only applicable when the same packet of bits are transmitted in each retransmission. Post-combining, however, can be applicable for both HARQ-CC and HARQ-IR systems.

V. Reduce complexity pre-combining

The complexity of MAP detector for a MIMO system, even with single transmission, is very high. Therefore, various low complexity MIMO detector has been proposed, see [8] and the reference therein for a review of MIMO detection algorithms. Those algorithms can be directly applied to post-combining scheme. However, for pre-combining scheme, especially when the bit rearrangement is adopted, more care will be taken.

When retransmission occurs, the combining scheme perform detection over the correlated bits based on several versions of received signals from multiple retransmissions. The euclidean distance (ED) for multiple transmission can be represented by

$$|\tilde{\mathbf{Y}} - \tilde{\mathbf{H}}\tilde{\mathbf{D}}|^2 = \sum_{r=0}^{R} |\mathbf{Y}^{(r)} - \mathbf{H}^{(r)}\mathbf{d}^{(r)}|^2 \tag{5}$$

The (5) may suggest to use R MIMO detectors in parallel to solve the problem. If this works, then any MIMO detectors can be applied to the HARQ system directly. However, since multiple transmission vector $\mathbf{d}^{(r)}$ represent the same bits for HARQ-CC, we can not perform parallel MIMO detectors *independently*.

Therefore, some modification needs to be done especially when bit rearrangement is used for retransmission. There are two cases to consider. 1) bit-to-symbol alignment is preserved, that is, each symbol in multiple transmission represents the same bits (maybe in different form). Such case happens when no bit rearrangement or only BSI is used for bit rearrangement. 2) bit-to-symbol alignment is not fixed, that is, the bits in one symbol may be spread to several different symbols in another transmission. For example, in BSA scheme, d_1 contains bit $i_{1,1}i_{1,2}q_{1,1}q_{1,2}$ at initial transmission and contains bit $i_{1,1}i_{2,2}q_{1,1}q_{2,2}$ at the first retransmission.

A. Pre-combining based on QRD-M

For the first case, [9] propose a parallel sphere decoder with a special candidate enumeration unit to solve the problem. Here we present a parallel QRD-M algorithm with only minor modification.

For QRD-M, we need to search the minimal of (5) over the tree. To do this, First, we perform the QR decomposition over the channel matrix $\mathbf{H}^{(r)}$ such that $\mathbf{H}^{(r)} = \mathbf{Q}^{(r)}\mathbf{R}^{(r)}$, $\mathbf{Q}^{(r)} \in C^{N_r \times N_r}$ is a unitary matrix and

$$\mathbf{R}^{(r)} = \begin{bmatrix} \mathbf{T}^{(r)} \\ \mathbf{0}_{N_r - N_t, N_t} \end{bmatrix}$$

where $\mathbf{T}^{(r)} \in C^{N_t \times N_t}$ is an upper-triangular matrix. Then we have

$$\mathbf{r}^{(r)} = \mathbf{Q}^{(r)H}\mathbf{y}^{(r)} = \mathbf{Q}^{(r)H}\mathbf{H}^{(r)}\mathbf{d}^{(r)} + \mathbf{Q}^{(r)H}\mathbf{n} = \mathbf{R}^{(r)}\mathbf{d}^{(r)} + \mathbf{Q}^{(r)H}\mathbf{n} \tag{6}$$

where $(\cdot)^H$ denotes the Hermitian operator. For ease of disposition, we assume that $N_t = N_r$ (thus $\mathbf{R} = \mathbf{T}$) hereafter.

With QR decomposition, R parallel trees are formed. Since bit-to-symbol alignment is fixed, we could arrange the parallel nodes in the trees corresponding to the symbols with same bits. In this way, we could run a parallel QRD-M search in the parallel trees. We first illustrate the parallel QRD-M algorithm using parallel trees shown in Fig. 3, assuming that $N_t = N_r = 4, M_c = 2, M = 2$ and $R = 2$. The nodes at the top represents the transmitted symbol $d_4^{(r)}$. Since $d_4^{(r)}$ may take 4 different values, there are 4 branches connected to the $d_4^{(r)}$ node, each representing a possible value of $d_4^{(r)}$. These branches are connected to the nodes $d_3^{(r)}$ at the next layer, etc. Hence, each path in the tree corresponds to a transmitted sequence $(d_4^{(r)}, d_3^{(r)}, d_2^{(r)}, d_1^{(r)})$. We define the path metric associated with each path as

$$\begin{aligned} |\omega^{(r)}|^2 &= \sum_{k=1}^{N_t} |r_k^{(r)} - \sum_{l=k}^{N_t} t_{k,l}^{(r)} d_l^{(r)}|^2 - \tfrac{2\sigma^2}{R} \sum_{k=1}^{N_t} \log p(d_k^{(r)}) \\ &= \sum_{k=1}^{N_t} |(r_k^{(r)} - t_{k,k}^{(r)} d_k^{(r)}) - \sum_{l=k+1}^{N_t} t_{k,l}^{(r)} d_l^{(r)}|^2 - \tfrac{2\sigma^2}{R} \sum_{k=1}^{N_t} \log p(d_k^{(r)}) \end{aligned} \tag{7}$$

where $r_k^{(r)}, t_{k,l}^{(r)}, d_l^{(r)}$ are the entries of $\mathbf{r}^{(r)}, \mathbf{T}^{(r)}, \mathbf{d}^{(r)}$, respectively, and $p(d_k^{(r)})$ is prior probability of symbol $d_k^{(r)}$ provided by SISO decoder. Note in last term we divide the log prior probability by R to ensure the prior probablities are counted only once in the whole path metric.

In parallel QRD-M, we compute the accumulated path metric $\lambda_i^{(r)}$ at each tree in parallel. The M branches are chosen from branches from the branches with smallest The parallel QRD-M reduces the tree search complexity by keeping only M branches that have the smallest accumulated path metric $\sum_{r=0}^{R} \lambda_i^{(r)}$ at each tree level i, which is basically a breath-first tree search. For example, as shown in Fig. 3, at level 1, only 2 out of 4 possible $d_4^{(r)}$ with smallest

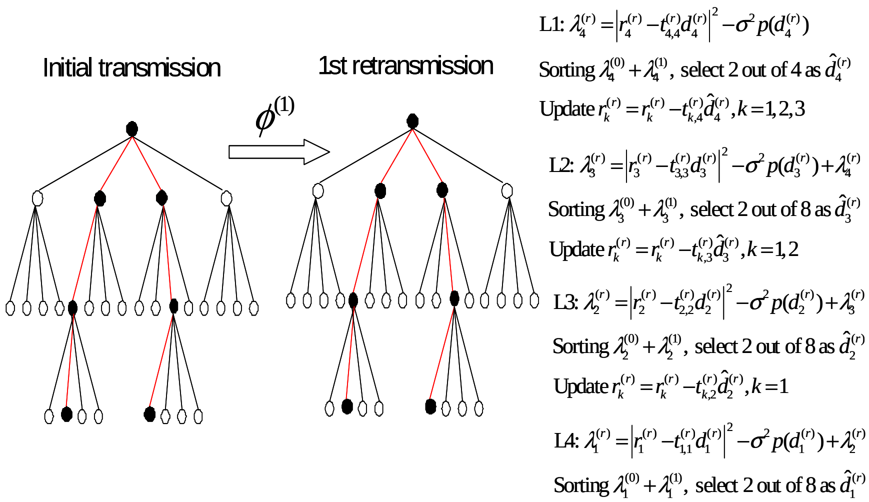


Fig. 3. Parallel QRD-M $N_t = 4, M_c = 4, M = 2, R = 2$. $\lambda_i^{(r)}$ denotes the accumulated path metric corresponding to the transmitted sequence $(d_4^{(r)}, d_3^{(r)}, \cdots, d_1^{(r)})$ at the r-th transmission.

$\lambda_4^{(r)}) = |r_4^{(r)} - t_{4,4}^{(r)} d_4^{(r)})|^2 - 2\sigma^2 \log p(d_4^{(r)})/2$ associated with $d_4^{(r)})$ are selected. With each selected $\hat{d}_4^{(r)}$, we update $r_k^{(r)} = r_k^{(r)} - t_{k,4}^{(r)} \hat{d}_4^{(r)}, 1 \leq k \leq 3$ which are used to compute the accumulated path metric $\lambda_3^{(r)}$ at the next level. This computation is carried out level by level until level 4 to yield $\lambda_1^{(r)}$ which equals the path metric defined in (7). Finally, we obtain two paths with the smallest path metric which form the important sets $\mathcal{I}_Q$. Those sequences in the important sets are used to compute the bitwise LLR as input to the soft channel decoder.

It is not hard to see that parallel QRD-M can only work for the case when bit-to-symbol alignment is fixed to ensure the same bits are processed at each level of the parallel trees. Similarly, other tree search based detectors have the same problem. For more general bit-rearrangement schemes, we need to consider other solution.

B. Pre-combining based on MCMC

Bit-wise MCMC can avoid this problem. The MCMC detector uses Gibbs sampler to generate a list of L most likely transmitted vectors. The basic idea of MCMC detector is The detail description of MCMC MIMO detector is given in [11]. In MCMC detector, a key step is to generating new sample (new state $\mathbf{b}^{[n]}$) using conditional probability $p(\mathbf{b}^{[n]}|\mathbf{b}^{[n-1]}, \mathbf{Y})$ as state transition probability from the old sample (old state $\mathbf{b}^{[n-1]}$). Bit-wise MCMC allows only one bit change between the old and new state (say b_i between $\mathbf{b}^{[n-1]}$ and $\mathbf{b}^{[n]}$). In applying MCMC detector to HARQ, the conditional probability is computed as following:

$$\begin{aligned} &P(\mathbf{b}^{[n]}|\mathbf{b}^{[n-1]}, \tilde{\mathbf{Y}}, \tilde{\mathbf{H}}) = P(b_i = b|\mathbf{b}_{-i}^{[n-1]}, \tilde{\mathbf{Y}}, \tilde{\mathbf{H}}) \\ &\propto p(\tilde{\mathbf{Y}}|\mathbf{b}_{-i}^{[n-1]}, b_i = b, \tilde{\mathbf{H}}) P(b_i = b) \end{aligned} \tag{8}$$

Let $\mathbf{d}^{[n-1,i],(r)} = \psi\{\phi^{(r)}(\mathbf{b}_{-i}^{[n-1]}, b_i = b)\}$, the calculation of (8) is equivalent to compute

$$\sum_{r=0}^{R} |\mathbf{Y}^{(r)} - \mathbf{H}^{(r)}\mathbf{d}^{[n-1,i],(r)}|^2 \tag{9}$$

Therefore, the application of MCMC detector to HARQ is equivalent to run Gibbs sampler with modified ED calculation. To show how the Gibbs sampler works for HARQ system, we examine the example shown in Table III where $R = 2, N_t = 2$ and 4PAM is used. Thus the number of bits in one MIMO detection is 4. We put the two bits mapped to one symbol in a pair of curly brackets. The bit rearrangement function is defined as $\phi^{(1)}(\{b_0, b_1, b_2, b_3\}) = \{b_0, b_3, b_2, b_1\}$. At time 0, the samples are initialized using some methods to obtain $b_k^{[0]}$. At time i, the Gibbs sampler updates the j-th bit of the symbol sequence in the initial transmission based on other bits obtained before according to the conditional probability of (8) . The corresponding symbol sequence for the 1st transmission is obtained through bit rearrangement function. Note we do not put any constraint on the bit rearrangement function. Therefore, MCMC detector can be applied to any bit rearrangement strategies.

TABLE III

SAMPLES UPDATING IN GIBBS SAMPLER FOR HARQ

Iteration n	Time t	Update sample	Initial transmission $\overset{\phi}{\Rightarrow}$ 1st transmission	
0	0		$\{b_0^{[0]}b_1^{[0]}\}, \{b_2^{[0]}b_3^{[0]}\}$	$\{b_0^{[0]}b_3^{[0]}\}, \{b_2^{[0]}b_1^{[0]}\}$
1	1	b_0	$\{b_0^{[1]}b_1^{[0]}\}, \{b_2^{[0]}b_3^{[0]}\}$	$\{b_0^{[1]}b_3^{[0]}\}, \{b_2^{[0]}b_1^{[0]}\}$
	2	b_1	$\{b_0^{[1]}b_1^{[1]}\}, \{b_2^{[0]}b_3^{[0]}\}$	$\{b_0^{[1]}b_3^{[0]}\}, \{b_2^{[0]}b_1^{[1]}\}$
	3	b_2	$\{b_0^{[1]}b_1^{[1]}\}, \{b_2^{[1]}b_3^{[0]}\}$	$\{b_0^{[1]}b_3^{[0]}\}, \{b_2^{[1]}b_1^{[1]}\}$
	4	b_3	$\{b_0^{[1]}b_1^{[1]}\}, \{b_2^{[1]}b_3^{[1]}\}$	$\{b_0^{[1]}b_3^{[1]}\}, \{b_2^{[1]}b_1^{[1]}\}$

C. Pre-combining based on QRMC

While bitwise MCMC detector can be easily modified for HARQ system, it has some problems at high SNR region which have been observed in previous literature [8], [12]. In [12], a MMSE initialized MCMC (MMSE-MC) is proposed to partially mitigate the high SNR problem. A more complete solution called hybrid QRD-M MCMC detector is suggested in [8]. Therefore, here we also apply the QRMC to the the pre-combining scheme for MIMO HARQ.

The main idea of QRMC detector is to perform QRD-M detection with small M first, followed by MCMC detection. The most likely transmitted vector found by the QRD-M algorithm is used as one of the initial vectors to initiate MCMC detection.

To make hybrid QRMC work for general bit rearrangement schemes, we first run a QRD-M detector just over the current transmission. The obtained bit sequence with minimal ED will be used to initialize the MCMC detector. The MCMC dector is then run over the multiple transmissions according to the description above. In this way, we perform a near optimal pre-combining scheme to improve the performance of HARQ MIMO-OFDM system.

VI. Performance results

In this section, we evaluate the performance of various bit rearrangement strategies using proposed pre-combing receiver and compare it with the QRD-M based post-combining scheme. Table IV shows the system parameter of a IEEE 802.16e system we followed in the simulation. The bit rearrangement strategies we studied are given in Table V which are basically different combinations of BRFs presented in section III. Note the strategies 2 is the one proposed in [7].

Figure **??** depicts the bit error rate (BER) versus the SNR for a ITU Pedestrian B channel with mobile speed 3km/h. It represents a typically slow fading channel. It shows the significant gain (6dB maximal) has been obtained using proposed pre-combining receiver. Moreover, the strategy 5 achieves the best performance. Using proposed combining scheme, it provides about 7.5dB gain over the strategy 1 (no bit rearrangement) and 0.5dB gain over strategy 2, i.e, SSA or SSC provide additional gain in a slow fading environment. Also the gap between strategies 3, 4, 5 is negligible. It shows the combining of SSA and SSC does not improve the performance than using the SSA and SSC alone.

Figure **??** shows the BER performance for the ITU Vehicular A channel with mobile speed 120km/h. This channel changes much faster and is a fast fading channel. Again, the proposed

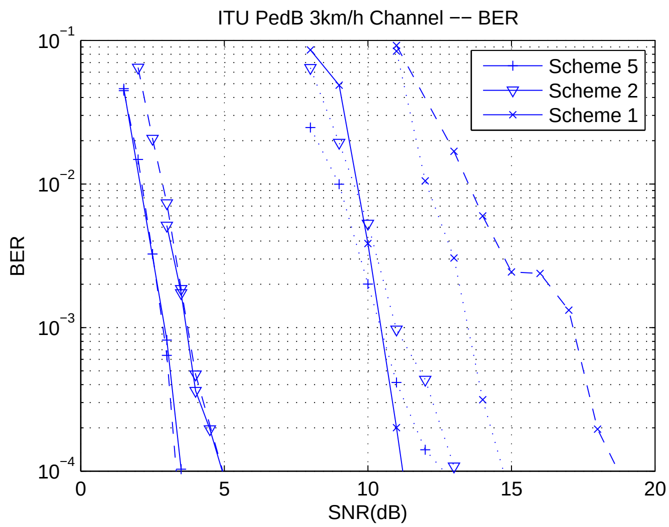


Fig. 4. Performance comparison of various HARQ schemes for ITU Ped B channel with speed 3km/h (solid line uses QRMC pre-combining, dashed line uses MMSE-MC precombining and dotted line uses QRD-M bitwise post-combining)

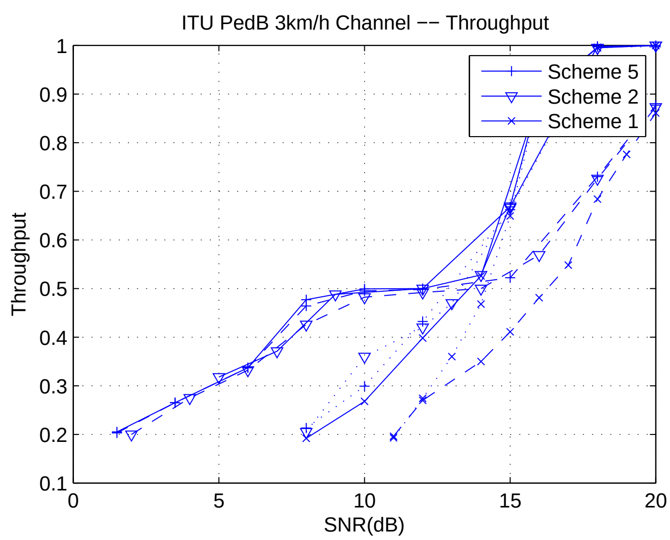


Fig. 5. Performance comparison of various HARQ schemes for ITU Ped B channel with speed 3km/h (solid line uses QRMC pre-combining, dashed line uses MMSE-MC precombining and dotted line uses QRD-M bitwise post-combining)

TABLE IV

SYSTEM PARAMETERS

Parameter	Value
Channel bandwidth	10 MHz
Number of subcarriers	1024
Subcarrier permutation	PUSC
Cyclic prefix	1/8
Channel coding	Convolutional turbo codes
Modulation	16QAM
MIMO mode	4x4 Spatial multiplexing
Multipath channel	ITU PedB/VehA
MS speed	3.6/120 *km/hr*
H-ARQ type	Type I (CC)
# of retransmissions	4

TABLE V

BIT REARRANGEMENT SCHEMES

Strategies	Strategy NO.
No bit rearrangement	1
BSI & BSA	2
BSI & BSA & SSA	3
BSI & BSA & SSC	4
BSI & BSA & SSA & SSC	5

combining scheme achieves significant gain (3dB) compared with QRD-M based post-combining scheme but the gain is much smaller than the slow fading scenario. In addition, no additional gain can be obtained by introducing SSA or SSC in fast fading channel.

Comparing the Figure **??** and **??**, we can see almost the same performance is obtained in both channels when proper bit rearrangement strategies are used. It implies that using proper bit arrangement strategies can compensate the diversity loss due to the slowing time-varing of the channel. .

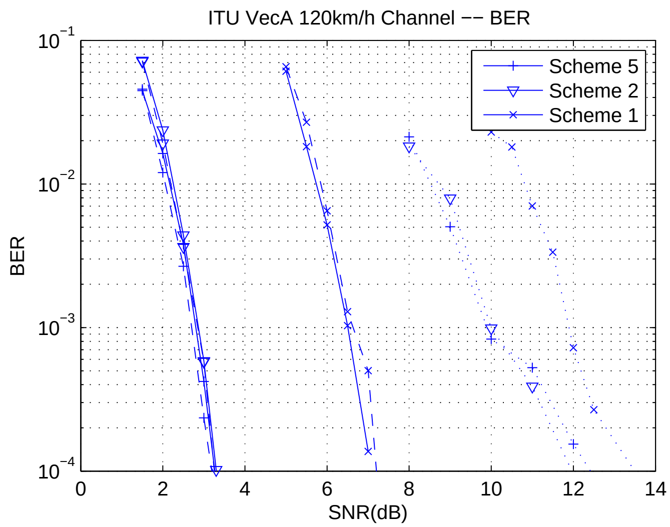


Fig. 6. Performance comparison of various HARQ schemes for ITU Vec A channel with speed 120km/h (solid line uses QRMC pre-combining, dashed line uses MMSE-MC precombining and dotted line uses QRD-M bitwise post-combining)

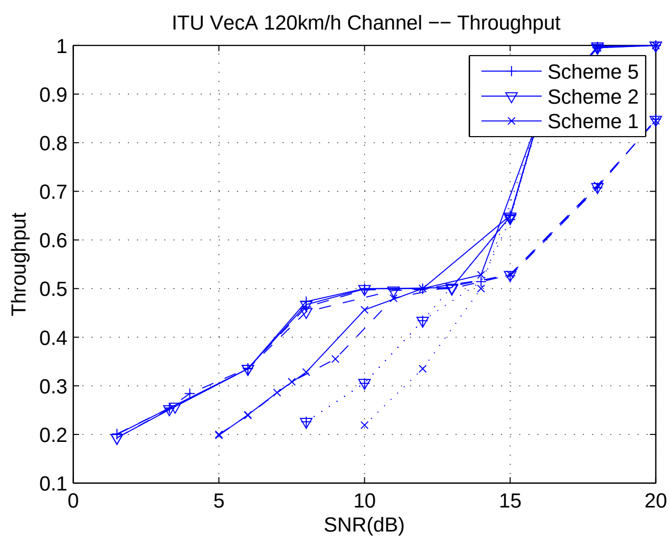


Fig. 7. Performance comparison of various HARQ schemes for ITU Vec A channel with speed 120km/h (solid line uses QRMC pre-combining, dashed line uses MMSE-MC precombining and dotted line uses QRD-M bitwise post-combining)

VII. Conclusion

In this paper, we propose a new near optimal pre-combining scheme for type I HARQ in MIMO-OFDM system. The proposed combining scheme is applied to various bit rearrangement strategies. The significant gain has been obtained for both slow and fast fading channels comparing with the system using near optimal QRD-M detector as a bitwise post-combining scheme at the cost of increased storage requirement. The performance of various bit rearrangement strategies using the proposed the combining scheme is also evaluated and compared. It is shown that proper bit-rearrangement can greatly improve the system performance even in slow fading channel. The new combining scheme has flexible structure and can be applied to any bit rearrangement strategies which in general do not work using other known pre-combining scheme. Due to its flexible structure, it is also very convenient to be applied to those systems using adaptive modulation during retransmissions.

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