An Algorithm for Iterative Detection and Decoding MIMO-OFDM HARQ with Antenna Scheduling

KyooHyun Kim, SeungWon Kang, Manar Mohaisen, and KyungHi Chang
The Graduate School of Information Technology & Telecommunications, Inha University,
253 Yonghyun-Dong, Nam-Gu, Incheon 402-751, Korea
[e-mail: makewish79@paran.com, swkang79@hanmail.net, lemanar@hotmail.com,
khchang@inha.ac.kr]
*Corresponding author: KyungHi Chang

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Abstract

In this paper, a multiple-input-multiple-output (MIMO) hybrid-automatic repeat request (HARQ) algorithm with antenna scheduling is proposed. It retransmits the packet using scheduled transmit antennas according to the state of the communication link, instead of retransmitting the packet via the same antennas. As a result, a combination of conventional HARQ systems, viz. chase combining (CC) and incremental redundancy (IR) are used to achieve better performance and lower redundancy. The proposed MIMO-OFDM HARQ system with antenna scheduling is shown to be superior to conventional MIMO HARQ systems, due to its spatial diversity gain.

Keywords: MIMO-OFDM, sphere decoding, chase combining, incremental redundancy, antenna scheduling

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1. Introduction

MIMO systems that employ several transmit and receive antennas at both ends have much higher capacity than traditional single antenna systems. However, MIMO systems incur co-antenna interference (CAI) and frequency-selective fading. CAI, which is a major disadvantage of MIMO systems, can be mitigated by employing an iterative detection and decoding (IDD) receiver. The intrinsic concept of IDD is the iterative exchange of information between the detector and the decoder, until there is no further performance improvement [1]. The robustness against frequency-selective fading can be improved by combining OFDM with MIMO techniques [2][3].

Many existing data communication systems utilize HARQ schemes, which combine forward error correction (FEC) and ARQ protocols. There are many versions of HARQ schemes. Among them, chase combining (CC) and incremental redundancy (IR) are the predominant applications. If the original transmission fails, CC re-sends an exact copy, and it is combined with the original transmission, while IR re-sends packets coded at a lower code rate, and combines these with the originally transmitted packets [4][5]. In a conventional MIMO HARQ system, if the channel status of the communication link for the antenna continues to operate in unfavorable conditions, retransmitted packets can also be erroneous, which increases the number of retransmissions and degrades system performance. Therefore, to solve these problems, a MIMO HARQ algorithm with antenna scheduling is proposed in this paper. Furthermore, to increase the data rate and improve the performance further, the IDD-based Bell laboratory layered space-time (I-BLAST) system is utilized with the proposed MIMO HARQ algorithm.

This paper is organized as follows: In Section 2, the I-BLAST system and sphere decoding (SD) algorithm are briefly described. In Section 3, details of the I-BLAST system with the proposed HARQ algorithm with antenna scheduling is described. In Section 4, the performance of the I-BLAST system and the proposed HARQ algorithm are evaluated by computer simulation. Finally, in Section 5, conclusions are provided.

2. I-BLAST System

A major factor degrading the performance of the BLAST system is the aforementioned CAI, which is caused by transmitting the signal using n_T antennas. However, this disadvantage of MIMO can be overcome via an IDD receiver, which is known as I-BLAST system.

2.1 Transmitter of I-BLAST System

Fig. 1 shows the transmitter architecture of the I-BLAST system with n_T transmitting antennas, including a conventional V-BLAST (Vertical BLAST) system with a diagonal space (DS) interleaver. The input bit sequence **b** is de-multiplexed and encoded by an error-correcting code, to generate the coded bit sequence **c**, which is further interleaved by a diagonal space bit interleaver, to generate an interleaved coded bit sequence $\mathcal{C}$.

Fig. 2 shows the diagonal space interleaver based on diagonal layering of each independently coded sub-stream. Note, unlike D-BLAST (Diagonal BLAST), no boundary wastage occurs [5]. The interleaved sequence $\mathcal{C}$ has q coded bits in each transmitting antenna branch, which are mapped by M-QAM, thus, **a** has a total of $q \cdot n_T$ coded bits, where $q = \log_2 M$.

An n_T – dimensional signal vector $\mathbf{a} = [a_1, a_2, \ldots, a_{n_T}]^T$ is forwarded to the OFDM modulator, which performs a virtual carrier insertion, an inverse fast Fourier transform (IFFT), then, a guard interval (GI) insertion. Finally, it is transmitted via n_T antennas.

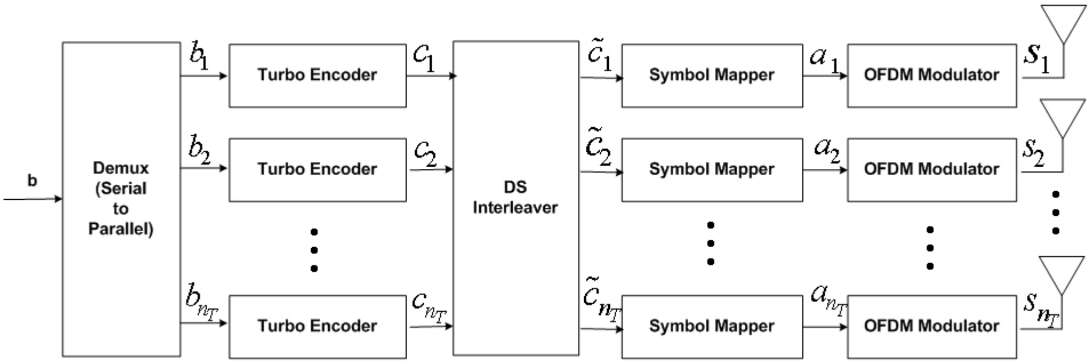
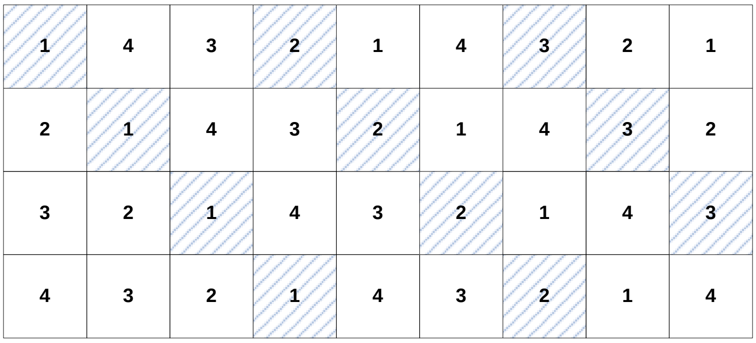


Fig. 1. Transmitter architecture of I-BLAST system.



1	4	3	2	1	4	3	2	1
2	1	4	3	2	1	4	3	2
3	2	1	4	3	2	1	4	3
4	3	2	1	4	3	2	1	4

Fig. 2. Diagonal space interleaver.

2.2 Receiver of I-BLAST System

Fig. 3 shows the architecture of the iterative receiver used in I-BLAST. The main components of the receiver are an inner soft-input soft-output (SISO) decoder, a DS (de-) interleaver and a n_R parallel outer SISO decoder.

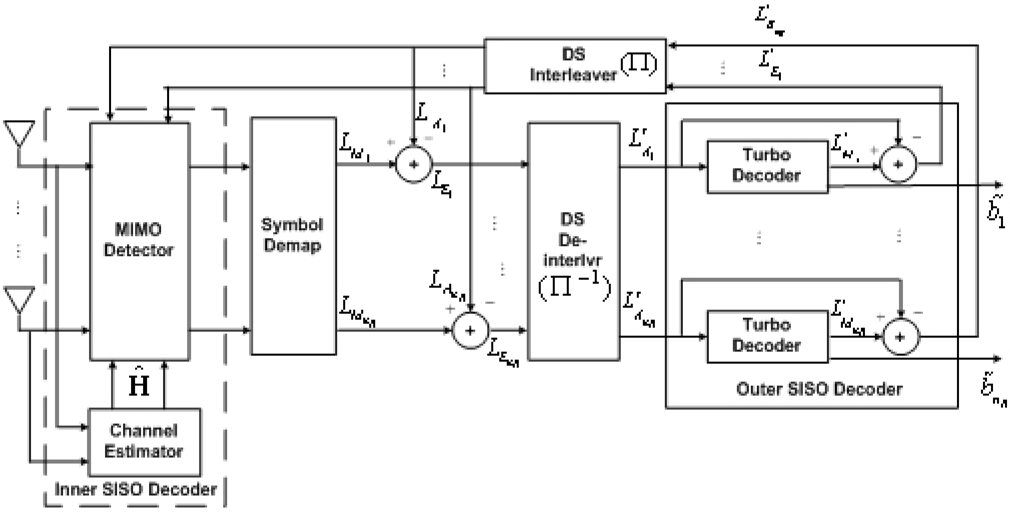


Fig. 3. Receiver architecture of I-BLAST system

Assuming an $n_T \times 1$ OFDM symbol vector **s** is transmitted over an $n_T \times n_R$ MIMO channel **H**, the $n_R \times 1$ received signal vector **r** is expressed by Eq. (1).

$$\mathbf{r}^{(k)} = \mathbf{H}\mathbf{s}^{(k)} + \mathbf{n}^{(k)} \tag{1}$$

In Eq. (1), k is a sub-carrier index and **n** is $n_R \times 1$ additive white Gaussian noise with a mean of zero and a variance of $\sigma^2 \mathbf{I}_{n_R}$.

The LLR (Log Likelihood Ratio) of the interleaved bit sequence $\tilde{\mathbf{c}} = [\tilde{c}_1,\ \tilde{c}_2,\ \ldots,\ \tilde{c}_{q \cdot n_T}]^T$ is defined by Eq. (2) [1][6].

$$L(\tilde{c}_i) = \log \frac{\Pr[\tilde{c}_i = 1 \mid \mathbf{r}]}{\Pr[\tilde{c}_i = 0 \mid \mathbf{r}]}, \qquad i = 1, 2, \ldots, q \cdot n_T$$

using Bayes's rule,

$$\begin{aligned} &= \log \frac{\Pr[\mathbf{r} \mid \tilde{c}_i = 1] \Pr[\tilde{c}_i = 1]}{\Pr[\mathbf{r} \mid \tilde{c}_i = 0] \Pr[\tilde{c}_i = 0]} \\ &= \log \frac{\sum_{\mathbf{s} \in S_i^1} \exp(-\frac{1}{2\sigma^2} \left\| \mathbf{r} - \mathbf{H}\mathbf{s} \right\|^2) \Pr[\mathbf{s}]}{\sum_{\mathbf{s} \in S_i^0} \exp(-\frac{1}{2\sigma^2} \left\| \mathbf{r} - \mathbf{H}\mathbf{s} \right\|^2) \Pr[\mathbf{s}]} \end{aligned} \tag{2}$$

In Eq. (2), $\Pr[\tilde{c}_i = b]$, where b has a value of 0 or 1, is the intrinsic information about a coded bit, $\tilde{c}_i$, and S_i^c is expressed as Eq. (3).

$$S_i^c = \{\mu(\tilde{\mathbf{c}}) \mid \tilde{\mathbf{c}} = [\tilde{c}_1,\ \tilde{c}_2,\ \ldots,\ \tilde{c}_{q \cdot n_T}],\ \tilde{c} = c\} \tag{3}$$

Here, $\mu(\cdot)$ denotes a modulation function. During the first iteration, the initial intrinsic probabilities of all symbol bits are assumed to be 0.5 (i.e., equal probabilities of 0 and 1). Using the approximation of $\sum P_i \approx \max\{P_i\}$, Eq. (2) is expressed as Eq. (4) [6].

$$\begin{aligned} L(\tilde{c}_i) \cong & \frac{1}{2\sigma^2} [\min_{\mathbf{s} \in S_i^1} \{ \left\| \mathbf{r} - \mathbf{H}\mathbf{s} \right\|^2 - \log \Pr[\mathbf{s}] \}] \\ & - \frac{1}{2\sigma^2} [\min_{\mathbf{s} \in S_i^0} \{ \left\| \mathbf{r} - \mathbf{H}\mathbf{s} \right\|^2 - \log \Pr[\mathbf{s}] \}] \end{aligned} \tag{4}$$

The a-priori LLR value $L_E(\tilde{c}_i)$ obtained by Eq. (4) is de-interleaved and used for the outer SISO decoder. The value of $\tilde{c}_i$ in Eq. (4) is c_i after the DS de-interleaver (Π^{-1}), and is $\tilde{c}_i$

again after the DS interleaver (Π) in the feedback path. The relationship between LLRs $L'(c)$ and $L(\tilde{c})$ is provided by Eq. (5).

$$\begin{aligned} L'(c) &= \Pi^{-1}\{L(\tilde{c})\} \\ L(\tilde{c}) &= \Pi\{L'(c)\} \end{aligned} \tag{5}$$

Based on the a-priori information, $L'_A(c_i)$, the outer SISO decoder yields a-posteriori information, $L'_M(c_i)$, as expressed by Eq. (6).

$$L'_M(c_i) = L'_A(c_i) + L'_E(c_i) \tag{6}$$

In this equation, $L'_E(c_i)$ is termed extrinsic information and is obtained via decoding. This value is interleaved and calculated for Eq. (7).

$$\Pr[\mathbf{s}] = \prod_{i=1}^{q \cdot n_T} \frac{[\exp\{L_E(\tilde{c}_i)\}]^{\tilde{c}_i}}{1+\exp[L_E(\tilde{c}_i)]} \tag{7}$$

Eq. (7) is used as a-priori probability $L'_A(c_i)$ in Eq. (6) in the next iteration. As the number of iterations increases, the performance of the I-BLAST system is improved, due to the more accurate a priori probability in Eq. (7) [6].

2.3 MIMO Detection Algorithm

Maximum likelihood (ML) achieves the best performance among all detection algorithms. The concept underpinning ML detection involves finding a solution s which minimizes the following function:

$$\hat{\mathbf{s}}_{ML} = \arg\min_{\mathbf{s}\in\mathbf{\Omega}^{n_T}} \left\| \mathrm{r} - \mathbf{Hs} \right\|^2 . \tag{8}$$

ML is statistically optimal. On the other hand, it needs an exhaustive search of all candidate solutions, which is impractical.

Sphere decoding (SD) was the first algorithm introduced and adopted as a MIMO detection algorithm for reducing ML detection complexity and achieving near-ML performance [7][8]. Though the complexity of SD is a random variable, it can efficiently reduce the ML complexity while achieving the required performance. In SD, as in every ML complexity-reduction algorithm, we are aiming to eliminate branches (burn nodes) which have accumulative square distances exceeding a pre-defined maximum search square radius. To define the accumulative square distance (square accumulative metric) based on Eq. (8), the channel matrix **H** is decomposed into the product of unitary and upper triangular matrices, respectively, via **QR**-decomposition [9]. Furthermore, a search accumulative metric (a search sphere radius) is set, such that any solution (in any detection stage) which exceeds its bounds is discarded. Based on these adaptations, Eq. (8) can be expressed as follows:

$$\begin{aligned}\hat{\mathbf{s}}_{ML} &= \arg\min_{\mathbf{s}\in\mathbf{\Omega}^{n_T}} \left\{ \sum_{j=1}^{n_T} \left| z_j - \sum_{i=j}^{n_T} R_{j,i}\, s_i \right|^2 \right\} \\ &= \arg\min_{\mathbf{s}\in\mathbf{\Omega}^{n_T}} \left\{ \left\| \mathbf{R}\left(\hat{\mathbf{s}} - \mathbf{s}\right) \right\|^2 \right\} \\ &= \arg\min_{\mathbf{s}\in\mathbf{\Omega}^{n_T}} \left\{ \sum_{j=1}^{n_T} \left| R_{j,j}\left(\hat{s}_j - s_j\right) + \sum_{i=j+1}^{n_T} R_{j,i}\left(\hat{s}_i - s_i\right) \right|^2 \right\} \end{aligned} \quad . \tag{9}$$

Finally, SD finds the best solution satisfying Eq. (10).

$$\sum_{j=1}^{n_T} \left| R_{j,j}\left(\hat{s}_j - s_j\right) + \sum_{i=j+1}^{n_T} R_{j,i}\left(\hat{s}_i - s_i\right) \right|^2 \le d^2 \tag{10}$$

Where $R_{i,j}$ is the intersection element of the i-th row with the j-th column of **R**. $\mathbf{z} = \mathbf{Q}^H\mathbf{r}$, and $\hat{s}_i$ is the estimate of the i-th transmitted signal.

The selected search radius plays a very important role in determining the complexity/performance tradeoff, such that as the radius increases, there are a higher number of solutions, thus, the complexity increases. On the other hand, a small radius can result in an empty set, such that the radius has to be tuned (to a larger value) and detection is reinitiated. In this paper, we set the square radius to the Babai point square distance, which can be defined as the left-hand side of Eq. (10), where signals are detected independently.

3. MIMO HARQ Algorithm with Antenna Scheduling

Conventional MIMO HARQ algorithms retransmit packets using the same antennas, if the transceiver receives a negative acknowledgement (NAK) in its feedback link. If the channel status of the communication link for the antenna remains in an adverse condition, the retransmitted packet may continue to be erroneous, and the number of retransmissions continuously increases, leading to further degradation of system performance.

In this section, a MIMO HARQ algorithm with antenna scheduling is proposed. The proposed algorithm retransmits a packet using scheduled antennas according to the state of the communication link, instead of retransmitting the packet via the same antennas. A stop-and-wait (SAW) ARQ scheme is assumed, which transmits the next packet after receiving an ACK or NAK.

3.1 System Model

Fig. 4 shows the I-BLAST transmitter architecture with the proposed MIMO HARQ algorithm with antenna scheduling. The main components of the transmitter are a cyclic redundancy check (CRC) encoder, conventional transmit I-BLAST blocks, a transmit antenna selection block, and a resource scheduler. The CRC encoder attaches the CRC to help the receiver to determine whether there is an error in the received packet. The Tx antenna selection block selects L_T transmitting antenna branches, with high channel-sum values according to Eq. (12), among all n_T antennas. Utilizing feedback information, the resource scheduler determines the code rate of the turbo encoder, as well as the antennas that transmit the signal. In **Fig. 4**, $\{\Im_{T_n}, ACK\}$, $j = 1, 2, \ldots, n_T$ are the channel sums and ACK values, respectively.

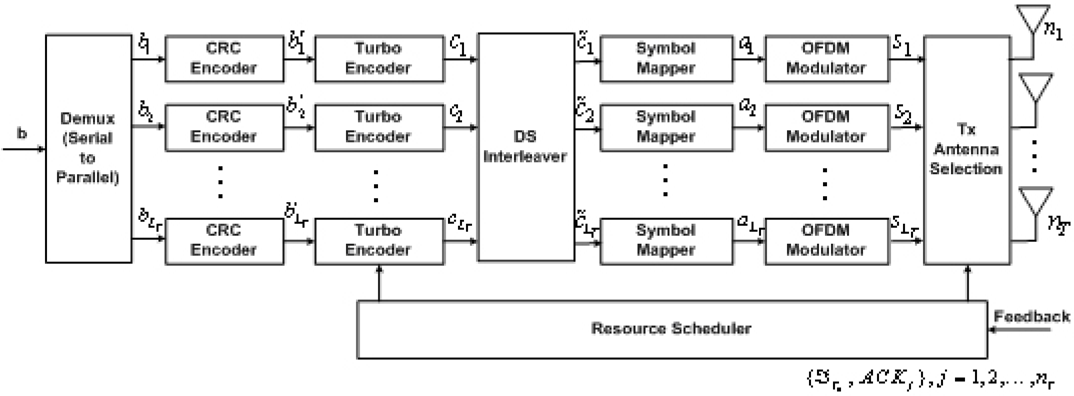


Fig. 4. Transmitter architecture of I-BLAST with proposed HARQ algorithm with antenna scheduling.

Fig. 5(b) shows the receiver architecture of I-BLAST with the proposed MIMO HARQ algorithm adapted to antenna scheduling. The main components of the receiver are the conventional receive blocks of I-BLAST, the receive antenna selection block, and the CRC decoder. Estimated channel values are used to calculate the channel sum and select $L_R \times L_T$ antennas.

A joint transmit / receive antenna selection algorithm, depicted in **Fig. 5(a)**, is used to select $L_R \times L_T$ antennas among $n_R \times n_T$ antennas. The CRC decoder extracts the CRC sequence from the turbo-decoded bit sequence and determines whether there is an error.

3.2 Proposed MIMO HARQ Algorithm for Antenna Scheduling

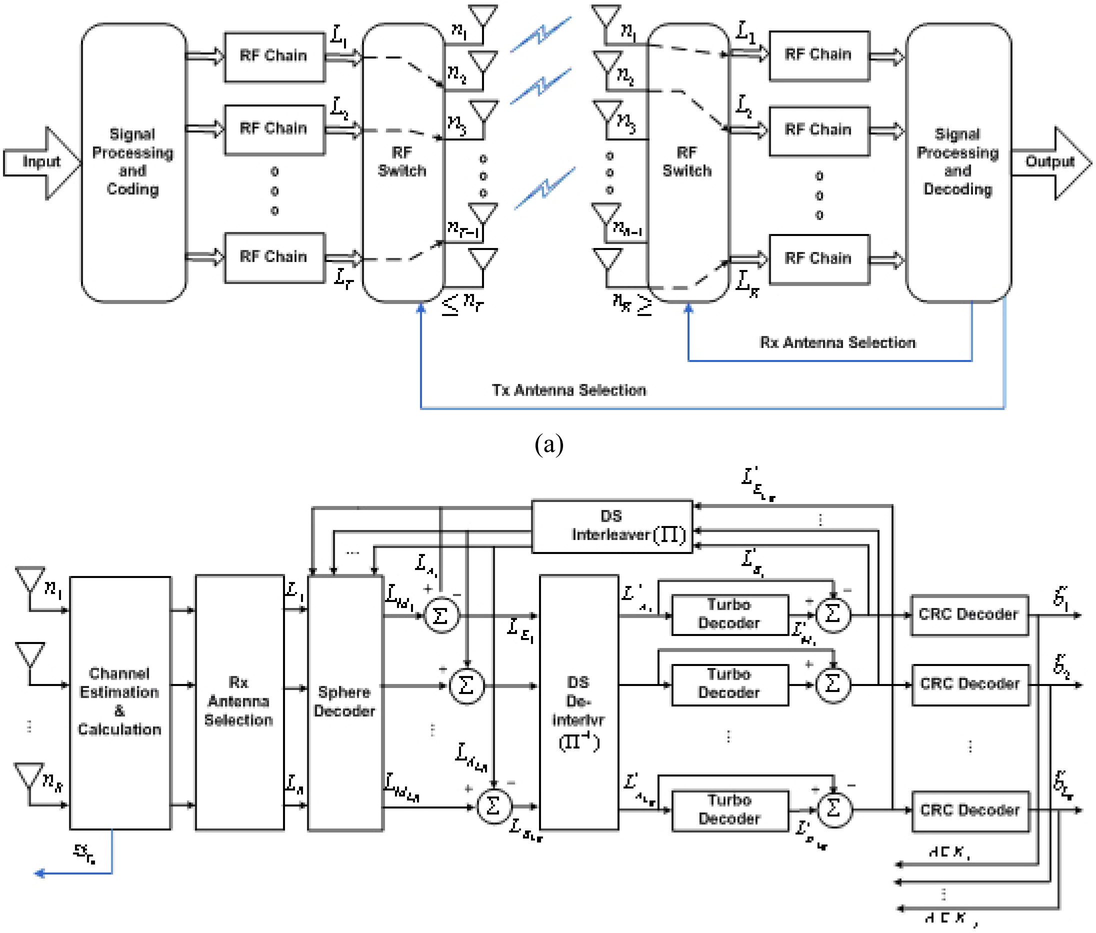
Fig. 6 shows a flow chart of the proposed MIMO HARQ algorithm, for the downlink. The transmitter receives the feedback information of the channel sum values and ACK values, according to Eq. (11). Then, it determines the retransmission strategy, as follows:

$$\{\Im_{T_n}, ACK_j\}, \quad j = 1, 2, \ldots, n_T \tag{11}$$

Initially, among n_T antennas L_T Tx antennas are selected according to the order of the channel sum values $\Im_{T_n}$ in Eq. (11). Also selected are the antennas for retransmission according to the order of the channel sum values.

The scheduler at the transmitter side then checks the ACK value of each antenna.

- If all Tx antennas receive a value of 1 (NAK), then an incremental redundancy scheme is employed for robust packet retransmission over the communication link. In this case, all L_T antennas are utilized for the retransmission.
- If the number of ACKs is greater than or equal to the number of NAKs, the scheduler retransmits via chase combining using the same number of ACK antennas as the number of NAKs.
- If the number of NAKs is greater than the number of ACKs, the scheduler retransmits via chase combining using the ACK antennas, and incremental redundancy using the number NAK antennas given by: (number of NAK antennas – number of ACK antennas).



(a)

(b)

Fig. 5. (a) Joint Tx/Rx antenna selection block. (b) Receiver architecture of I-BLAST with proposed HARQ algorithm with antenna scheduling.

On the receiver side, channel sum values are obtained by Eq. (12).

$$
\begin{aligned}
\Im_{T_n}^{Sum} &= \sum_{m=1}^{n_R} \left| H_{mn} \right|, \; n = 1,\ 2,\ \ldots,\ n_T \\
\Im_{R_m}^{Sum} &= \sum_{n=1}^{n_T} \left| H_{mn} \right|, \; m = 1,\ 2,\ \ldots,\ n_R
\end{aligned}
\tag{12}
$$

After calculating the channel sum values, the Rx antenna selector selects L_R Rx antennas among n_R antennas, which are selected according to the order of the channel sum values $\Im_{R_m}$ in Eq. (12). Then, the receiver determines whether the received packet is a retransmitted packet or not.

For receiving a retransmitted packet, the receiver combines the retransmitted packet with the originally received packet in the buffer, then, CRC decoding is performed. Determination of whether there is an error using Eq. (13) and (14) [10].

(a) (b)

Fig. 6. (a) Proposed MIMO HARQ algorithm at downlink transmitter side. (b) Proposed MIMO HARQ algorithm at downlink receiver side.

$$Error\ Check : \left[\frac{P(x)}{G(x)} = Q(x) + R(x) \right] \tag{13}$$

$$ACK = \begin{cases} 0: \text{ No Error} & \text{if } R(x) = 0 \\ 1: \text{ Error} & \text{if } R(x) \neq 0 \end{cases} \tag{14}$$

In Eq. (13), $P(x)$, $G(x)$, $Q(x)$, and $R(x)$ are the information polynomial, CRC polynomial, quota, and remainder, respectively. If the remainder $R(x)$ is zero, there are no transmission errors.

4. Simulation Results

In this section, the performance of I-BLAST is evaluated via computer simulation, and an optimal iteration number for IDD is determined. The performances of MIMO detection algorithms are compared, and the performance of a conventional MIMO-HARQ algorithm is also compared to the proposed antenna scheduling-adapted MIMO-HARQ algorithm, to show the gain due to spatial diversity.

The MIMO channel model used for simulation is the spatial channel model extension (SCM-E) of the 3rd generation partnership project ad-hoc group (3GPP AHG) [11].The simulation parameters are summarized in **Table 1**. **Fig. 7** shows the structure of sub-frame construction and pilot assignment per antenna, which is employed in 3GPP LTE (Long Term Evolution) standardization [12].

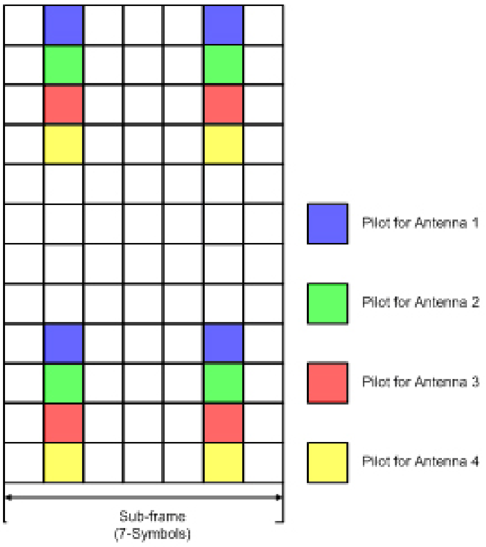


Fig. 7. Structure of the pilot assignment per antennas

Table 1. Simulation parameters

Parameter	Value
Carrier Frequency (GHz)	2
Bandwidth of Operation (MHz)	20
Number of FFT Points	2,048
Cyclic Prefix	146
Modulation	QPSK, 16QAM
Sub-frame Duration (ms)	0.5
OFDM Symbols per Sub-frame	7
Mobile Speed (km/h)	120
MIMO Fading Channel Model	SCM-E Sub-urban Macro
Channel Estimation	Practical
MIMO Detection	ZF, MMSE, SIC, SD
Tx / Rx Antenna Configuration (Selected Antenna Configuration)	6 x 6 (4 x 4)
Tx / Rx Antenna Distance	10λ / 0.5λ
Channel Coding	Turbo Coding
Mother Code Rate	2/3
HARQ Algorithm	CC, IR, Proposed

Retransmission Code Rate of IR	3/5, 8/15, 1/2, 2/5
Max Retransmission Number	4

4.1 Performance of I-BLAST System

Fig. 8 and **9** show the 4 x 4 QPSK uncoded BER and the 4 x 4 16QAM coded BER performance of the I-BLAST system according to the iteration number.

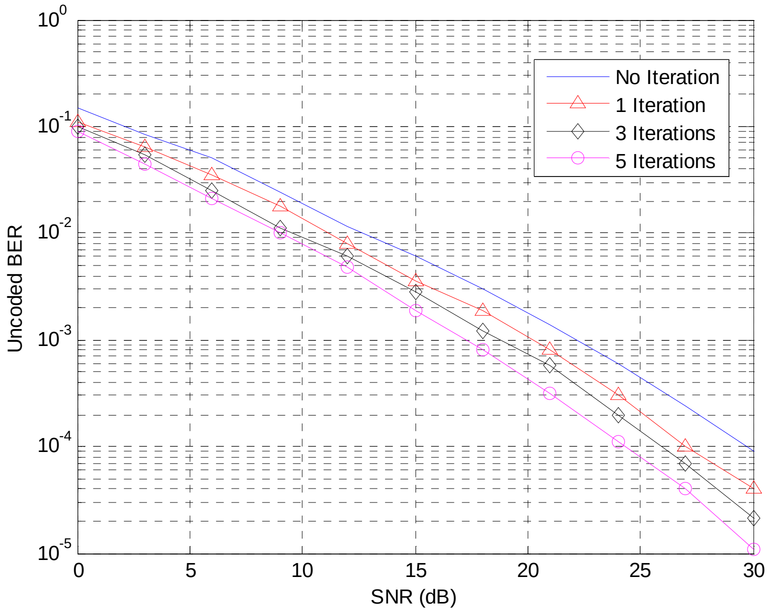


Fig. 8. 4x4 QPSK uncoded BER performance of I-BLAST system according to iteration number

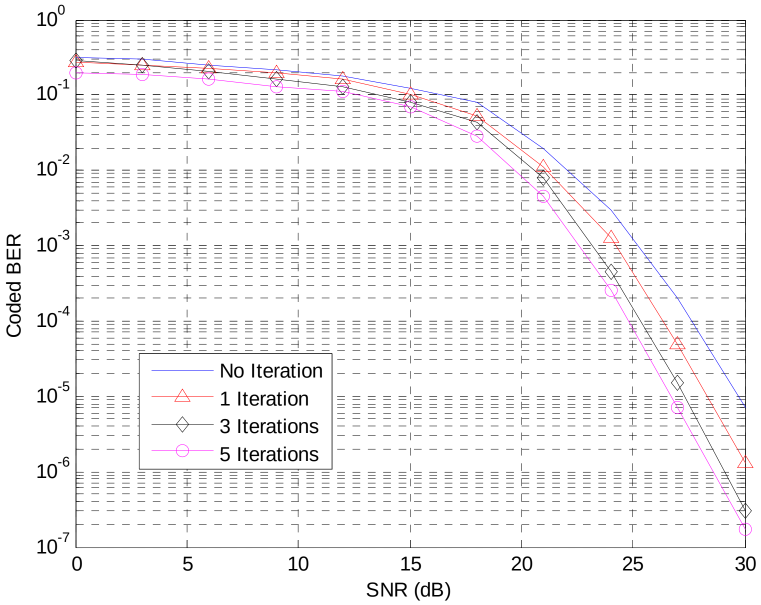


Fig. 9. 4x4 16QAM coded BER performance of I-BLAST system according to iteration number

In the case of the uncoded QPSK in **Fig. 8**, for a target BER of 10^{-4}, SNR gains for three and five iterations are about 4.8 dB and 5.2 dB, respectively, compared to the case with no iteration. In the case of the coded 16QAM shown in **Fig. 9**, for a target BER of 10^{-5}, SNR gains for three and five iterations are about 3 dB and 2.7 dB, respectively, compared to the case with no iteration. Based on these results, it is clear that the detection error, channel estimation error and interference components are largely eliminated for three iterations. Thus, the optimal number of iterations is three.

4.2 Performance of MIMO Detectors

The 4 x 4 16QAM coded BER performance of the I-BLAST system was tested for MIMO detections with zero forcing (ZF), minimum mean-square error (MMSE), ZF successive interference cancellation (ZF-SIC), MMSE-SIC, and the SD as shown in **Fig. 10**. It was found that SD shows the best performance; ZF is the simplest, but had the worst performance, as was expected. For a target BER of 10^{-5}, SD gains more than 10 dB of SNR, compared to ZF detection. The ZF-SIC and MMSE-SIC schemes are superior to the ZF and MMSE schemes, due to the interference cancellation strategy.

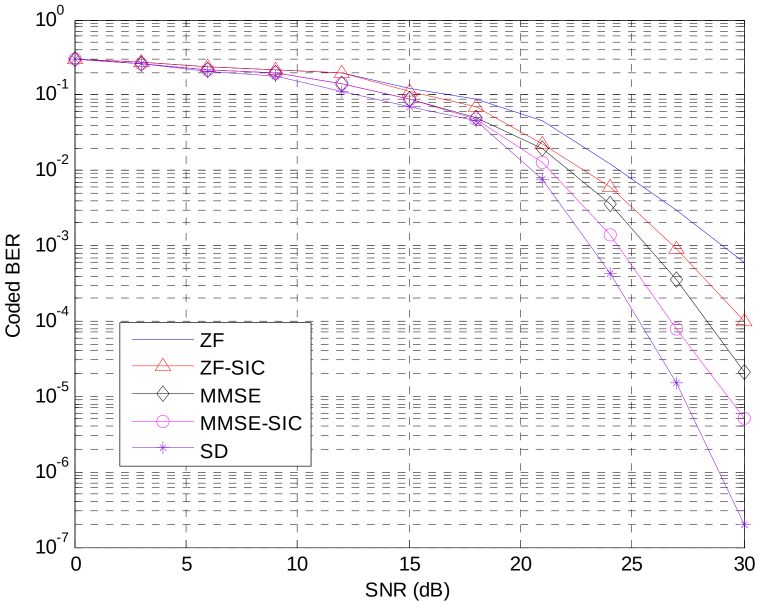


Fig. 10. 4x4 16QAM coded BER performance of I-BLAST according to MIMO detection method

4.3 Performance of Proposed MIMO HARQ Algorithm with Antenna Scheduling

In this subsection, the performance of the transmit/receive antenna selection scheme and the proposed HARQ scheme with antenna scheduling is discussed.

Antenna selection as described in this subsection is based on the channel sum in Eq. (12).

Fig. 11 and **12** show the performance of the I-BLAST system with the proposed HARQ algorithm with antenna scheduling. In this simulation, the stop-and-wait (SAW) ARQ scheme is used. The SAW ARQ scheme transmits a packet after receiving an ACK or NAK from the receiver, and the new packet is not transmitted during retransmission. In **Fig. 11**, for a target BER of 10^{-5}, the proposed algorithm is superior to the antenna scheduling schemes using conventional chase combining and incremental redundancy; the gain in SNR is 3 dB and 0.5 dB, respectively. In **Fig. 12**, for a target PER of 10^{-3}, the gain for the two aforementioned schemes using the proposed algorithm is 2 dB and 0.7 dB, respectively. The gain for the proposed HARQ algorithm results from spatial diversity, due to the proposed antenna scheduling scheme.

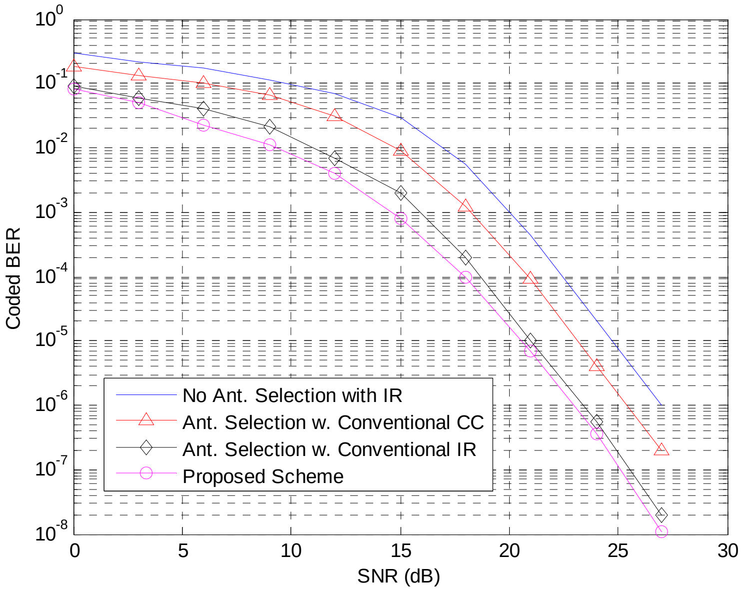


Fig. 11. 4x4 16QAM coded BER performance of I-BLAST with HARQ and antenna scheduling

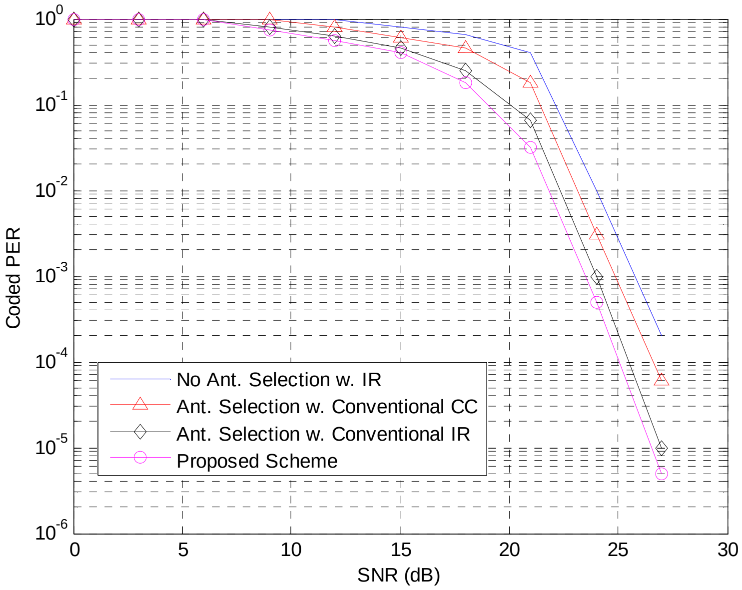


Fig. 12. 4x4 16QAM coded PER performance of I-BLAST with HARQ and antenna scheduling

5. Conclusion

In this paper, a MIMO HARQ algorithm with antenna scheduling was proposed. It retransmits a packet using scheduled antennas according to the state of the communication link, instead of retransmitting the packet using the same antennas. In the MIMO HARQ system with I-BLAST, the optimal number of iterations was determined, and the performances of detection schemes such as sphere decoding were compared. In the proposed antenna scheduling scheme, a channel sum type of antenna selection scheme was adopted, which considers the tradeoff between performance and complexity. Finally, the proposed MIMO HARQ system based on antenna scheduling was shown to be superior to conventional MIMO HARQ systems, due to its spatial diversity gain.

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KyooHyun Kim received the BS degree in electronic engineering from Inha University, Incheon, Korea, in 2005. He received the MS degree in the Graduate School of Information Technology and Telecommunications, Inha University, Incheon, Korea, in 2007. He is currently with Mobile Communications (MC), LG Electronics Co., Ltd. His research interests include 4G wireless communication systems, iterative decoding and detection, and detection schemes for spatial multiplexing MIMO systems.

SeungWon Kang received the BS degree in electronics engineering from Inha University, Incheon, Korea, in 2005. He received the MS degree in the Graduate School of Information Technology and Telecommunications, Inha University, Incheon, Korea, in 2007. He is currently with the Telecommunication Network (TN), Samsung Electronics Co., Ltd. His research interests include 4G wireless communication systems, MIMO techniques, turbo equalizer, WiBro, and wavelet-based OFDM systems.

Manar Mohaisen received the B.S. in electrical engineering from the University of Gaza, Gaza, Palestine, in 2001. From 2001 to 2003, he was with the Palestinian Telecommunications Company, where he worked as cell-planning engineer. He received the M.S. degree in communication and signal processing from the University of Nice-Sophia Antiplois, Sophia Antipolis, France, in 2005. Currently he is pursuing the Ph.D. degree in the Graduate School of Information Technology and Telecommunications at Inha University, Incheon, Korea. His research interests include 3GPP LTE systems, interference cancellation in OFDM system, and detection schemes for spatial multiplexing MIMO systems.

KyungHi Chang received the BS and MS degrees in electronics engineering from Yonsei University, Seoul, Korea, in 1985 and 1987, respectively. He received his Ph.D. degree in electrical engineering from Texas A&M University, College Station, Texas, in 1992. From 1989 to 1990, he was with the Samsung Advanced Institute of Technology (SAIT) as a Member of Research Staff and was involved in digital signal processing system design. From 1992 to 2003, he was with the Electronics and Telecommunications Research Institute (ETRI) as a Principal Member of Technical Staff. During this period, he led the design teams of WCDMA UE modem and 4G radio transmission technology (RTT). He is currently with The Graduate School of Information Technology and Telecommunications, Inha University, where he has been an Associate Professor since 2003. His current research interests include RTT design for IMT-Advanced & 3GPP LTE systems, WMAN system design, cognitive radio, cross-layer design, cooperative relaying system, and RFID/USN. Dr. Chang has served as a Senior Member of IEEE since 1998. Currently he is an Editor in Chief for the Korean Institute of Communication Sciences (KICS) Proceedings. He has also served as an Editor of ITU-R TG8/1 IMT.MOD, and he is currently an International IT Standardization Expert of Telecommunications Technology Association (TTA).