Space-Time Coding과 낮은 복잡도의 복호 방법을 사용한 효과적인 Hybrid ARQ 기법

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Efficient Hybrid ARQ with Space-Time Coding and Low-Complexity Decoding

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요 약

본 논문에서는 Space-Time code를 사용하는 다중 안테나 시스템에서 Hybrid automatic retransmission request (HARQ)의 처리능력(throughput)을 향상시키는 기법을 제안한다. 우선 낮은 복잡도를 가지는 Hard decision decoding (HDD) 방법에서 신뢰도 정보를 이용하여 복호 성능을 높일 수 있는 알고리즘을 제안한다. 이렇게 제안된 알고리즘을 HARQ 프로토콜을 사용하는 시스템에 사용하여 낮은 복잡도를 유지하면서 전체 처리능력을 향상시킬 수 있도록 한다. 제안된 기법의 성능을 확인하기 위하여 처리능력을 수학적으로 분석하였으며, 모의실험을 통해 AWGN 채널 및 페이딩이 있는 다중입력 다중출력 채널뿐만 아니라 Impulse 잡음이 있는 환경에서도 성능이 향상됨을 확인하였다.

Key Words : Space-Time coding, Hybrid ARQ, HDD

ABSTRACT

We aim at increasing the throughput of the hybrid automatic retransmission request (HARQ) protocol in Space-Time (ST) coded multi-antenna transmission systems. By utilizing reliability information at the decoder, we obtain an improved probability of successful decoding, which enhances the overall system throughput at low-complexity. Simulations and analytical results demonstrate the performance of our scheme in impulse noise environment as well as AWGN and fading multi-input multi-ouput (MIMO) channels.

I. Introduction

Demand in wireless connectivity has led to major advances providing reliable high data rates over wireless channels. Diversity is typically employed to mitigate channel fading effects in time, frequency, or space [1],[7],[8],[10]. Recently, full-diversity and full-rate (FDFR) Space-Time (ST) codes with any number of transmit- and receive-antennas in MIMO fading channels have been proposed to achieve high performance and high data rates [7]. Although ST coding has well appreciated merits in coping with channel effects, hybrid automatic retransmission request (HARQ) is typically employed to improve the low throughput of ARQ-only protocols [8].

※ 본 연구는 정보통신부 대학 IT 연구센터 육성.지원사업의 연구결과로 수행되었습니다.
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논문번호 : KICS2005-07-303, 접수일자 : 2005년 8월 2일

On the other hand, algebraic hard decision decoding (HDD) algorithms are popular because they lead to low complexity at the receiver [4]. While optimal (unquantized) soft decision decoding (SDD) algorithms provide up to 3 dB coding gain relative to HDD [10], HDD offers the distinct advantages of low decoding complexity and robustness to channel-induced interference such as jamming, impulsive noise, and fading; see e.g., [5] and references therein. In addition, it has been shown that by resorting to reliability information, the error resilience of HDD can be increased while retaining low complexity [5],[12].

Many HARQ schemes with various forward error correction (FEC) algorithms, such as linear block codes, convolutional codes, and turbo codes [3],[15],[16], have been pursued to increase the system throughput [2],[6],[13]. Recently, HARQ schemes combined with ST codes have been proposed [8],[14]. Other researches have shown that the MIMO system can be effective in HARQ protocols [9].

In this paper, we wed the merits of HARQ, ST coding, and HDD to increase the system throughput while maintaining low-complexity. First, we combine all received copies of the same packet by using maximum ratio combining (MRC). Combining packets has the effect of increasing the number of receiver antennas with every retransmission, which means that the number of transmissions per packet equals the diversity order. Therefore, with HARQ and ST coding, transmit and receive diversity is established by increasing the diversity order per retransmission. Moreover, our development of an efficient HDD algorithm using reliability information provides improved error performance at low-complexity.

This combination allows the receiver to perform reliable decoding and hence reduce the probability for retransmission. In this paper, we assume that only acknowledgment (ACK) or negative acknowledgment (NAK) is fed back to the transmitter and the channel state information (CSI) is not available at the transmitter. However, when the CSI feedback is allowed, we may further increase the

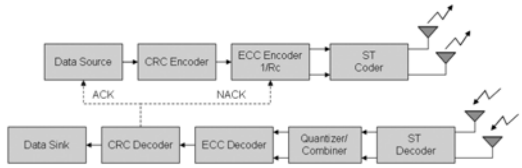

Figure 1. System model

spectral efficiency of the system in time-varying channel conditions by using adaptive modulation and coding (AMC), for which chosen modulation scheme and coding pair based on the channel condition are used [6].

The rest of the paper is organized as follows. In Section II, we describe our system model. The HARQ scheme using reliability information optimally and throughput analysis are shown in Section III, followed by simulation results and conclusions in Section IV and V.

Ⅱ. Preliminaries

Consider a HARQ system equipped with multi-antennas, as shown in Fig. 1. The cyclic redundancy check(CRC) coded packet $b := [b_1, \ldots, b_K]$ of length K is encoded using a rate R_c convolutional code (CC) to result in R_c encoded packets $c^{(j)} := [c_1^{(j)}, \ldots, c_K^{(j)}]$, where $j = 1, \ldots, R_c$, and $c_k^{(j)}$ is an element of GF(2). For simplicity, we will use R_c transmit antennas equal to the number of encoded packets (i.e., code rate). Although we use the convolutional code in this paper, we do not restrict ourselves to any particular coding system. Subsequently, encoded data packets are modulated using BPSK and then encoded using ST coding such as Alamouti [1] and FDFR codes [7]. ST coded packets are forwarded to each antenna and then transmitted over an AWGN channel with variance $\sigma^2 := N_o/2$.

At each receiver, we can obtain the ST decoded sequence $\tilde{c}^{(j)}(n) := [\tilde{c}_1^{(j)}(n), \ldots, \tilde{c}_K^{(j)}(n)]$ for the nth transmission of the same packet $\tilde{c}^{(j)}$, where $c_k^{(j)}(n) \sim N(\pm\sqrt{E}, N_o/(2n))$, and

E denotes the received energy per symbol. We will explain the reduced noise variance of $N_o/(2n)$ in the next section.

Assume that N copies, $\{\tilde{c}^{(j)}(n)\}_{n=1}^{N}$ have been received and CRC checked as unreliable. These packets are combined into a single packet by taking the average of the received values of each re-transmitted packet as follows:

$$r_k^{(j)}(N) = \frac{1}{N}\sum_{n=1}^{N} \tilde{c}_k^{(j)}(n), \tag{1}$$

where $r^{(j)}(N) := [r_1^{(j)}(N), \ldots, r_K^{(j)}(N)]$ for $j = 1, \ldots, R_c$.

Next, the quantizer compares $r_k^{(j)}(N)$ with a decision threshold (zero when a priori probabilities are equal, i.e., $P(0) = P(1) = 1/2$) to output the hard decision sequence $z^{(j)}(N) := [z_1^{(j)}(N), \ldots, z_K^{(j)}(N)]$ with $z_k^j(N) \in \{0, 1\}$. The decoder metric for $c^{(j)} \in C$, denoted by $\mathrm{CM}_{z^{(j)}}(c^{(j)})$, is the Hamming distance defined as:

$$\mathrm{CM}_{\mathbf{z}^{(j)}}(\mathbf{c}^{(j)}) = \sum_{j=1}^{R_c}\sum_{k=1}^{K}\left|c_k^{(j)} - z_k^{(j)}\right|. \tag{2}$$

Then, a codeword $c^{(j)}$ is more likely to be the transmitted codeword than another codeword $c^{(j)\prime}$ if and only if $\mathrm{CM}_{z^{(j)}}(c^{(j)}) \le \mathrm{CM}_{z^{(j)}}(c^{(j)\prime})$; i.e., maximum likelihood (ML) decoding always decodes $z^{(j)}$ to the codeword with the smallest Hamming distance.

However, $r_k^{(j)}(N)$ close to the decision threshold are likely unreliable, which motivates us to puncture those bits belonging to the interval $-\gamma_p < r_k^{(j)}(N) < \gamma_p$, where γ_p is a suitably chosen puncturing threshold, as shown in Fig. 2. The reliability information will be denoted as $\alpha^{(j)} := [\alpha_1^{(j)}, \ldots, \alpha_K^{(j)}]$, with $\alpha_k^{(j)} \in \{0, 1\}$, where $\alpha_k^{(j)} = 0$ indicates that the kth bit in $z^{(j)}(N)$ is unreliable. We call this algorithm *"receive-puncturing"* to differentiate it from the known trans mit-puncturing used to increase FEC rates [2].

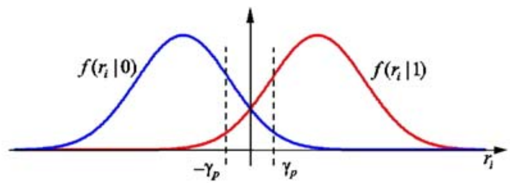

Figure 2. Conditional PDF

Packets $\{z^{(j)}(N)\}_{j=1}^{R_c}$ with corresponding reliability information $\alpha := [\alpha^{(1)}, \ldots, \alpha^{R_c}]$ are processed jointly by the Viterbi decoder to decode the original packet based on the metric

$$\mathrm{CM}_{\mathbf{z}}(\mathbf{c}, \boldsymbol{\alpha}) = \sum_{j=1}^{R_c}\sum_{k=1}^{K}\alpha_k^{(j)}\left|c_k^{(j)} - z_k^{(j)}\right|. \tag{3}$$

We observe from (3) that relative to the conventional HDD, the hardware complexity required by our approach entails just AND gates for implementing α.

If the combined packet turns out to be decoded unreliably by the CRC decoder, a retransmission is requested and the transmitter sends another copy $\{c^{(j)}\}_{j=1}^{R_c}$. The receiver continues to request and combine packets until successful decoding.

Our objective in this paper is to show that the *receive-puncturing* scheme employing reliability information can improve the throughput of HARQ for an ST coded system.

III. HARQ with Reliability

In this section, we optimize the decoder so that the system throughput of HARQ can be increased. In order to attain this goal, we rely on our previous work in [5], where the optimal puncturing threshold γ_p^* enables the codeword decision error probability (CEP) to be minimized at the output of the Viterbi decoder.

3.1 Optimal Puncturing Threshold

To derive γ_p^* that determines the entries of the reliability vector α, we define the *punctured* *code distance* as:

$$d(\boldsymbol{\alpha}) = \sum_{j=1}^{R_c} \sum_{k=1}^{K} \alpha_k^{(j)} \left| c_k^{(j)} - c_k^{(j)'} \right|, \quad (4)$$

where $c^{(j)}, c^{(j)'} \in C$ and $c^{(j)} \neq c^{(j)'}$. We note that $d(\alpha)$ is a random variable corresponding to the code distance after puncturing and takes values between 0 and d (original code distance). Although the puncturing scheme decreases the code distance, the error correction capability can increase due to the enhanced signal-to-noise ratio (SNR) [5]. Specifically, the probability of having $d(\alpha) = m$ is given by

$$\Pr(d(\alpha) = m) = \binom{d}{m} p_d^{d-m}(\gamma_p)\{P_{cp}(\gamma_p) + P_{ep}(\gamma_p)\}^m, \quad (5)$$

where $P_{cp}(\gamma_p)$, $P_{ep}(\gamma_p)$, and $P_p(\gamma_p)$ denote the probabilities of correct symbol, error, and puncturing, which are defined as:

$$P_{cp}(\gamma_p) = \sum_{i=0}^{1} P(i) \int_{S_{cp}^i} f(r \mid i) dr,$$
$$P_{ep}(\gamma_p) = \sum_{i=0}^{1} P(i) \int_{S_{ep}^i} f(r \mid i) dr,$$
$$P_p(\gamma_p) = \sum_{i=0}^{1} P(i) \int_{S_p^i} f(r \mid i) dr, \quad (6)$$

where $f(r|i)$ is the probability density function (PDF) of the observation r conditioned on signal class $i \in \{0,1\}$ as shown in Fig. 2, S_{cp}^i, S_{ep}^i, and S_p^i denote the region that yields $P_{cp}(\gamma_p)$, $P_{ep}(\gamma_p)$, and $P_p(\gamma_p)$, assuming that signal class $i \in \{0,1\}$ is transmitted. These probabilities obey the relationship $P_{cp}(\gamma_p) + P_{ep}(\gamma_p) + P_p(\gamma_p) = 1$. We notice that if $\gamma_p = 0$, then $P_p(0) = 0$, which leads to the conventional one, i.e., $P_{ep}(\gamma_p) + P_{cp}(\gamma_p) = 1$.

To derive the averaged CEP when the *receive-puncturing* scheme is employed, we assume that the all-zero codeword (AZC) is transmitted and the codeword being compared with the AZC has distance d from the AZC. Then, the CEP of the conventional HDD becomes

$$P_{CEP}(d) = \sum_{m=\lfloor (d+1)/2 \rfloor}^{d} \binom{d}{m} P_e^m P_c^{d-m}, \quad (7)$$

where $P_c = 1 - P_e$, and $P_e = P_{ep}(0)$ [10]. If a puncturing process is employed, for code distance d, the CEP can be given by

$$P_{CEP}(d_R(\boldsymbol{\alpha})) = \sum_{j=\left\lfloor \frac{d_R(\boldsymbol{\alpha})+1}{2} \right\rfloor}^{d_R(\boldsymbol{\alpha})} \binom{d_R(\boldsymbol{\alpha})}{j} \bar{P}_{ep}^j(\gamma_p) \bar{P}_{cp}^{d_R(\boldsymbol{\alpha})-j}(\gamma_p), \quad (8)$$

where $\overline{P_{cp}}(\gamma_p) := P_{cp}(\gamma_p)/(1 - P_p(\gamma_p))$, and $\overline{P_{ep}}(\gamma_p) := P_{ep}(\gamma_p)/(1 - P_p(\gamma_p))$ represent the effective correct symbol probability and symbol error probability, respectively. To this end, the CEP averaged with respect to $d(\alpha)$ can be alternatively expressed as [5]

$$\bar{P}_{CEP}(d) = \sum_{m=0}^{d} \Pr(d_R(\boldsymbol{\alpha}) = m) P_{CEP}(d_R(\boldsymbol{\alpha}) = m)$$
$$= \sum_{m=0}^{d} \binom{d}{m} P_p^{d-m}(\gamma_p) \sum_{k=\lfloor (m+1)/2 \rfloor}^{m} P_{ep}^k(\gamma_p) P_{cp}^{m-k}(\gamma_p). \quad (9)$$

We note if $\gamma_p = 0$, i.e., $P_p(\gamma_p) = 0$ in (9), then the average CEP (9) becomes the CEP of the conventional HDD. Since it is difficult to find γ_p^* minimizing $\overline{P}_{CEP}(d)$ in closed form, we use Chernoff's approximation [10], based on which, we obtain the following upper bound on CEP:

$$\bar{P}_{CEP}(d) \le \sum_{m=0}^{d} \binom{d}{m} P_p^{d-m}(\gamma_p) \left\{ 4P_{ep}(\gamma_p) P_{cp}(\gamma_p) \right\}^{\frac{m}{2}}$$
$$= \left(P_p(\gamma_p) + \sqrt{4P_{ep}(\gamma_p) P_{cp}(\gamma_p)} \right)^d. \quad (10)$$

By differentiating the bounded $\overline{P}_{CEP}(d)$ in (10) with respect to γ_p, and setting it equal to zero, the optimal threshold γ_p^* is found to obey

$$\frac{P_{ep}(\gamma_p^*)}{P_{cp}(\gamma_p^*)} = \left\{ \frac{P'_{ep}(\gamma_p^*)}{P'_{cp}(\gamma_p^*)} \right\}^2 , \tag{11}$$

where prime denotes differentiation. We note that (11) does not depend on the code distance d. Choosing the threshold to satisfy (11) is shown in [5] to optimize the decoder performance. By simulation results, we will show that the solution of (11) is unique and the gap between the optimal solution found from (11) and the solution obtained by the numerical search of (9) is relatively small and converges to zero as d and/or SNR increase (for proof, see [5]).

3.2 Throughput Analysis

Now we will show that if we utilize the optimal puncturing threshold that we have derived, then the throughput of HARQ increases. To derive the throughput of HARQ, we assume an ideal retransmission protocol with no delay between packet retransmissions, and the error/delay-free feedback channel. The throughput Γ is then a function of the average number of transmissions N_r by the HARQ protocol.

Let $P(D_u(n))$, $P(D_d(n))$, and $P(D_c(n))$ be the probabilities of a combined received sequence after n transmissions that contain undetected errors, detected errors, and no errors, respectively. Note that $P(D_d(n))$ is equivalent to the event of a retransmission request. These obey the relationship:

$$P(D_u(n)) + P(D_d(n)) + P(D_c(n)) = 1. \tag{12}$$

Assuming that $P(D_u(n))$ is negligible for most CRC codes, N_r can be expressed as

$$\begin{aligned} N_r &= 1 + P(D_d(1)) + P(D_d(1), D_d(2)) + \cdots \\ &\quad + P(D_d(1), D_d(2), \ldots, D_d(n)) + \cdots . \end{aligned} \tag{13}$$

The joint probability can be lower and upper bounded by [2]

$$1 + \sum_{n=1}^{\infty} \prod_{i=1}^{n} P(D_d(i)) \le N_r \le 1 + \sum_{n=1}^{\infty} P(D_d(n)) . \tag{14}$$

By using the fact that $P(D_c(n)) = (1 - P_b(n))^K \ge (1 - P(E(n)))^K$, where $P_b(n)$ denotes BER on the nth transmission, $P(D_d(n))$ can also be upper bounded as:

$$P(D_d(n)) \approx 1 - P(D_c(n)) \le 1 - \left(1 - P(E(n))\right)^K , \tag{15}$$

where $P(E(n))$ is the probability of a decoding error event during Viterbi decoding on the nth transmission. Thus, the upper bound of N_r in (14) is given by

$$N_r \le 1 + \sum_{n=1}^{\infty} \left\{ 1 - \left(1 - P(E(n))\right)^K \right\} . \tag{16}$$

Given the free distance d_{free} of the convolutional code, $P(E)(= P(E(1)))$ can also be bounded by [10]

$$P(E) \le \sum_{d=d_{free}}^{\infty} a_d P_{CEP}(d) , \tag{17}$$

where the distance spectra a_d represents the number of paths with distance d. Notice that the upper bounded CEP of the convolutional HDD is given by

$$P_{CEP}(d) \le \left[4 P_e P_c \right]^{d/2} , \tag{18}$$

and the average CEP of our algorithm with the optimal puncturing threshold γ_p^* is

$$\bar{P}_{CEP}(d) \le \left(P_p(\gamma_p^*) + \sqrt{4 P_{ep}(\gamma_p^*) P_{cp}(\gamma_p^*)} \right)^d . \tag{19}$$

Let us now describe how $P(E(n))$ is determined. To this end, we will use the variance of the received symbol $c_k^{(j)}(n)$ and the effective SNR after n packets are combined:

$$\begin{aligned} \mathrm{var}\left[c_k^{(j)}(n) \right] &= \frac{1}{n} \mathrm{var}\left[c_k^{(j)}(1) \right], \\ \mathrm{SNR}(n) &= n\, \mathrm{SNR}, \end{aligned} \tag{20}$$

where $\mathrm{SNR} := E/\sigma_{\omega}^2$ represents SNR without combining. Consequently, the MRC of n packets increases the effective received SNR by a factor n.

Let $P_{CEP}(d;n)$ and $\overline{P}_{CEP}(d;n)$ be the CEP of the conventional HDD and the CEP of our scheme with n combined packets, respectively. $P_{CEP}(d;n)$ can be calculated by plugging P_e with the reduced noise variance by a factor n into $P_{CEP}(d)$ in (18), while $\overline{P}_{CEP}(d;n)$ can be obtained by choosing the optimal puncturing threshold corresponding to the increased SNR. Recalling that our optimal puncturing thresholds minimize the CEP, we deduce that

$$\overline{P}_{CEP}(d;n) \le P_{CEP}(d;n), \qquad (21)$$

which leads to lower $P(E(n))$ for our algorithm.

Based on this derivation of the average number of transmissions, we can evaluate a lower bound of our throughput performance. Because we utilized a full-rate (rate 1) ST code (e.g., Alamouti code [1]), the throughput can be expressed as $\Gamma := 1/N_r$. Therefore, our assertion that the *receive-puncturing* scheme enhances system throughput has been confirmed.

Ⅳ. Simulation Results

To verify performance, we conduct simulations for both AWGN and Rayleigh fading channels. In all experiments, we use the Alamouti ST code with two transmit and one receive antennas and encode each data block (K = 20000) using a rate $R_c = 1/2$ convolutional code (133,171) with constraint length 7, and minimum code distance 10.

In Fig. 3, we illustrate the puncturing threshold versus CEP. We also plot the minimum points found by simulation and the points corresponding to the puncturing threshold of (11), for which the CEP difference is quite small. It is observed that there exist optimal thresholds that minimize the CEP for each SNR. Noting that $P_{CEP}(d)$ of the conventional HDD corresponds to $\overline{P}_{CEP}(0)$, we deduce the improvement of our scheme.

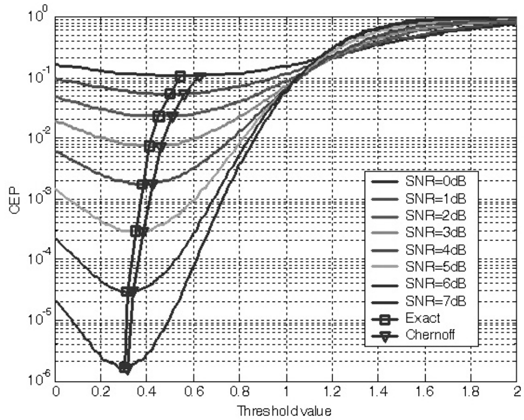

Figure 3. Puncturing threshold vs. CEP

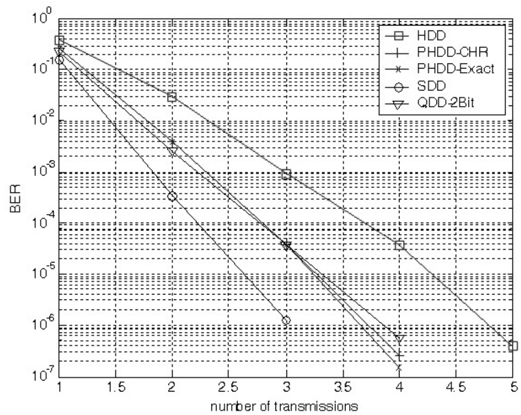

Figure 4. BER vs. number of transmissions in AWGN (SNR = 0 dB)

BER performance versus the average number of transmissions for SNR = 0 dB is shown in Fig. 4. Because the number of bits in our decoding algorithm (punctured (P)HDD) is 2-bit (1-bit for HDD plus 1-bit reliability information), we compare with 2-bit quantized SDD (QDD-2 bit). It is observed that our BER is similar to the 2-bit quantized SDD, or even better when the number of transmissions increases. However, one should recall that the complexity of our algorithm is lower than that of the 2-bit quantized SDD. We also note that the gaps between the performance using the optimal thresholds calculated by the Chernoff bound of CEP as in (10) and those by the exact CEP as in (9), i.e., PHDD-CHR and PHDD-Exact, are relatively small.

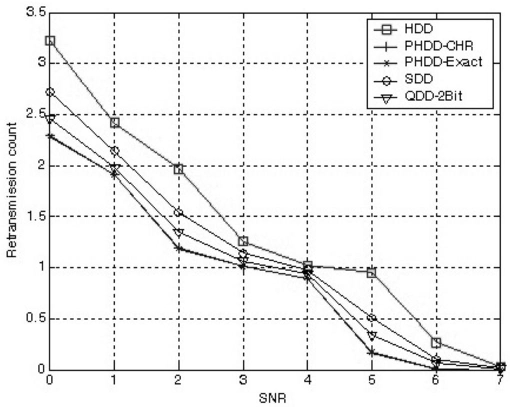

Figure 5. Number of transmissions vs. SNR in AWGN

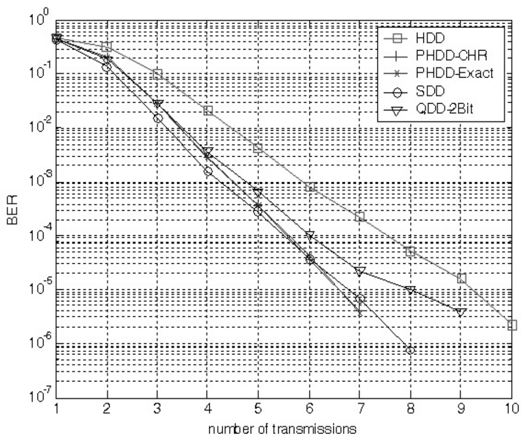

Figure 6. Number of transmissions vs. BER in fading channel (SNR = 6 dB)

Figure 5 depicts the average number of transmissions versus SNR. As expected from Fig. 4, the number of retransmissions is similar to the 2-bit quantized SDD. It is seen that at high SNR, where on the average only one transmission is needed, the error performances of the 4 schemes tend to coincide.

For the Rayleigh fading channel described by Jakes' model with a Doppler frequency 100 Hz for 11 MHz symbol transmission on a 2.4 GHz carrier frequency, we assume that the channel information is perfectly knownto the receiver and zero forcing (ZF) equalization is used. As shown in Fig. 6, our scheme performs well compared with SDD, conventional HDD, and 2-bit quantized SDD.

We also conduct simulations in impulse environment. In many real-world communication systems, non-Gaussian impulsive noise limits the achievable system performance [11]. Impulsive noise is modeled as [17],[18]

$$\eta(n) = a(n)A(n), \tag{22}$$

where $a(n)$ is a binary independent identically distributed occurrence process with $\Pr[a(n)=1] = \varepsilon$, $\Pr[a(n)=0] = 1-\varepsilon$, and ε is the arrival probability; whereas $A(n)$ is a process, with symmetric amplitude distribution, which is uncorrelated with $a(n)$. The variance of $A(n)$ is chosen to be substantially greater than that of $\overline{c_k^{(j)}}(n)$ to represent impulse noise. The mean value of $\eta(n)$ is zero and its variance is $\mathrm{var}[\eta(n)] = \varepsilon \cdot \mathrm{var}[A(n)] = \sigma_\eta^2$. In this simulation, we used $\varepsilon = 10^{-2}$ and $\mathrm{var}[A(n)] = 10^4/12$ as in [18]. The presence of the additional Gaussian measurement noise does not violate the assumptions in our derivation, since it simply adds an additional independent Gaussian component to the received signal. As shown in Figs. 7 and 8, our method shows better performance than SDD and 2-bit quantized SDD in impulse environment because ours has the optimized threshold which is determined by reliability of the received signal.

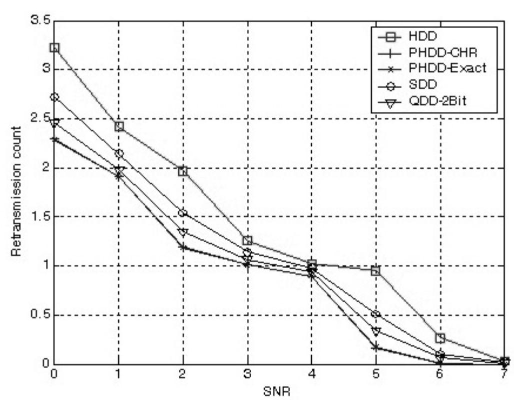

Figure 7. Number of transmissions vs. SNR impulsive noise environment

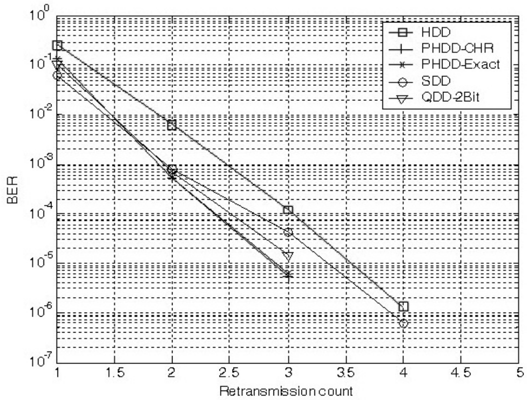

Figure 8. Number of transmissions vs. BER in impulsive noise environment (SNR = 1 dB)

V. Conclusions

We have derived an efficient HARQ protocol equipped with ST coding to enable transmit and receive diversity. By using reliability information, we relied on *receive-puncturing* to further increase the overall system throughput at low decoding complexity thanks to HDD. Simulations illustrated the merits of our algorithm for both AWGN and fading channels as well as impulse noise channels.

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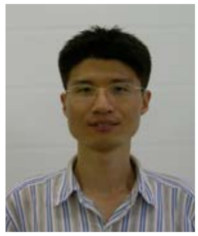

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