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Re:	Call for Comments and Contributions on Project 802.16m System Description Document (SDD) issued on 2008-06-16 (IEEE 802.16m-08/024) Topic: Hybrid ARQ (PHY Aspects)
Abstract	This contribution illustrates Hybrid ARQ physical layer architecture and performance enhancement for high mobility scenario. Five performance enhancement methods are described: 1, increasing the number of convolutional turbo code block sizes; 2. modifying bit selection method for HARQ IR; 3. constellation rearrangement; 4. joint MIMO HARQ scheme; 5. symbol repetition in high mobility.
Purpose	For discussion and approval by IEEE 802.16m TG
Notice	
Release	
Patent Policy	

The HARQ Physical Layer Architecture

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1 Introduction

This contribution provides a general HARQ physical layer architecture. HARQ improves reception performance by incorporating previous incorrectly-decoded bursts and the new burst to process. In IEEE 802.16e [1], Chase combining and incremental redundancy are introduced in HARQ mechanism. However there are still channel coding issues and performance enhancement methods for IEEE 802.16e HARQ scheme. The proposed architecture covers these performance enhancement methods.

Five HARQ physical layer performance enhancement methods are described in this contribution.

1. Increasing the number of block sizes reduces throughput loss. The IEEE 802.16e convolutional turbo code (CTC) block sizes are few. When the MAC PDU does not fit one of the block sizes, padding dummy bits to fit block size is used and it causes throughput loss. This method provides at least 10% throughput improvement.
2. Modifying IEEE 802.16e CTC IR bit selection method provides extra performance gain. The proposed selection method provides extra 0.2dB performance gain comparing to IEEE 802.16e HARQ IR symbol selection mechanism.
3. Changing constellation and carrier mapping on adjacent transmissions improves error rate performance. Our example indicates 5dB performance gain.
4. MIMO technologies can provide diversity gain and spatial-multiplexing gain. Robust MIMO scheme, STBC/SFBC, could be applied for retransmissions to acquire better diversity gain, i.e. link quality. 1 dB performance gain is achievable while 42% memory reduction on symbol level combing is available.
5. In high mobility, symbol repetition benefits system throughput due to accurate inter-carrier interference cancellation method applicable. This method provides 5dB performance gain at 350km/hr.

2 HARQ physical layer architecture

Fig. 1 illustrates the proposed HARQ architecture. In IEEE 802.16e [1], HARQ mechanism is mainly classified into two categories: Chase combining (CC) and incremental redundancy (IR). CC features less storage of multiple redundancy version transmissions due to identical retransmitted redundancy version. IR renders better performance due to more parity bits and less systematic bits transmitted. In order to achieve better performance and higher throughput based on these two categories, some mechanisms or methods are introduced.

First, we propose to increase the number of CTC block sizes to reduce extra redundant bits and reach better bandwidth efficiency and higher throughput. Second, shifted IR redundancy version selection is proposed to enhance the performance of IR. Third, bit rearrangement re-permutes transmitted redundancy version to provide better performance for CC. Fourth, in different transmission and retransmissions, MIMO scheme could be different to achieve better performance. Fifth, if the mobility is high, symbol repetition achieves better channel estimation performance. Following sections will describe these mechanisms.

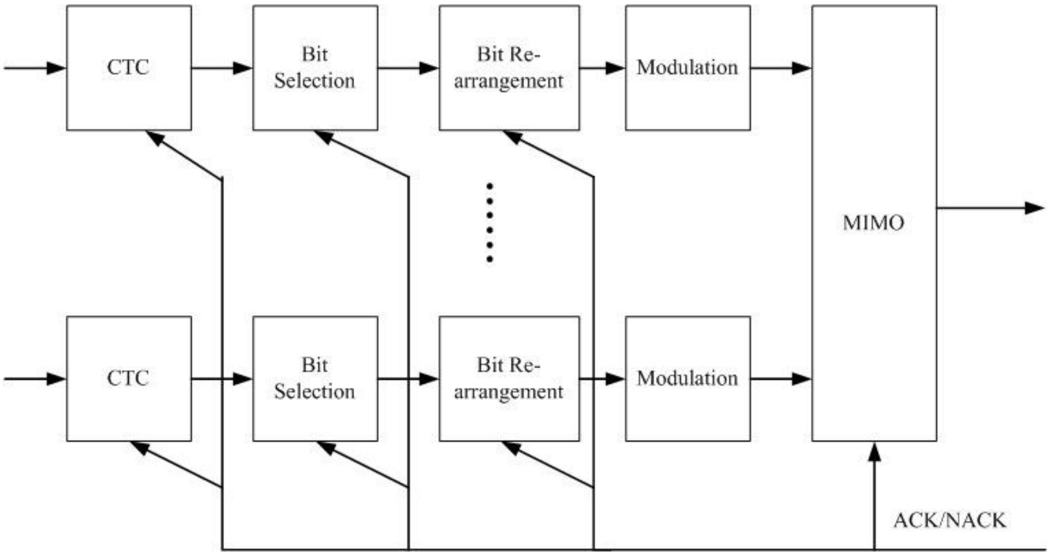


Fig. 1: HARQ physical layer architecture.

3 IEEE 802.16e CTC encoding and the associated subpacket generation method

In IEEE 802.16 [1], CTC applies double binary circular recursive systematic convolutional code and Fig. 2 shows the CTC encoder. The input symbol of this code is composed of a bit pair (A_i, B_i), where A_i and B_i are the *i*th bits of the input sequences *A* and *B*. The output are *A*, *B* ,Y_1 ,Y_2 ,W_1 ,W_2. Y_1 and W_1 correspond to the input sequences *A* and *B*; Y_2 and W_2 correspond to the permuted input sequences *A* and *B*.

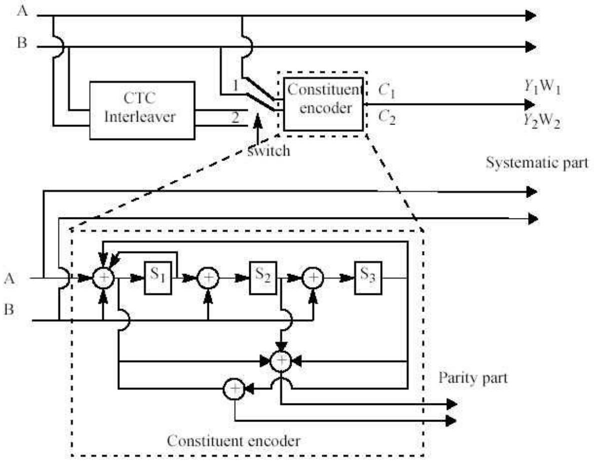


Fig. 2: Double binary turbo code.

Fig. 3 plots subpacket generation method. This method generates code sequence by the CTC encoder. Then the channel interleaver permutes the generated code sequence. After interleaving, the puncturing block selects the associated selected symbols. Fig. 4 shows the channel interleaving. This channel interleaving applies subblock interleaving on sequences A, B ,Y_1 ,Y_2 ,W_1 ,W_2, respectively. Then subblock interleaved sequences Y_1 and Y_2 are inter-block permuted into sequences Y_1' and Y_2' and subblock interleaved sequences W_1 and W_2 are inter-block permuted into sequences W_1' and W_2'. Puncturing block selects symbols according to the sequence order A, B ,Y_1' ,Y_2' ,W_1' ,W_2', where Table 1 shows the parameters and Table 2 shows the selected symbols.

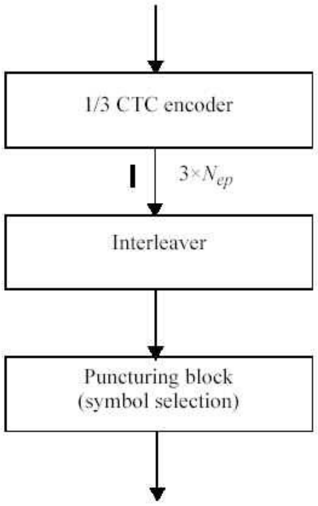


Fig. 3: Subpacket generation method.

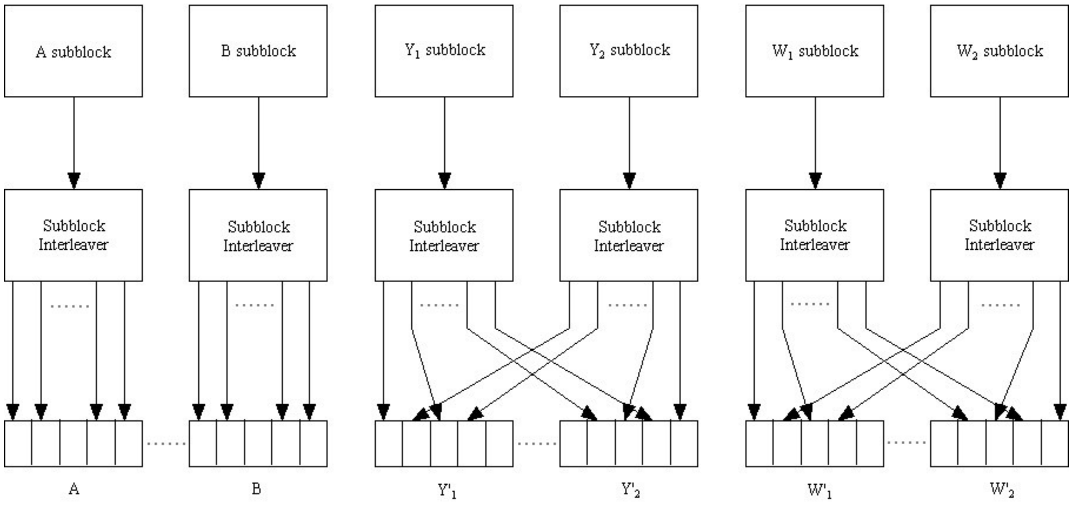


Fig. 4: Channel interleaver.

H-ARQ mechanism chooses transmission bits from the puncturing block shown in Fig. 3. If NACK is received, HARQ will choose the redundancy version from the block. If identical redundancy version is chosen, the CC mechanism is applied; otherwise IR mechanism is applied.

Table 1: Parameters for the H-ARQ CTC in IEEE 802.16 8.4.9.2.3.4.4 [1].

k	be the subpacket index when HARQ is enabled. $k = 0$ for the first transmission and increases by one for the next subpacket. $k = 0$ when HARQ is not used. When there are more than one FEC block in a burst, the subpacket index for each FEC block shall be the same.

N_{EP}	be the number of bits in the encoder packet (before encoding).
N_{SCHk}	be the number of the concatenated slots for the subpacket defined in Table 560 for the non-HARQ CTC scheme defined in 8.4.9.2.3.1 and be the same as the *Nsch* that is indicated in the Allocation IE for the HARQ CTC scheme defined in 8.4.9.2.3.5.
m_k	be the modulation order for the *k*-th subpacket ($m_k = 2$ for QPSK, 4 for 16-QAM, and 6 for 64-QAM).
$SPID_k$	be the subpacket ID for the *k*-th subpacket, (for the first subpacket, $\text{SPID}_{k=0} = 0$).

Table 2: The symbol selection in IEEE 802.16 8.4.9.2.3.4.4 [1].

$$S_{k,i} = (F_k + i) \bmod (3N_{EP})$$

where

$$i = 1,2,\ldots, L_k - 1$$

$$L_k = 48 \cdot N_{SCHk} \cdot m_k$$

$$F_k = (SPID_k \cdot L_k) \bmod (3 \cdot N_{EP})$$

4 Increase HARQ performance through applying larger block size and finer granularity for CTC

The advantage of increasing the maximum transmitted bits per transmission is shown and applying larger code block size seems nature. Contribution [2] indicates 8.75% average throughput improvement coming from larger data block size of CTC. 3GPP LTE [3] applies turbo code with maximum block size 6144 bits to support high throughput per user by less HARQ channels. It also reduces MAC PDU overhead. However channel coding design in [1] introduces throughput and performance issues.

Throughput issue comes from the padding applied for HARQ CTC subpacket generation. Below cited two sections in [1].

- 8.4.9.2.3.5 IR HARQ support
  - The procedure of HARQ CTC subpacket generation is as follows: padding, CTC addition, fragmentation, randomization, and CTC encoding.
- 8.4.9.2.3.5.1 Padding
  - MAC PDU (or concatenated MAC PDUs) is a basic unit processed in this channel coding and modulation blocks. When the size of MAC PDU (or concatenated MAC PDUs) is not the element in the alloowed set for HARQ, ones are padded at the end of MAC PDU (or concatenated MAC PDUs). The amount of the padding is the same as the difference between the size of the PDU (or concatenated MAC PDUs) and the smallest element is the allowed set that is not less than the size of the PDU (or concatenated MAC PDUs) . The padded packet is input into the CRC encoding block.
  - The allowed set is {32, 80, 128, 176, 272, 368, 464, 944, 1904, 2864, 3824, 4784, 9584, 14384, 19184, 23984} bits.

We assume the length of MAC PDU is uniformly distributed. Under this assumption, Table 3 shows average 415 bits padded for maximum N_{EP} =4800 bits; it equivalent to average 17.02% resource wasted for the padding. Fig. 5: Throughput loss due to the padding. further shows the associated throughput loss corresponding to various MAC PDU lengths for maximum N_{EP}=4800 bits. In some cases, the throughput loss may be 50%. Due to lack of interleaver sizes, the throughput loss is significant.

Increasing the number of allowed set (number of interleavers) reduces the throughput loss and padded bits resulted form the padding. We provide extra 15 kinds of interleaver sizes ranging from 480 to 4800 bits. We also assume the length of MAC PDU is uniformly distributed. Table 3 shows average 113 bits padded for maximum N_{EP} =4800 bits; it equivalent to average 6.55% resource wasted for the padding. Fig. 5: Throughput loss due to the padding. further shows the associated throughput loss corresponding to various MAC PDU lengths for maximum N_{EP}=4800 bits. Above 480 bits, the throughput loss decreases a lot and the wasted resource is less.

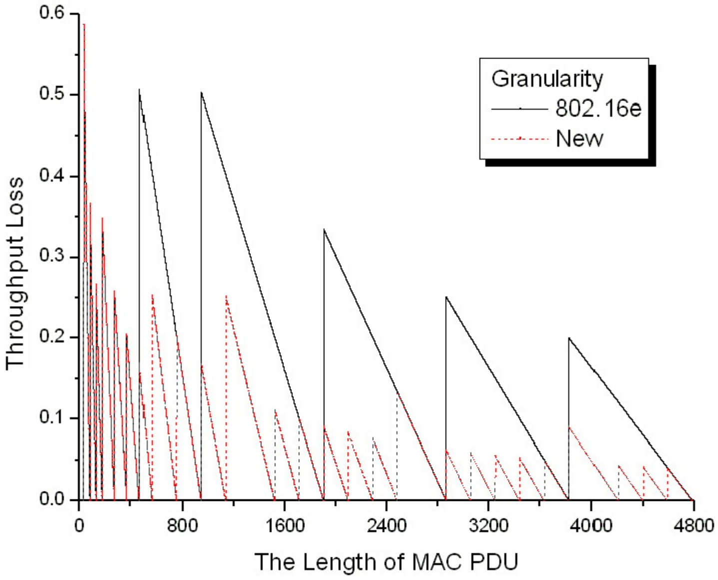


Fig. 5: Throughput loss due to the padding.

Table 3: Average padded bits and throughput loss corresponding to maximum N_{EP}=2400 symbols (4800 bits).

	IEEE 802.16e	New Pattern
Padded bits	415 bits	113 bits
Throughput loss	17.02%	6.55%

5 HARQ IR symbol selection

5.1 Shifted IR symbol selection mechanism

Parity bits of turbo coding provide better error correction capability. Systematic bits of turbo coding provide extrinsic information in the initial iterative decoding rounds. More systematic bits or larger power on systematic bits will make the iterative decoding fast converged. However the parity bits determine the minimum distance and error correction capability. When considered frame error rate is low, parity bits dominates error rate performance. For the HARQ, the first transmission will choose systematic bits. When IR is applied, the systematic bits would also be firstly chosen when all turbo code bits are selected as shown in Table 2. Since systematic bits are all selected in the first transmission and parity bits determine error correction capability, we propose first selecting parity bits after all parity bits are selected.

Table 2 shows the modified symbol selection method. This method shifts the selected symbol by X if complete codeword is selected, i.e. $S_{k,j} = (F_k + i + X)\mathrm{mod}(3N_{EP}),\ F_k + i \geq 3N_{EP}$. In general $X=N_{EP}$. We also can set $X=0.5N_{EP}$ or $1.5N_{EP}$. When $X=N_{EP}$, this method firstly selects parity bits to enhance the distance property if a complete codeword is selected. Our simulation will show the error rate performance.

Table 4: The modified symbol selection method.

$$S_{k,i} = \begin{cases} (F_k + i)\,\mathrm{mod}(3N_{EP}) & F_k + i < 3 \cdot N_{EP} \\ (F_k + i + N_{EP})\,\mathrm{mod}(3N_{EP}) & F_k + i \geq 3 \cdot N_{EP} \end{cases}, \text{ where}$$

$$i = 1,2,\ldots,L_k - 1$$

$$L_k = 48 \cdot N_{SCHk} \cdot m_k$$

$$F_k = (SPID_k \cdot L_k)\,\mathrm{mod}(3 \cdot N_{EP})$$

This method provides more kinds of redundancy versions comparing with IEEE 802.16e symbol selection method. If we consider the case $L_k=1.5\ N_{EP}$, IEEE 802.16e symbol selection method only provides two kinds of redundancy versions but our method can provide four kinds of redundancy versions. Our method gives transmitter more degree of freedom in selecting retransmission symbols.

5.2 Simulation results

This part evaluates the error rate performance for both symbol selection methods and relative parameters are shown in Table 5. Denote by IR1 and IR2 the IEEE 802.16 symbol selection method and our symbol selection method respectively. We assume $SPID_k$=0,1,2,3 for k=0,1,2,3. The compared symbol lengths per transmission are $2N_{EP}$, $1.5N_{EP}$ and $1.2N_{EP}$, where the code rates for the first transmission are 1/2, 2/3 and 5/6. N_{EP}=2400 symbols (4800 bits). We simulate the error rate performance of the 1st 2nd, for the cases $2N_{EP}$, $1.5N_{EP}$, respectively, and 2nd and 3rd retransmission for the cases $1.2N_{EP}$. The H-ARQ IR mechanism chooses the selected symbols corresponding to the kth transmission according to $SPID_k$. The simulation environment is AWGN channel. Log-MAP algorithm with 8 iterations is performed in our simulation.

Table 5: Simulation parameters.

Symbol length per transmission	Number of transmission	Mode	Equivalent rate
$2N_{EP}$	2	IR1	1/4
		IR2	1/4
$1.5N_{EP}$	3	IR1	2/9
		IR2	2/9

1.2N_{EP}	3	IR1	5/18
		IR2	5/18
	4	IR1	5/24
		IR2	5/24

Fig. 6-9 shows the error rate performance. Blue curve denotes IEEE 802.16 symbol selection method and red cure denotes the new symbol selection method. Obviously, that new symbol selection method provides 0.2-0.3 performance gain to the IEEE 802.16 symbol selection method at frame error rate=10^{-4}.

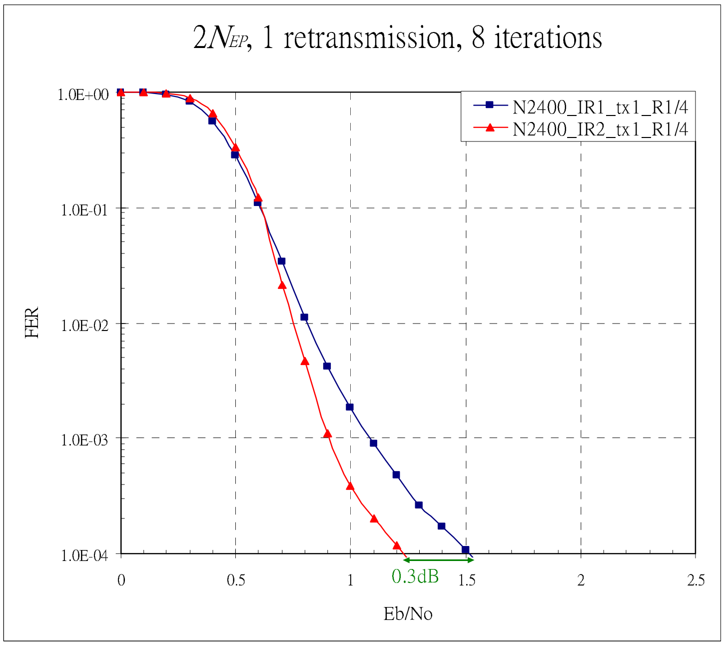


Fig. 6: N_{EP}=2400 with two transmissions whose length is 2 N_{EP}.

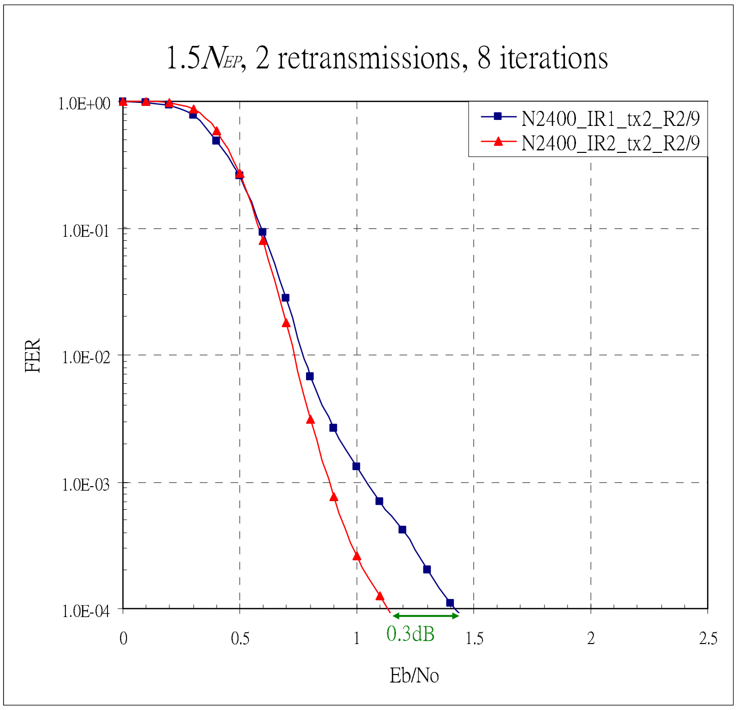


Fig. 7: N_{EP}=2400 with three transmissions whose length is 1.5 N_{EP}.

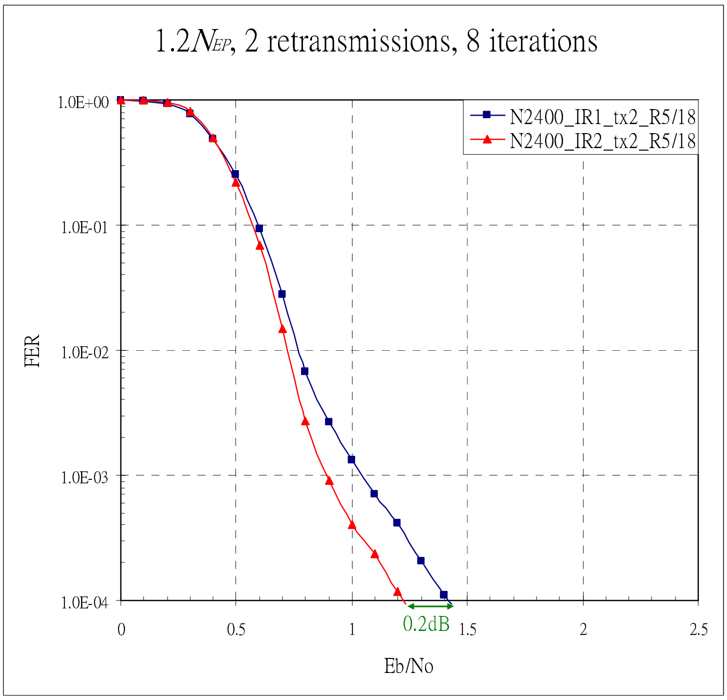


Fig. 8: N_{EP}=2400 with three transmissions whose length is 1.2 N_{EP}.

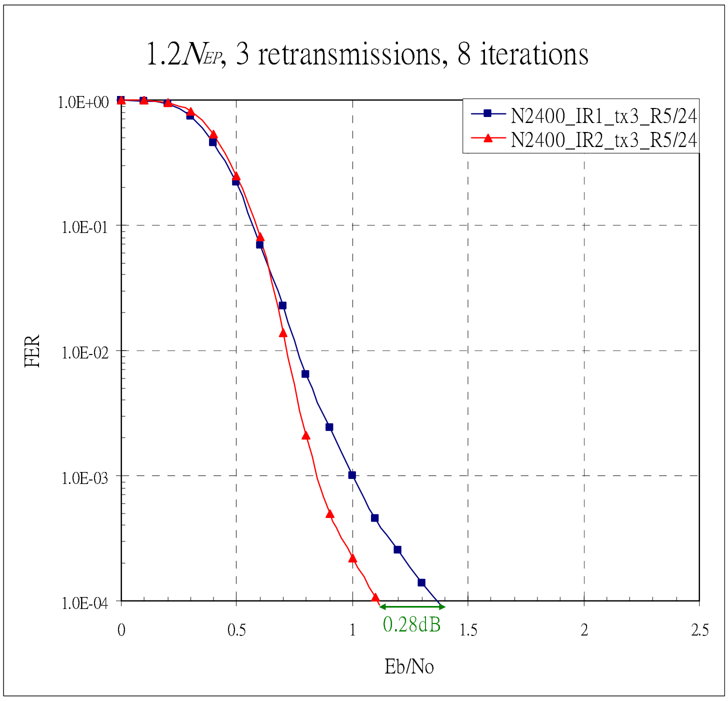


Fig. 9: N_{EP}=2400 with four transmissions whose length is 1.2 N_{EP}.

6 HARQ CC bit-rearrangement

6.1 Bit-rearrangement mechanism

HARQ can exploit both constellation and frequency diversity gain. Constellation rearrangement concept [4][5] is introduced to improve HARQ CC performance with 16QAM and 64QAM. This concept rearranges constellation at each retransmission. Therefore, the bits protected less can be protected well at the next retransmission. The constellation effect could be averaged and all code bits are more equally affected. However, the bits are restricted to be rearranged within a modulation symbol. Taking 16QAM for example, every four bits are rearranged to change each bit's reliability. Over the fading channel the constellation rearrangement just may supply constellation diversity, but not support the frequency diversity. In order to acquire frequency diversity gain, inter-symbol rearrangement is introduced.

Bit rearrangement method has both constellation and frequency diversity gain. This method rearranges coded bits across the whole subpacket. We introduce a bit rearranger before the modulator as shown in [4]. Bit rearranger performs bit-level interleaving and bit inverting. It applies various rearrangement patterns in different transmissions. Since bits are interleaved in different retransmissions, the mapping on the 16QAM and 64QAM are different in different retransmissions and protection on these coded bits would be more uniform. By the way, the coded bits mapped in the same symbol would be mapped to different symbol in the next transmission and frequency diversity gain is available.

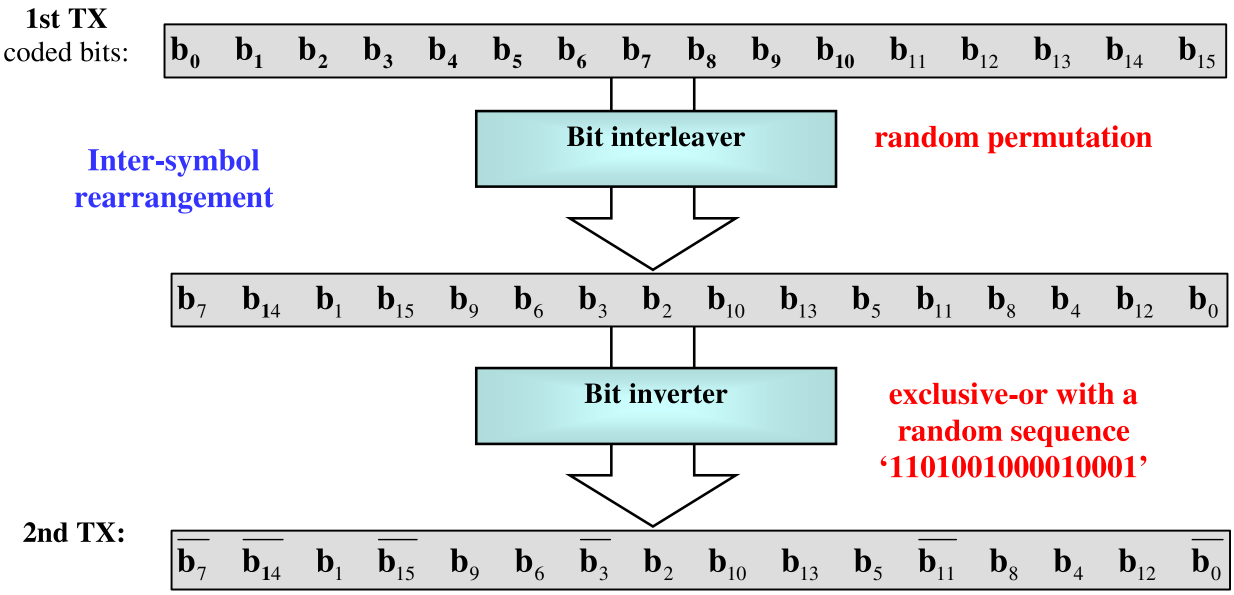


Fig. 10: An example of the bit rearrangement performed by bit rearranger.

An example of bit rearranger is shown in Fig. 10 and 16 bits sequence is used. The bit rearranger performs bit-level interleaving and bit inverting. The bit rearranger permutes these bits on bit level and inverts several bits which are labeled by bar on the symbol.

Every time, as transmitter received NACK from the receiver, transmitter applies different bit rearrangement method to retransmit the coded burst. Since each retransmission applies different bit rearrangement, coded bits may be mapped from LSB to MSB or MSB to LSB corresponding to each symbol mapping. The output SNR corresponding to different transmission would vary and it provides constellation diversity gain. By the way, the bit interleaver would map identical coded bit to different carriers on different retransmissions and it provides frequency diversity gain.

6.2 Simulation results

Fig. 11 and Fig. 12 illustrate some simulation results. CTC (convolutional turbo code) specified in IEEE 802.16e [1], 480 bits, code rate=1/2 with 16QAM and 64QAM is used. Typical urban fading channel is considered [6]. Bit interleaver and bit inverter are randomly selected in these results. The CTC decoder adopts linear Log-MAP and eight turbo decoding iterations. Fig. 11 shows that the bit rearrangement method outperforms the conventional HARQ CC with 16QAM about 1.7 dB, 2.5 dB and 3 dB at FER=10^{-3} after two, three and four transmissions respectively. Fig. 12 shows that the error performance of bit rearrangement surpasses the conventional HARQ CC with 64QAM about 2.7, 3.9, 4.6, 5.4 and 6 dB at FER=10^{-3} after two, three, four, five and six transmissions respectively. The proposed method significantly improves the capability of HARQ.

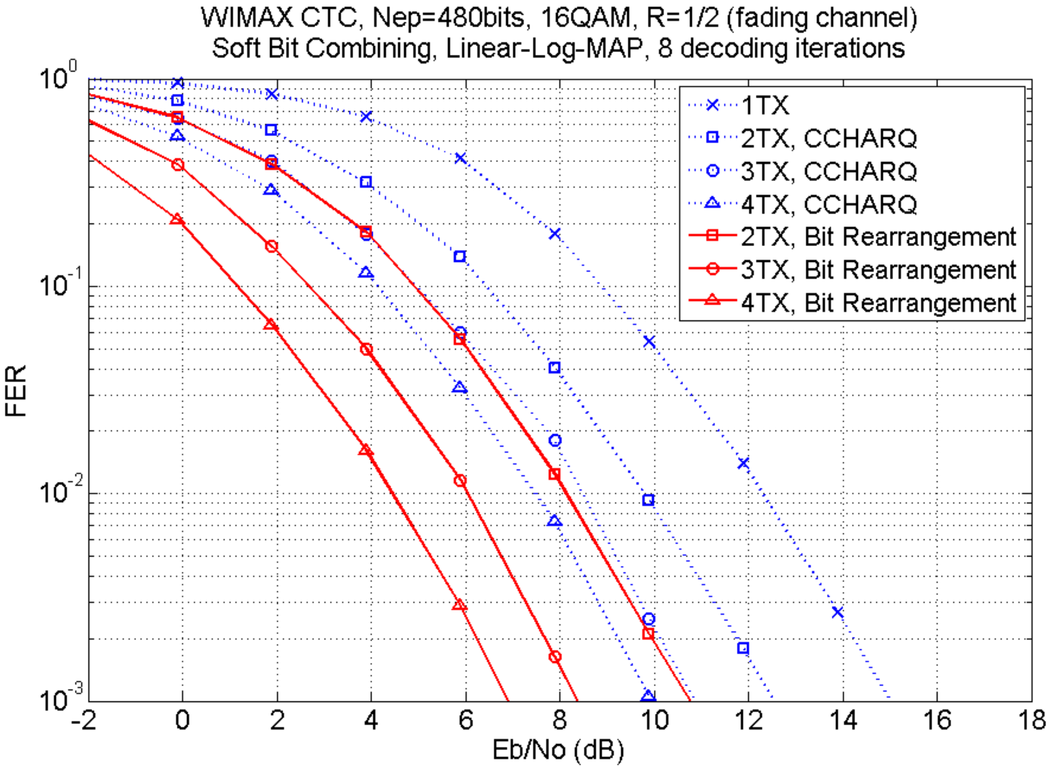


Fig. 11: Simulation result of the bit rearrangement for 16QAM over fading channel

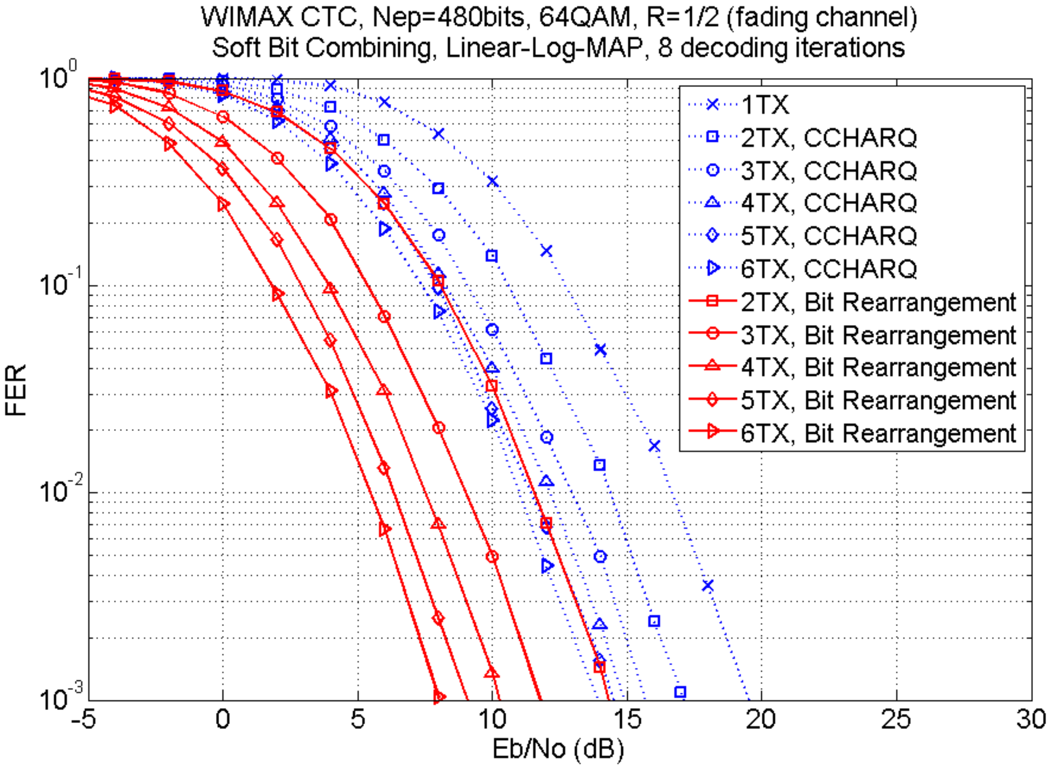


Fig. 12: Simulation result of the bit rearrangement for 64QAM over fading channel

7 Joint MIMO technology HARQ mechanism

In HARQ, the combining can be performed before or after detectors. Bit-level combining is a straightforward method after detectors. On the other hand, symbol-level joint detection is another approach for MIMO HARQ scheme and joint MIMO technology HARQ mechanism is also introduced in IEEE C802.16m-07/172 [7].

This section describes joint MIMO HARQ mechanism. This applies spatial multiplexing for the first transmission and STBC/SFBC for the retransmission. The scheme provides about 1dB performance gain when SNR ranges from 5-25dB. 42% memory reduction is also available when symbol-level combining is used.

7.1 Proposed Joint Symbol-Level Combining before MIMO Equalization

Fig. 15 illustrates the proposed joint MIMO HARQ scheme which exploits both throughput gain through spatial-multiplexing (SM) and diversity gain through STBC/SFBC. This HARQ scheme applies SM for the first transmission and STBC/SFBC for following retransmissions. SM improves HARQ transmission throughput and STBC/SFBC enhances link quality for the following retransmission; the associated storage is reduced by less retransmission.

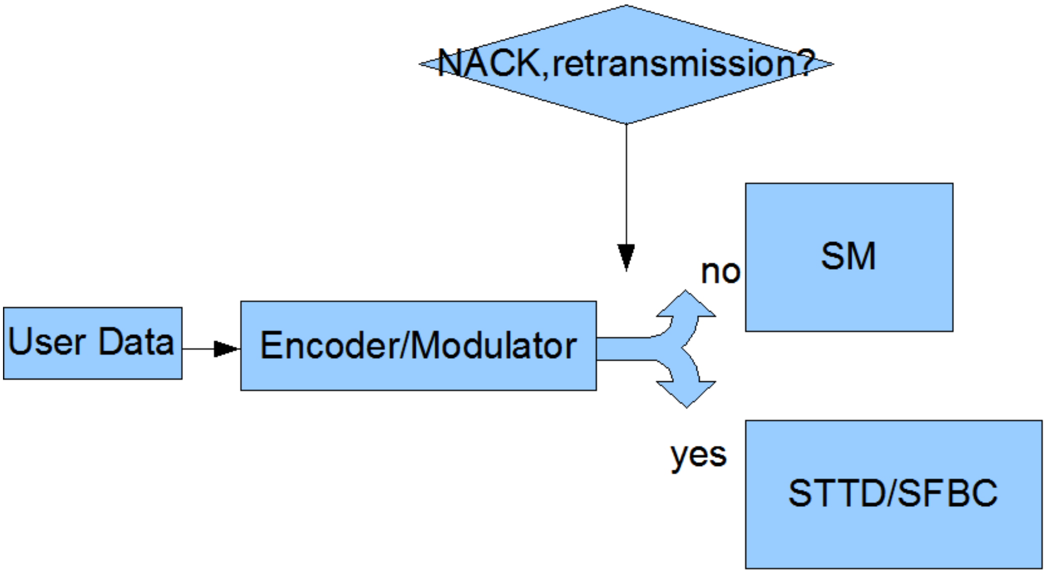


Figure 15. HARQ Transmission/Retransmission Block Diagram

The received signal model at the original transmission for MIMO is:

$$\mathbf{y}^0 = \mathbf{H}^0\mathbf{s}^0 + \mathbf{n}^0,$$

where

$\mathbf{y}^0$ is the received signal for the 0-th transmission

$\mathbf{H}^0$ is the MIMO channel for the 0-th transmission,

$\mathbf{s}^0$ is the transmit signal for the 0-th transmission, $\mathbf{s}^0 = \begin{bmatrix} s_0^0 \\ s_1^0 \end{bmatrix}$, where t and m of s_m^t stands for t-th transmission and m-th constellation signal, respectively.

$\mathbf{n}^0$ is the noise plus interference vector for the 0-th transmission.

The retransmission applies STBC/SFBC and half of information of the first transmission is transmitted at each retransmission. Every other retransmission will send different versions. Retransmitting half information improves spectrum efficiency and the other half information could be decoded through applying successive information cancellation (SIC) to first transmission.

The received signal is shown as follows. If the original transmission of a particular packet is $\mathbf{s}^0$, the re-transmitted packet $\mathbf{s}^1$ during two consecutive OFDMA symbols is:

$\mathbf{s}^1$ in symbol 0:

$$\mathbf{s}^1_{(symbol\ 0)} = \begin{bmatrix} s_0^1 \\ s_2^0 \end{bmatrix},$$

and $\mathbf{s}^1$ in symbol 1:

$$\mathbf{s}^1_{(symbol\ 1)} = \begin{bmatrix} -s_2^{0*} \\ s_0^{1*} \end{bmatrix}.$$

The corresponding joint detection blcok diagram is shown in Fig. 16.

Figure 16. Joint MIMO Detection Diagram

With this arrangement, the overall spectrum efficiency is the same as to retransmit the whole packet $\mathbf{s}^0$ again.

With this HARQ retransmission scheme, the required receiver complexity can be greatly reduced because only simple STBC/MRC or SFBC/MRC joint detector is used. We also demonstrate that with this retransmission scheme, the required buffer for symbol level joint detection can also be greatly reduced. The memory saving analysis will be shown later.

.While expanding to a 4 Tx and 4 Rx system, we suggest to retransmission with a rate-2 Matrix B with half information retransmission or a rate-1 Matrix A with one-forth information retransmission instead of a rate-4 Matrix C retransmission.

7.2 Memory Saving Analysis

This implementation scheme not only save the computation power (low complexity with STBC and MRC Equalizer in retransmission combining), but also requires less memory buffer. Traditional HARQ retransmits the same SM pattern and receiver may combine the signal in symbol level with the joint detector. The retransmitted signal and the original one should be all buffered. For example, if the maximum number of retransmission is 4 (including the original transmission), and the resource block is 4 OFDM symbols with 180 subcarriers in a 2x2 SM. Assuming each channel path element, received signal element and detected one (subcarrier) all counted as one unit. Moreover, one complex number will be counted as two units, and one real numerical number will be counted as one unit. Then the total required memory units are:

Received signal: (original transmission + first retransmission + second retransmission)

$$4(symbols)\times 180(sub)\times 2(rx)\times 3(orig.+retrans.)\times 2(I/Q)=8640\text{ units}$$

Channel response: (original transmission + first retransmission + second retransmission)

$$4(symbols)\times 180(subcarriers)\times 4(paths)\times 3(orig.+retrans.)\times 2(I/Q)=17280\text{ units}$$

Hence, the total required buffer size is: 8640+17280 = 25920 units.

Our proposal can only buffer the signal of the original transmission, retransmission of each time with just one symbol (only STBC) and the combined results.

Received signal: (original transmission +one symbol of each retransmission (STBC))

$$(4(symbols)\times 180(subcarriers)\times 2(rx)+1(symbols)\times 180(subcarriers)\times 2(rx))\times 2(I/Q)=3600\text{ units}$$

Channel response: (original transmission +one symbol of each retransmission (STBC))

$$(4(symbols)\times 180(subcarriers)\times 4(paths)+1(symbols)\times 180(subcarriers)\times 4(paths))\times 2(I/Q)=7200$$
units

Equalized channel response of STBC/MRC or SFBC/MRC: (whole packet size)

$$4(symbols)\times 180(subcarriers)\times 2(rx)=1440\text{ units}$$

Equalized received signal of STBC/MRC or SFBC/MRC: (whole packet size)

$$4(symbols)\times 180(subcarriers)\times 2(rx)\times 2(I/Q)=2880\text{ units}$$

Hence the total required buffer memory is 3600+7200+1440+2880 = 15120 units. The saved memory is about:

$$\frac{25920-15120}{25920}=42\%$$

7.3 Simulation Results

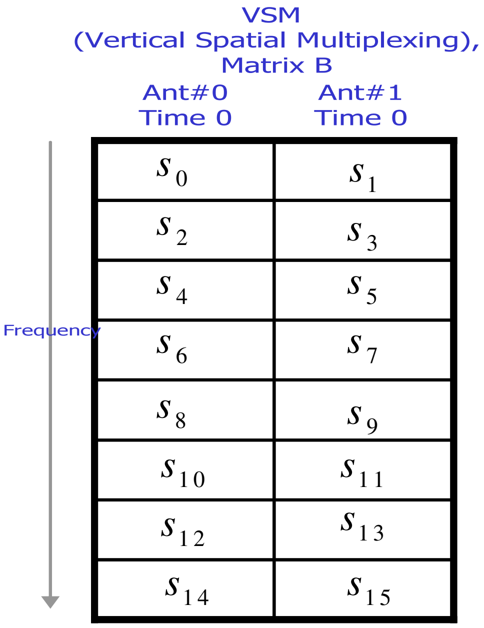
In the contribution, simulation results for Spatial Multiplexing with 2 Tx and 2 Rx are presented. Simulation parameters are as following:

Table 6. Simulation Parameters

Carrier Frequency (GHz)	2.5
Subcarrier Spacing (kHz)	10.94
Channel Model	ITU-PB3, ITU-VA60
Channel Correlation	Low 10000 packets
FFT Size	1024

Guard Interval	1024/8=128
Permutation	PUSC
Channel Coding	Convolutional Turbo Coding (Vertical) 4 iterations
Coding Rate	id0~id7
Packet Size	60 slots (2 symbols by 30 subchannels per antenna)
HARQ	Maximum 3 retransmission (including original transmission)
Channel Estimation	ideal
MIMO Detector	Linear MMSE Detector

The transmission pattern is the same as defined in 802.16e Vertical Matrix B. A simplified table is shown below:



Ant#0 Time 0	Ant#1 Time 0
s_0	s_1
s_2	s_3
s_4	s_5
s_6	s_7
s_8	s_9
s_{10}	s_{11}
s_{12}	s_{13}
s_{14}	s_{15}

Figure 17. Original Matrix B Pattern

In each retransmission, the spectrum usage is the same as the first transmission and only half of the original transmitted pattern will be retransmitted. Thus the consecutive retransmission patterns are:

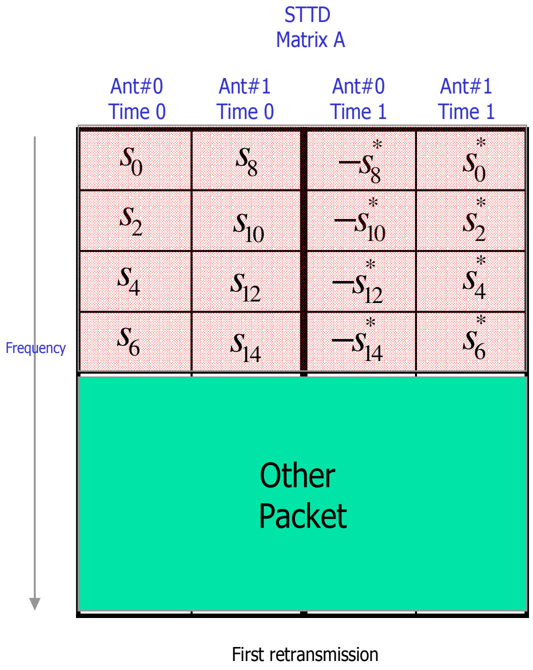


Figure 18. First Retransmission Matrix A Pattern

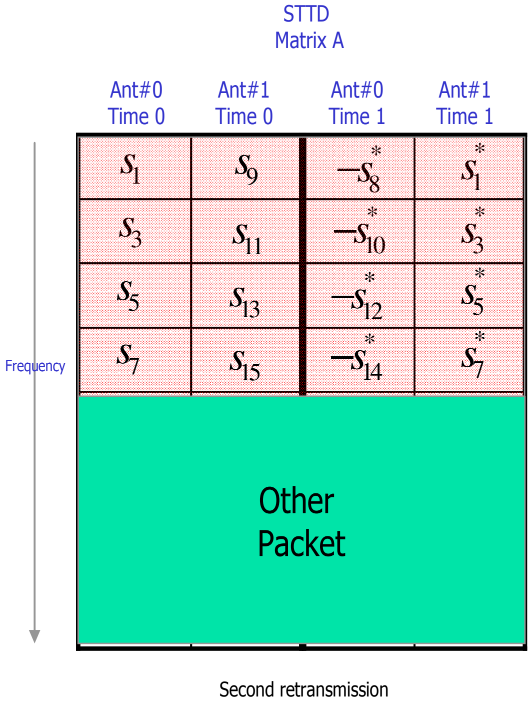


Figure 19. Second Retransmission Matrix A Pattern

The Simulation results of Throughput and Packet Error Rate are shown in the following figures.

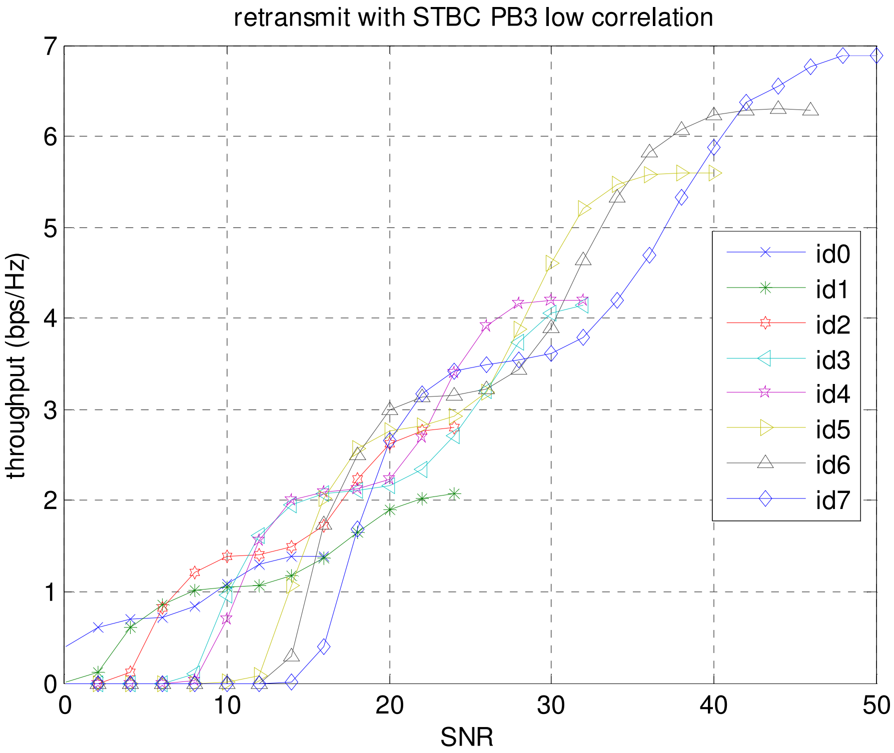


Figure 20. Throughput of retransmission with Matrix A

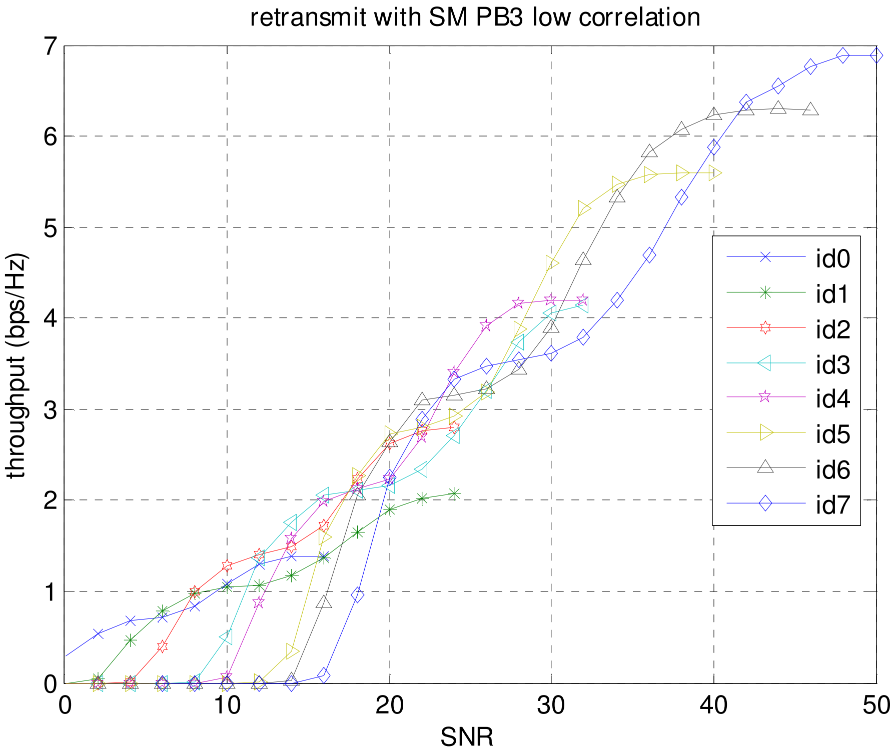


Figure 21. Throughput of retransmission with Matrix B

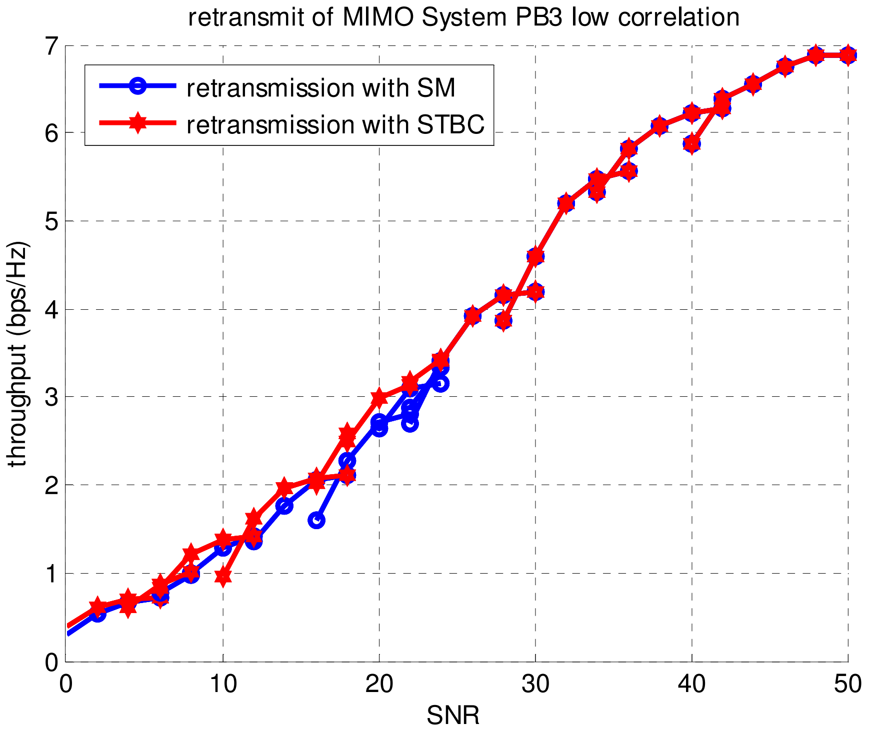


Figure 22. Throughput Comparison of Retransmission with Matrix A and B

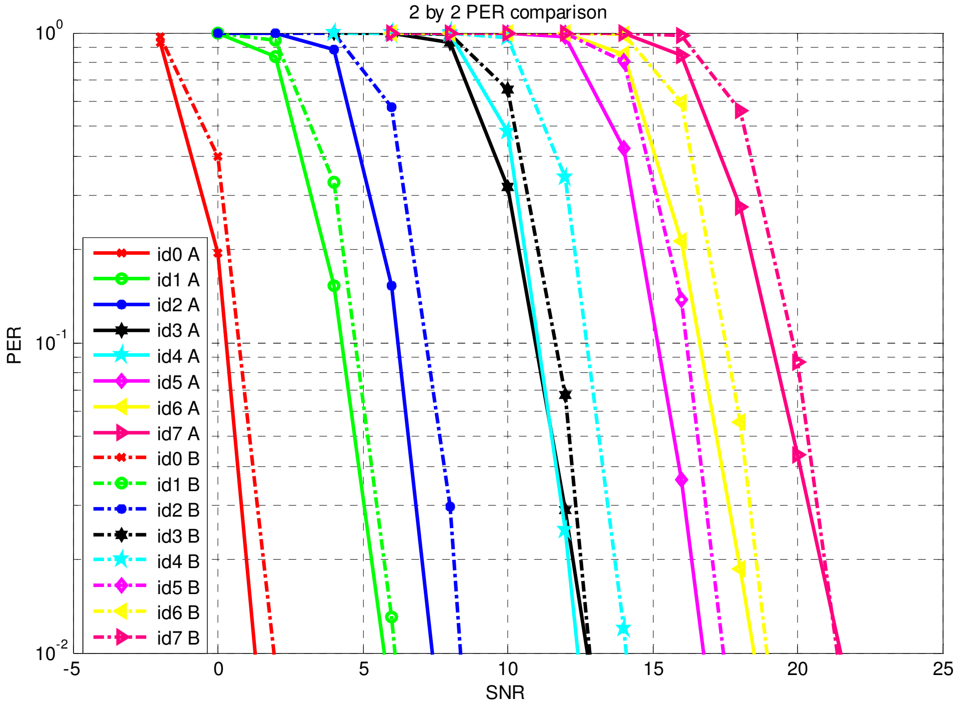


Figure 23. Packet Error Rate Comparison of Retransmission with Matrix A and B

From the result of Figure 22, it shows throughput of our proposed retransmission with Matrix A pattern always better than retransmission with Matrix B pattern no matter in what SNR regions and Coding Rate. Moreover, the implementation complexity of the proposed transceiver design has low power and low memory buffer advantages. In fading channel, BER with 1e-1 is considered. In Figure 23, the proposed retransmission with Matrix A is shown as coding rate A, and retransmission with Matrix B is shown as coding rate B. The PER of retransmission with Matrix A saves about 0.5dB in id0 and id1, saves about 1dB in id2, and saves with up to about 2 dB in id4. To summarize, our proposed retransmission with Matrix A (STBC or SFBC) can save receiver computation power, memory buffer and even gain transmission power in the same PER requirements.

8 Symbol repetition for high mobility scenario

This section presents a HARQ scheme with ICI cancellation for high-mobility users. The mobility information can be obtained via many methods, such as GPS or estimation based on RSSI or CINR. For example, the MS measures CINR by the common pilots broadcasted by the BS periodically. Then, the velocity information may be estimated roughly via the variation of CINR by the MS. If the MS is in high mobility scenario, the MS feedbacks the request of the high-mobility HARQ scheme to the BS. Figure 24 displays the flow chart of the proposed HARQ retransmission scheme. A packet is first appended with the cyclic redundancy check (CRC) code, which is used for error detection. Based on the CRC decoding, if the receiver decodes the packet correctly, then the receiver sends an ACK to the transmitter as a delivery confirmation signal indicating a correct reception, otherwise the receiver sends a NACK and request an additional retransmission to provide the receiver a successful packet reception. Conventional HARQ scheme combines the received packets by using maximal-ratio combining scheme, which achieves the maximum signal-to-noise power ratio.

The frequency domain received signal of an N-point FFT OFDM system in a time-varying, frequency-selective multipath fading channel can be expressed as

$$Y_m = \sum_{k=0}^{N-1} X_k \sum_{l=1}^{L} \frac{1}{N} \sum_{n=0}^{N-1} h_{l,n} \exp\left[\frac{j2\pi n(k-m)}{N}\right] \exp\left(\frac{-j2\pi k\tau_l}{N}\right) + \sum_{n=0}^{N-1} z_n \exp\left(\frac{-j2\pi nm}{N}\right) \text{for } m = 0, 1, 2, \cdots, N-1,$$

where X_k is the complex-valued transmitted signal for the k-th subcarrier, L is the number of multipath, $h_{l,n}$ is the complex-valued channel gain of the l-th path at n-th sample, τ_l represents the tap-delay of the l-th path, and z_n stands for an additive white Gaussian noise sample. Assume the channel variation is linear over the interval of an OFDM symbol, i.e. $h_{l,n} = \alpha_l n + \beta_l$ for $l = 1, 2, \cdots, L$, where α_l and β_l are constants. The frequency domain received signal can be rewritten as

$$Y_m = H_m X_m + \sum_{k=0, k\neq m}^{N-1} C_{k-m} H'_k X_k + Z_m,\ m = 0, 1, 2, \cdots, N-1$$

where

$$H_m = \sum_{l=1}^{L} \left(\alpha_l \frac{N-1}{2} + \beta_l\right) \exp\left(\frac{-j2\pi m\tau_l}{N}\right)$$

$$H'_k = \sum_{l=1}^{L} \alpha_l \exp\left(\frac{-j2\pi k\tau_l}{N}\right)$$

$$C_{k-m} = \frac{-1}{1-\exp\left[\frac{j2\pi(k-m)}{N}\right]}$$

$$Z_m = \sum_{n=0}^{N-1} z_n \exp\left(\frac{-j2\pi nm}{N}\right)$$

To mitigate the ICI effects, the proposed HARQ scheme permutes the data packets according the rule shown in Table 7, where $X_m = -X_{m+1}$ for $m = 0, 2, 4, \cdots, N-2$. Combining the signals, we have

$$\begin{aligned}\tilde{Y}_m &= Y_m - Y_{m+1} \\ &= (H_m + H_{m+1} - C_1 H'_{m+1} - C_{-1} H'_m) X_m + \sum_{k=0, k\neq \frac{m}{2}}^{\frac{N}{2}-1} \left[(C_{2k-m} - C_{2k-m-1}) H'_{2k} - (C_{2k-m+1} - C_{2k-m}) H'_{2k+1} \right] X_{2k} + Z_m\end{aligned},$$

for $m = 0, 2, 4, \cdots, N-2$. The ICI effects can be reduced significantly [10]. To compensate the rate loss, the proposed HARQ scheme uses two antennas. Table 8 shows the permutation rule for the proposed scheme.

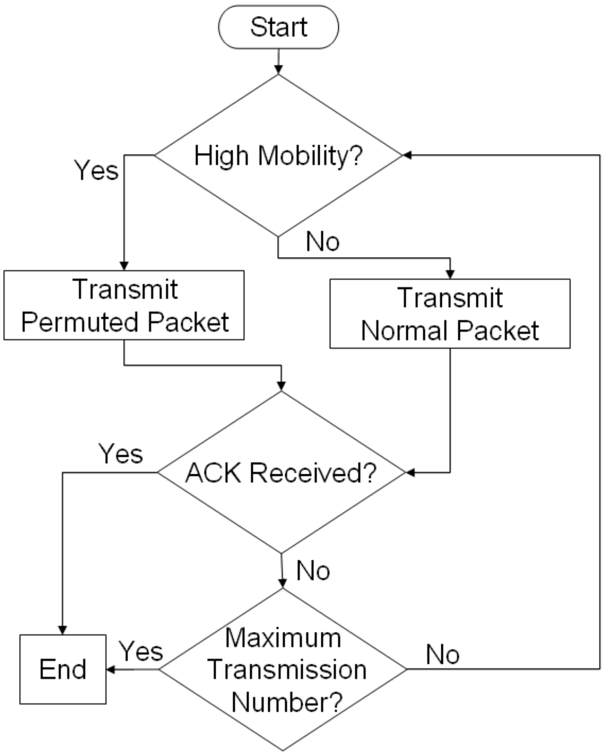


Figure 24. Flow chart of proposed HARQ scheme

Table 7. Packet Permutation Rule for the Proposed HARQ Scheme

	f_0	f_1	f_2	f_3	…	f_{N-2}	f_{N-1}
Original packet	X_0	$-X_0$	X_1	$-X_1$	…	$X_{\frac{N}{2}-1}$	$-X_{\frac{N}{2}-1}$
	$X_{\frac{N}{2}}$	$-X_{\frac{N}{2}}$	$X_{\frac{N}{2}+1}$	$-X_{\frac{N}{2}+1}$	…	X_{N-1}	$-X_{N-1}$
Retransmitted packets	X_0	$-X_0$	X_1	$-X_1$	…	$X_{\frac{N}{2}-1}$	$-X_{\frac{N}{2}-1}$
	$X_{\frac{N}{2}}$	$-X_{\frac{N}{2}}$	$X_{\frac{N}{2}+1}$	$-X_{\frac{N}{2}+1}$	…	X_{N-1}	$-X_{N-1}$

Table 8. Packet Permutation Rule for the Proposed HARQ Scheme with two Antennas

		f_0	f_1	f_2	f_3	…	f_{N-2}	f_{N-1}
Antenna 1	Original packet	X_0	$-X_0$	X_1	$-X_1$	…	$X_{\frac{N}{2}-1}$	$-X_{\frac{N}{2}-1}$
	Retransmitted packets	X_0	$-X_0$	X_1	$-X_1$	…	$X_{\frac{N}{2}-1}$	$-X_{\frac{N}{2}-1}$
Antenna 2	Original packet	$X_{\frac{N}{2}}$	$-X_{\frac{N}{2}}$	$X_{\frac{N}{2}+1}$	$-X_{\frac{N}{2}+1}$	…	X_{N-1}	$-X_{N-1}$
	Retransmitted packets	$X_{\frac{N}{2}}$	$-X_{\frac{N}{2}}$	$X_{\frac{N}{2}+1}$	$-X_{\frac{N}{2}+1}$	…	X_{N-1}	$-X_{N-1}$

Simulation results for SISO are presented. Table 9 shows the parameters used in the link-level simulations. Figures 25, 26, and 27 show the performance comparisons, where 16m stands for the proposed scheme. The proposed scheme outperforms the scheme of 16e. Although, conventionally, communication systems tend to used lower-order modulation schemes in high mobility scenarios, the proposed method can be applied to high data rate applications, such as video on demand, when mobile uses are in high mobility.

Table 9. Parameters of Link-level Simulation

Carrier frequency	2.5GHz
Operating Bandwidth	11.2MHz
FFT Size	1024
Guard Interval	1024/8=128
Resource Block Size	18 sub-carriers × 6 symbols
Channel Coding	CTCs
Packet Size	48 Resource Blocks
HARQ	Chase Combining, Maximum 2 Frames retransmission delay
Channel	ITU Veh A 350km/h
User Mobility	350km/hr

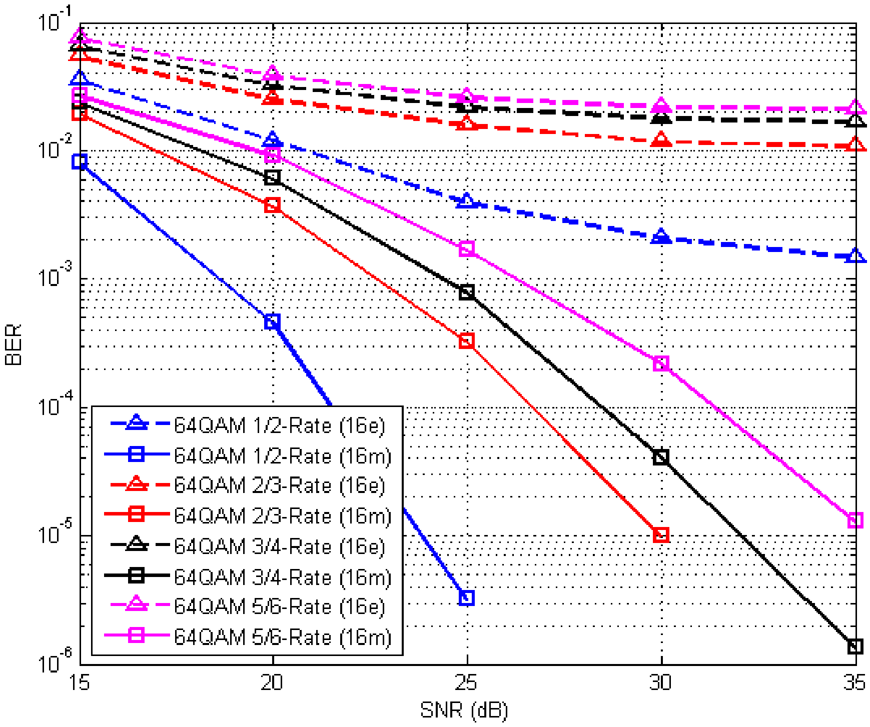


Fig. 25. BER performance comparison

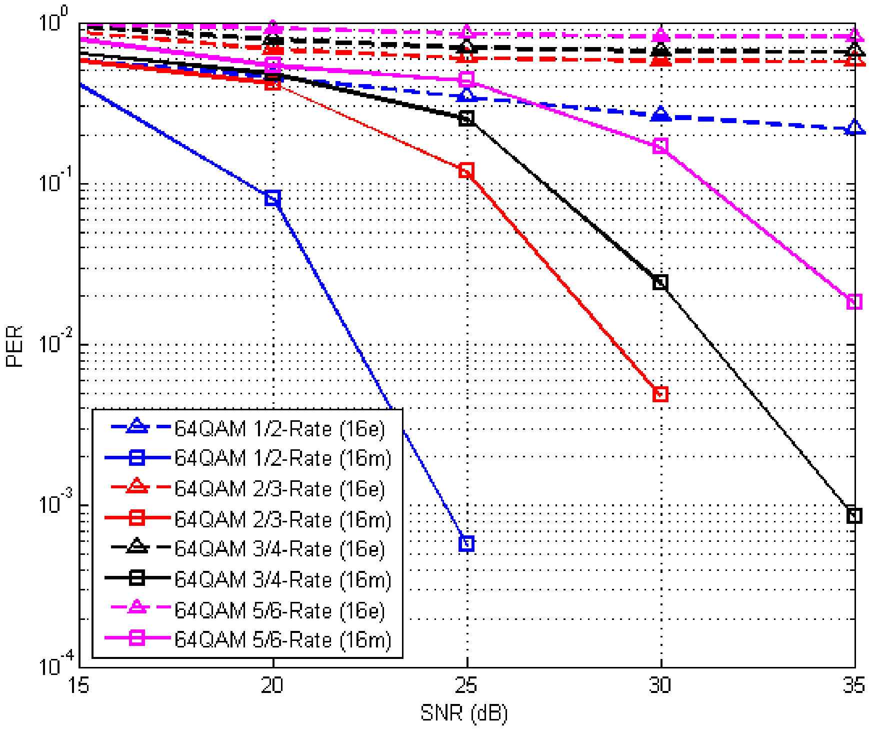


Fig. 26. PER performance comparison

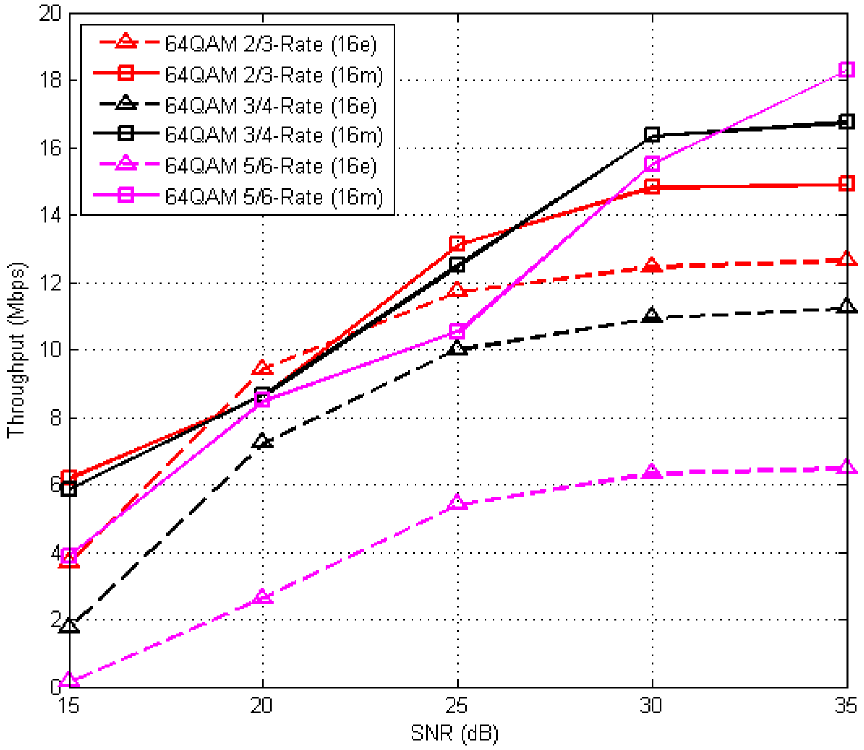


Fig. 27. Throughput performance comparison

9 Conclusions

This contribution describes HARQ functionalities and the enhancement methods for IEEE 802.16e HARQ. First, we introduce increasing the number of CTC block sizes to improve system throughput. Second, modifying CTC IR bit selection method for IEEE 802.16e renders better error rate performance. Third, bit rearrangement improves reception performance. Fourth, joint MIMO HARQ mechanism provides performance gain. Fifth, symbol repetition enhances error rate performance in high mobility.

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============================== TEXT Proposal ==============================

HARQ improves reception performance by incorporating previous incorrectly-decoded bursts and the new burst to process. Fig. X illustrates the proposed HARQ architecture. Chase combining (CC) and incremental redundancy (IR) are basic two functionalities. CC requires less storage on receiver side and IR provides better error rate performance. IR bit selection method shall select parity bits at first in the following retransmissions. CC shall incorporate with bit rearrangement to enhance error rate performance. Reliable MIMO scheme, e.g. STBC/SFBC, could be used in the retransmission for better reception performance.

In order to avoid the throughput loss due to the padding used in IEEE 802.16e, the number of CTC block sizes shall be large enough to reduce throughput loss below than [8%, TBD].

For high mobility scenario, symbol repetition is used to mitigate inter-carrier interference.

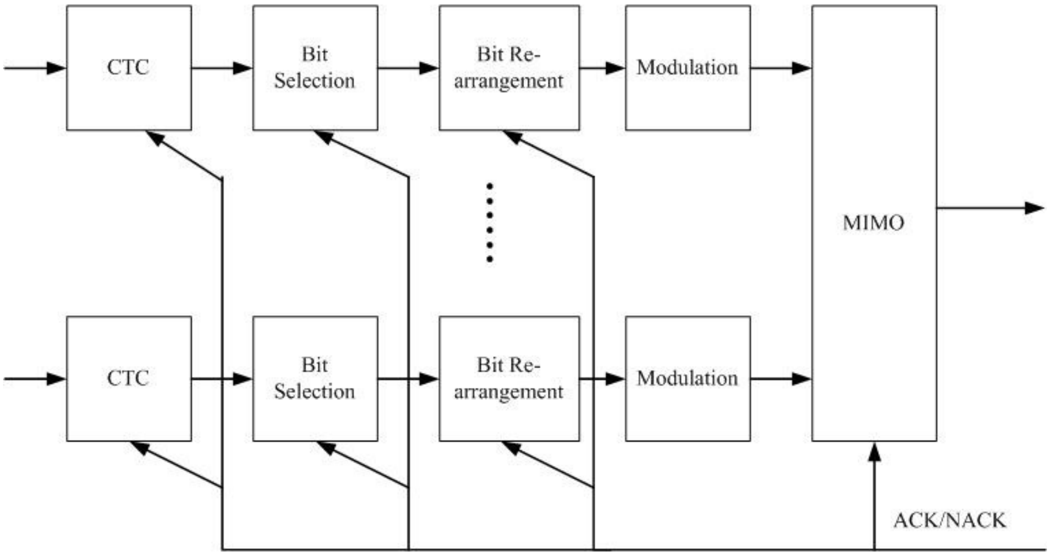

Fig.X: HARQ physical layer architecture.