Collaborative HARQ Schemes for Cooperative Diversity Communications in Wireless Networks

Kun Pang

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School of Electrical & Information Engineering
The University of Sydney

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To my family

Abstract

Wireless technology is experiencing spectacular developments, due to the emergence of interactive and digital multimedia applications as well as rapid advances in the highly integrated systems. For the next-generation mobile communication systems, one can expect wireless connectivity between any devices at any time and anywhere with a range of multimedia contents. A key requirement in such systems is the availability of high-speed and robust communication links. Unfortunately, communications over wireless channels inherently suffer from a number of fundamental physical limitations, such as multipath fading, scarce radio spectrum, and limited battery power supply for mobile devices.

Cooperative diversity (CD) technology is a promising solution for future wireless communication systems to achieve broader coverage and to mitigate wireless channels' impairments without the need to use high power at the transmitter. In general, cooperative relaying systems have a source node multicasting a message to a number of cooperative relays, which in turn resend a processed version message to an intended destination node. The destination node combines the signal received from the relays, and takes into account the source's original signal to decode the message. The CD communication systems exploit two fundamental features of the wireless medium: its broadcast nature and its ability to achieve diversity through independent channels.

A variety of relaying protocols have been considered and utilized in cooperative wireless networks. Amplify and forward (AAF) and decode and forward (DAF) are two popular protocols, frequently used in the cooperative systems. In the AAF mode, the relay amplifies the received signal prior to retransmission. In the DAF mode, the relay fully decodes the received signal, re-encodes and forwards it to the destination. Due to the retransmission without decoding, AAF has the shortcoming that noise accumulated in the received signal is amplified at the transmission. DAF suffers from decoding errors that can lead to severe

error propagation. To further enhance the quality of service (QoS) of CD communication systems, hybrid Automatic Repeat-reQuest (HARQ) protocols have been proposed. Thus, if the destination requires an ARQ retransmission, it could come from one of relays rather than the source node.

This thesis proposes an improved HARQ scheme with an adaptive relaying protocol (ARP). Focusing on the HARQ as a central theme, we start by introducing the concept of ARP. Then we use it as the basis for designing three types of HARQ schemes, denoted by HARQ I-ARP, HARQ II-ARP and HARQ III-ARP. We describe the relaying protocols, (both AAF and DAF), and their operations, including channel access between the source and relay, the feedback scheme, and the combining methods at the receivers.

To investigate the benefits of the proposed HARQ scheme, we analyze its frame error rate (FER) and throughput performance over a quasi-static fading channel. We can compare these with the reference methods, HARQ with AAF (HARQ-AAF) and HARQ with perfect distributed turbo codes (DTC), for which correct decoding is always assumed at the relay (HARQ-perfect DTC). It is shown that the proposed HARQ-ARP scheme can always performs better than the HARQ-AAF scheme. As the signal-to-noise ratio (SNR) of the channel between the source and relay increases, the performance of the proposed HARQ-ARP scheme approaches that of the HARQ-perfect DTC scheme.

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Statement of Originality

All material presented in this thesis is the original work of the author, unless otherwise stated. The content of this thesis has not been previously submitted for examination as part of any academic qualifications. Most of the results contained herein have been submitted for publication, in journals or conferences of international standing. The author's contribution in terms of published material is listed in the next section, "Publications".

The original motivation to pursue research in this field was provided by my thesis supervisors, Professor Branka Vucetic and Dr Yonghui Li in the School of Electrical and Information Engineering, the University of Sydney, Australia.

Kun Pang
Telecommunication Laboratory
School of of Electrical and Information Engineering
The University of Sydney
New South Wales, Australia
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Publications

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List of Acronyms

2G	Second-Generation
3G	Third-Generation
4G	Fourth-Generation
AAF	Amplify and Forward
ACK	A Positive Acknowledgement
AM	Amplitude Modulation
APP	A Posteriori Probability
ARP	Adaptive Relaying Protocol
ARQ	Automatic Repeat-reQuest
AWGN	Additive White Gaussian Noise
BER	Bit Error Rate
BPSK	Binary Phase Shift Keying
CD	Cooperative Diversity
CPC	Complementary Punctured Convolutional
CRC	Cyclic Redundancy Check
CSI	Channel State Information
DAF	Decode and Forward
DAF-DTC	Decode and Forward with Distributed Turbo Coding
DTC	Distributed Turbo Coding
EI	Extrinsic Information
FCS	Frame Check Sequence
FEC	Forward Error Control
FER	Frame Error Rate
FSK	Frequency-Shift Keying
FSM	Finite State Machine
GBN	Go-Back-N

GSM	Global System for Mobile Communications
HARQ	Hybrid ARQ
IR	Increment Redundancy
LLR	Log-Likelihood Ratio
LOS	Line-of-Sight
LT	Luby Transform
MAP	Maximum a Posteriori Probability
MDF	Moment Generating Function
MIMO	Multiple-Input Multiple-Output
MLD	Maximum Likelihood Decoding
NAK	Negative Acknowledgement
pdf	probability density function
PEP	Pairwise Error Probability
QoS	Quality of Service
PSK	Phase-Shift Keying
RCPC	Rate Compatible Punctured Convolutional
RCPT	Rate Compatible Punctured Turbo
RSC	Recursive Systematic Convolutional
SISO	Soft-Input Soft-Output
SNR	Signal-to-Noise Ratio
SOVA	Soft Output Viterbi Algorithm
SR	Selective-Repeat
SW	Stop-and-Wait
TDD	Time Division Duplexing
UEP	Unequal Error Protection
VA	Viterbi Algorithm
WEP	Word Error Probability

Chapter 1

Introduction

This chapter describes the background and motivation for this research work by briefly introducing the field and explaining the principal research problems. A concise outline for the remainder of the thesis is provided at the end of the chapter.

1.1 Background

Pick up any newspaper today and it is a safe bet that you will find an article somewhere relating to mobile communications, which are affecting virtually everyone's life. To date, second-generation (2G) communication systems, such as the Global System for Mobile communications (GSM), have been widely implemented across the globe, providing high-quality speech service [1]. Since the evolution from 2G towards third-generation (3G) has not brought any substantial new service, it is not enough to encourage the customers to change their equipment [2]. Following the paradigm of generational changes, fourth-generation (4G) wireless and mobile networks are beginning to pave the way for the future.

Basically, many prophetic visions have appeared in literature presenting the future generation as the ultimate boundary of the wireless mobile communication without any limit in its potential, but not giving any practical designing rule and thus any definition of it. Recently, a pragmatic methodology, centered on a user-centric approach, which leads to a novel vision of the 4G and the definition of its key features and technological development, has been proposed [3, 4]. Along with this view, 4G will be a convergence platform that will provide clear advantages in terms of bandwidth, coverage, power consumption and spectrum usage, thus also offering a variety of new heterogeneous services. Although the core of this technology is still cellular, the network architecture will predominantly rely on short-range communication systems, wherein *the users may cooperate in a completely distributed or cellular-controlled fashion*. Therefore, the concept of node cooperation introduces a new form of diversity, *spatial diversity*, which results in an increased reliability of the communication, leading both to the extension of the coverage and the minimization of the power consumption. Furthermore, *cooperative transmission* strategies increase the end-to-end capacity and hence the spectral efficiency of the system.

Actually, research on spectrally efficient wired/wireless communications has been gaining momentum since the fundamental channel capacity findings of Shannon in 1948 [5]. In his discovery, Shannon showed that there is a parameter intrinsic to a channel, referred to as the channel capacity, that acts as the fundamental limit of the maximum information transfer rate over a noisy channel. Also, he showed that arbitrarily reliable communication can be achieved if we signal at information rates less than the channel capacity. However, Shannon only gave an existence proof of his theory, which did not indicate a constructive scheme that can approach this theoretical limit. Since the dawn of information theory, in the enduring years, a large amount of research was conducted into the construction of specific codes with good error-correcting capabilities and the development of efficient decoding algorithms for these codes.

Turbo codes [6], as one of the most powerful types of forward-error-control (FEC) codes, caused a great stir in the coding community and have prompted a great deal of research. A turbo code is formed from the parallel concatenation of two constituent codes separated by an interleaver. Each constituent code may be any type of FEC code used for conventional data communications. The interleaver is a critical part of turbo codes; which should have the capability of breaking up the low-weight input sequence so that the permuted sequence has a large distance between the 1's. Hence, even if the first encoder generates a low-weight output sequence due to a certain low-weight input pattern, the second encoder is likely to generate high-weight output sequences and the resulting overall codeword is still of high weight. Because of this feature, turbo codes have exceptionally good performance, particularly at moderate bit error rates (BER) and for large block lengths. In fact, for essentially any code rate and information block lengths greater than about 10^4 bits, turbo codes with iterative decoding can achieve BERs as low as 10^{-5} at signal-to-noise ratios (SNRs) within 1 dB of the Shannon limit [7].

Although turbo codes could achieve energy efficiencies within 1 dB of the Shannon capacity, as a FEC code, turbo codes still have inevitable defects. When a received packet is detected in error, it must be decoded, and the decoded packet has to be forwarded to the user regardless of the decoding result. When the channel condition is worse, turbo decoders might even produce worse results than uncoded systems, making it difficult to achieve high reliability in FEC schemes [8].

The drawback of the FEC system can be overcome if it is combined with a retransmission scheme, ARQ. Such a combination is referred to as a hybrid ARQ (HARQ) [9, 10]. A straightforward HARQ scheme simply combines FEC and ARQ schemes together, and a code is designed for simultaneous error correction and error detection. Once the received codeword is detected in error and the designed error correcting capability cannot correct the errors, the receiver requests a retransmission. This HARQ scheme is called type-I HARQ. To accommodate different error protection requirements, or a channel with unknown or time-varying parameters, a more flexible HARQ scheme is desirable. Unlike the type-I HARQ, which has a fixed code rate, the type-II HARQ scheme uses continuous rate variations to change from low to high error protection within a codeword. This is done by transmitting supplemental code symbols, when they are needed. Such a scheme can provide a higher throughput if the channel is quiet [7, 9].

Consequently, Hagenauer [11] introduced rate compatible punctured convolutional (RCPC) codes with such an application in mind. A family of codes with particular rates is obtained by puncturing a low rate code periodically. Therefore, RCPC-ARQ protocol falls into the so-called class of incremental redundancy (IR) codes, in which parity check digits are incrementally transmitted to adaptively meet the error performance requirements of the system. A few years later, the use of turbo codes in an ARQ protocol was proposed [12], and the rate-compatibility requirement of the turbo codes in a HARQ system was considered in [13] to achieve a high throughput over an additive white Gaussian (AWGN) channel.

It has been shown that turbo codes can provide tremendous coding gains in AWGN channels [6]. However, in fading environments, turbo codes lose much of their power, particularly in quasi-static fading channels [14]. To deal with fading effects, *spatial diversity* techniques have been provided to effectively mitigate the performance deterioration without imposing delay or bandwidth expansion [15]. Generally, spatial diversity is obtained when signals are transmitted from antennas separated far enough to experience independent fading channels.

The multiple-input multiple-output (MIMO) technique is a successful example that can provide both spatial and temporal diversities to effectively mitigate the detrimental effects of fading for point-to-point channels [16–18]. This system is implemented typically with an antenna array at both the transmitter and receiver in conjunction with the employment of space-time coding, aiming to create separate transmission paths subject to independent fades.

The advantages of the MIMO technique have been widely acknowledged. However, due to size, cost, or hardware limitations, most wireless handsets may not be able to support multiple transmission antennas. This problem restricted the performance gains and was solved by a new form of spatial-temporal diversity, referred to as cooperative diversity (CD) [19–21], wherein multiple users share antennas which form a virtual antenna array. A classic example of the CD technique can actually be traced back to the groundbreaking work on the relay channel in 1977 [22]. In this work, the author analyzed the capacity of the three-node network consisting of a source, a relay and a destination. It was assumed that all nodes operate in the same band, so the system can be decomposed into a broadcast channel from the viewpoint of the source and a multiple access channel from the viewpoint of the destination, as shown in Fig.1.1.

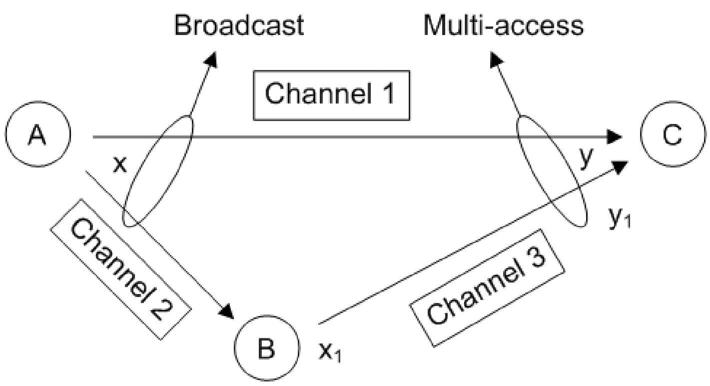

Figure 1.1: The relay channel

1.2 Research Problems and Methodology

To fully exploit spatial diversity in the CD communication systems, a variety of relaying protocols were proposed in [21]. These protocols blend different fixed relaying modes, specifically amplify and forward (AAF), which lets the relay amplify and retransmit this noise version to the destination, and decode and forward (DAF), which allows the relay to decode and re-encode the received message before forwarding it to the destination. With strategies based upon adapting to channel state information (CSI) between cooperating source terminals, selection relaying, as well as incremental relaying, which is exploiting limited feedback from the destination terminal, were also developed in [21].

In the above methods, the relay repeats the received message symbols. Recently, a different framework, called coded cooperation was extended by integrating the channel-coding into the existing cooperation scheme [23], where the symbols are not repeated by the relay. The idea of coded cooperation is to use the same overall rate for coding and transmission (thus no more system resources are used), however, the coded symbols are re-arranged between the source and relay such that better diversity is attained. In general, various channel coding methods can be used within this coded cooperation framework. For example, in [23] RCPC codes were employed. In [24] space-time codes were used and in [25] distributed turbo codes were introduced.

A simple feedback HARQ protocol can further improve communication reliability in cooperative communication systems. In such a system, the appropriate retransmitted signal could come from the relay rather than the source [26]. Some studies on HARQ with relaying protocols have appeared in literature. For example, in [27], a HARQ I-DAF scheme was proposed. In this scheme, upon the reception of a negative acknowledgement (NAK), the relay retransmits a copy of the original packet, received from the source, to the destination. In a HARQ II-DAF scheme [28], the source broadcasts odd-numbered symbols of the codeword, the relay first decodes the received message and then re-encodes it by using convolutional codes. It then sends the parity check symbols, which form even-numbered symbols of the codeword to the destination when it decodes correctly. In [29], a Pure-ARQ (which means that if a packet is wrongly decoded, the receiver discards this packet and asks for a retransmission from the source), HARQ I, HARQ II schemes with different cooperative strategies, including DAF and AAF protocols, have been further investigated.

A survey of the current state-of-the art in the CD communication systems shows that almost all the existing HARQ schemes are based on the fixed relaying protocols, either AAF or DAF. However, these HARQ schemes with an AAF or a DAF protocol suffer from either noise amplification or error propagation. In addition, there are limited studies on the performance of the HARQ schemes for the CD communication systems. Preliminary works on the analysis of HARQ schemes for cooperative communications have focused on the DAF [30] and IR protocols [31]. In [30], a performance analysis of the HARQ I-DAF scheme with/without using Chase combining [32] is presented. In [31], the IR protocol with RCPC code in a two-user cooperative diversity system is analyzed.

In this thesis, we propose an improved HARQ scheme with an adaptive relaying protocol (HARQ-ARP). The proposed HARQ-ARP scheme combines the retransmission mechanisms (repetition coding and incremental redundancy), the distributed turbo coding (DTC) and the adaptive relaying strategy, which is a combination of DAF and AAF. Based on the cyclic redundancy check (CRC) codes, the relay decides which one of the relaying protocols will be used during the retransmission. If the decoding result is correct, the relay uses a DAF protocol to interleave the decoded signal, re-encode and forward it to the destination; otherwise, the relay uses the AAF scheme. The ARP can effectively avoid the problem of both error propagation and noise amplification encountered in current cooperative communication systems. In addition, since the decoding of the ARP at the receiver is the same as for AAF and DAF, the decoding process of the ARP has the same complexity as non-adaptive relaying protocols. Moreover, the feedback scheme used in this thesis entails broadcasting a positive (ACK) or negative acknowledgement (NAK) from the destination to both the source and relay. The decision about which one of the two transmits in case of NAK is based on the number of NAKs.

The performance of the proposed ARP scheme in conjunction with type I, II and III HARQ protocols is also analyzed. We derive pairwise error probabilities (PEP) and word error probabilities (WEP) expressions for the proposed and reference schemes, including HARQ with AAF (HARQ-AAF) scheme and HARQ with perfect distributed turbo codes (HARQ-perfect DTC) scheme. Based on them, we develop a frame error rate (FER) expression and a general throughput expression, which can be applied in each type of HARQ scheme. The analytical results are validated by comparing with the simulation results. Both simulation and analytical results show that the proposed HARQ-ARP scheme can achieve a superior FER and throughput, relative to the reference methods in all SNR regions.

1.3 Thesis Outline

As a guide to reading this thesis, its structure and contribution are briefly summarized as follows.

Chapter 1 gives a short review of the history of the cooperative diversity communication system and the hybrid ARQ applications in this system, as well as the motivation of our work.

Chapter 2 provides some fundamental background related to this thesis. It covers the basic elements of a digital communication system, fading channels, two basic categories of error control schemes including FEC and ARQ, as well as their combination, a HARQ scheme.

Chapter 3 presents the concept of cooperative diversity and gives an introduction to the current solution, CD techniques. Some classic relaying protocols of the CD communication systems are briefly reviewed. The last part of this chapter introduces the coded cooperation schemes.

Chapters 4 and 5 contain the main contribution of the thesis. In Chapter 4, we propose three types of the HARQ-ARP protocols, including type I HARQ-ARP, type II HARQ-ARP and type III HARQ-ARP, in a two-hop CD communication system. Rather than the source-to-destination transmission, the relay takes part in the packet retransmission process. We propose an adaptive relaying protocol, which overcomes the disadvantages of the fixed one. The performance of the proposed and reference schemes is compared through simulations.

Chapter 5 deals with the performance analysis of our proposed HARQ-ARP scheme over a quasi-static fading channel. We calculate the PEPs, WEPs, FER and a general throughput expression, which can be applied in each type of HARQ scheme. In addition, the performance comparison between the proposed scheme and reference schemes is investigated by analytical and simulation results.

Chapter 6 concludes the thesis by summarizing the main results and discussing potential future work.

Chapter 2

Background

This chapter presents the ideas and techniques fundamental to the digital communication systems. We start by providing a brief introduction to wireless channel characteristics and fading channels. Then we proceed by reviewing two categories of techniques for controlling transmission errors in data transmission systems. First technique includes forward-error control (FEC), which covers redundancy check codes, convolutional codes and turbo codes as well as associated decoding algorithms such as maximum likelihood (ML) decoding, the Viterbi algorithm (VA) and iterative soft output Viterbi algorithm (SOVA) decoding. Second technique introduces automatic repeat-reQuest (ARQ) scheme and the proper combination of ARQ and FEC, which is referred to as a HARQ scheme.

2.1 Digital Communication Systems

Digital communication systems are becoming increasingly attractive because of the ever-growing demand for data communication and because digital transmission offers data-processing options and flexibilities not available with analog transmission. The block diagram of a typical digital communication system with basic elements is illustrated in Fig. 2.1 [33].

The source output may be either an analog signal, such as a video signal, or a digital signal, such as the output of a teletype machine, which is discrete in time and has a finite number of output characters.

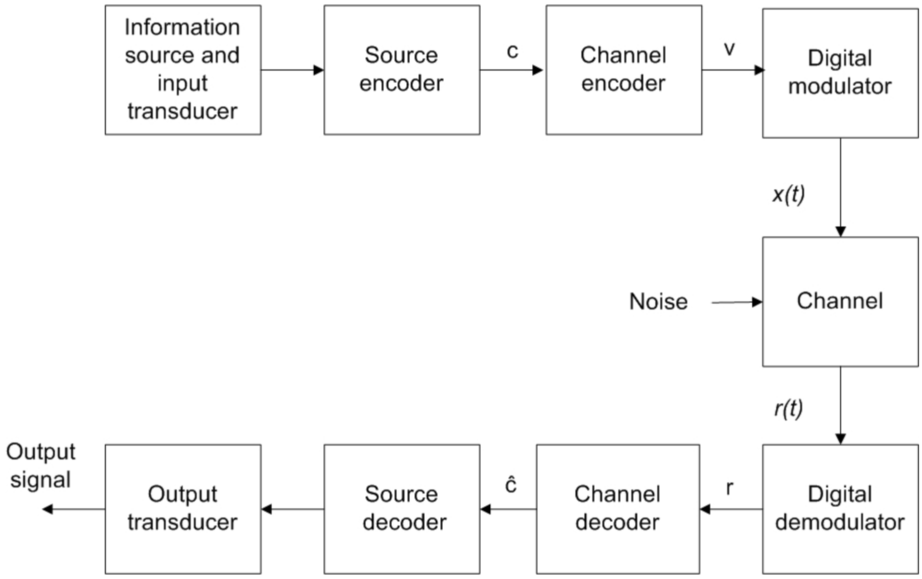

Figure 2.1: Block diagram of a typical digital communication system

To seek an efficient representation of the source output, the source encoder removes the unnecessary redundancy from the source output by converting the output of either an analog or a digital source into a sequence of binary digits, c, which is called information sequence. When c is passed to the channel encoder, the channel encoder introduces controlled redundancy into the binary information sequence, denoted by v, to overcome the effects of noise and interference encountered during the transmission through the channel. The added redundancy serves to increase the reliability of the received data and improves the fidelity of the received signal. The digital modulator will transform v into appropriate electrical waveforms suitable for transmission over the channel, based on the channel characteristics. Some of the most commonly used modulation techniques are amplitude modulation (AM), phase-shift keying (PSK) and frequency-shift keying (FSK). The modulated signal, $x(t)$, will then be sent to the channel for transmission.

The channel is the physical medium that is used to send the signal from the transmitter to the receiver. The channel can be either atmosphere (in wireless transmission), or other physical media, including wire lines, optical fiber cables and so on. During the transmission over the channel, the transmitted signal is inevitably corrupted in a random manner by various possible mechanisms, such as thermal noise generated by electronic devices and atmospheric noise. Additive white Gaussian noise [33] is most often used to model the noise in communication systems.

At the receiving end of the digital communication systems, the digital demodulator processes the corrupted transmitted waveform, $r(t)$, and reduces the waveforms to a sequence of numbers that represent estimates of the transmitted data symbols. The channel decoder attempts to reconstruct an estimate of the information sequence. As a final step, the source decoder accepts the output sequence, $\hat{c}$, from the channel decoder and attempts to reconstruct the original signal, with the knowledge of the source encoder used. Ideally, the produced estimated sequence is a replica of the source information. In this thesis, the focus is placed on mobile wireless channels characterized by fading, which will be elaborated in section 2.2.

2.2 Fading Channels

In wireless communications, the presence of reflectors in the environment surrounding a transmitter and receiver creates multiple paths. As a consequence, the receiver will receive the reflected, diffracted and scattered signals from all directions. The receiver sees the superposition of multiple copies of the transmitted signal, which experienced differences in attenuation delay and phase shift while traversing a different path; such a phenomenon is called multipath fading [33]. This can result in either constructive or destructive interference, amplifying or attenuating the signal power seen at the receiver.

2.2.1 Statistical Models for Fading Channels

Depending on the nature of the radio propagation environment, there are different models describing the statistical behavior of the multipath fading envelope. In this section, we will introduce Rayleigh and Rician fading models, used to describe signal variations in a multipath environment.

2.2.1.1 Rayleigh Fading

Rayleigh fading is the most applicable to model heavily built-up city centers where there is no line-of-sight (LOS) between the transmitter and receiver, and many buildings and other objects attenuate, reflect, refract and diffract the signal [34].

The Rayleigh distribution is frequently used to model multipath fading with no direct LOS path [35]. The square root of a sum of two zero-mean identically distributed Gaussian random variables has a Rayleigh distribution [33]. It can be assumed that the real and imaginary parts of the response are modeled by an independent and identically distributed zero-mean Gaussian process, so the amplitude of the response is the sum of two such processes. If r is defined as a Rayleigh distribution random variable, the probability density function (pdf) is given by [33]

$$p_{(R)}(r) = \frac{r}{\sigma^2} e^{-r^2/2\sigma^2}, \qquad r \geqslant 0, \tag{2.1}$$

where σ^2 is the variance of two zero-mean identically distributed Gaussian random variables.

2.2.1.2 Rician Fading

Rician fading is a stochastic model for radio propagation anomaly caused by partial cancelation of a radio signal. The received signal consists of a direct wave and a number of reflected waves. The direct wave is a stationary non-fading signal and is called the specular coherent component of the received signal [36]. The reflected waves are independent random multipath signals, which constitute the scattered component of the received signal. When the number of reflected waves becomes large, the scattered component can be characterized as a complex Gaussian random process with a zero mean and variance. If r is defined as a Rician distribution random variable, the pdf of Rician is shown as below [37]:

$$p(r) = \begin{cases} \frac{r}{\sigma^2} e^{-\frac{\left(r^2+A^2\right)}{2\sigma^2}} I_0\left(\frac{rA}{\sigma^2}\right) & r \geqslant 0, A \geqslant 0 \\ 0 & \text{otherwise,} \end{cases} \tag{2.2}$$

where σ^2 is the variance of two zero-mean identically distributed Gaussian random variables, A denotes the peak magnitude of the non-faded signal component and $I_0(\cdot)$ is the modified Bessel function of the first kind and zero order [38]. The Rician distribution is often described in terms of a parameter K, which is defined as the ratio of the power in the specular component to the power in the multipath signal, which is given by

$$K = \frac{A^2}{2\sigma^2}. \tag{2.3}$$

As we can see, when K approaches zero, the Rician pdf approaches a Rayleigh pdf as given in Eq.(2.1).

2.2.2 Slow and Fast Fading

In this thesis, we classify fading channels into fast and slow ones [35]. The term “fast fading” is used for describing channels in which $T_0 < T_s$, where T_0 is the channel coherence time, and T_s is the time duration in which the channel behaves in a correlated manner is short compared with the time duration of a symbol. Therefore, it can be expected that the fading character of the channel will change several times during the time span of a symbol, leading to distortion of the baseband pulse shape. Fast fading causes the baseband pulse to be distorted, often resulting in an irreducible error rate.

A channel is generally referred to as introducing slow fading if $T_0 > T_s$. Here, the time duration in which the channel behaves in a correlated manner is long compared with the time duration of a transmission symbol. Thus, one can expect that the channel state to virtually remain unchanged during the time in which a symbol is transmitted. The primary degradation in a slow-fading channel, as with flat fading, is loss in SNR. A similar fading model can be referred to as a quasi-static fading [39] if the fading coefficients change independently from one frame to another.

Clearly, slow and fast fading stand for two extremes of actual fading scenarios.

2.3 Error Control Scheme - Forward Error Correction

In recent years, there has been an increasing demand for efficient and reliable digital data transmission and storage systems. To reliably reproduce the data, a major concern of the system design is the control of errors. With his 1948 paper, "A Mathematical Theory of Communication", Shannon [5] demonstrated that, by proper encoding of the information, errors induced by a noisy channel or storage medium can be reduced to any desired level without sacrificing the rate of information transmission or storages, as long as the information rate is less than the capacity of the channel. The fundamental philosophical contribution inspired the subsequent research in the error control coding areas. One of these approaches is the application of coding, that is, the use of error-correcting or error-detecting codes [40–42].

The error-correcting codes are used to combat transmission errors in an FEC communication system by introducing more redundant parity bits. When the receiver detects the presence of errors in a received packet, it attempts to correct the errors at first. If the receiver fails to do so, the erroneous decoded message will be delivered to the user. However, for the error-detecting codes, which are used in an ARQ communication system, only a few parity-check bits are appended with the message. Later, the new packet is transmitted over the channel to the receiver. At the receiver, if the decoding is unsuccessful, then the retransmission is required.

In this section, we briefly reviews the classic error-correcting codes and the corresponding decoding methods, which serve as the basis for the development and design of the cooperative diversity systems.

2.3.1 Cyclic Redundancy Check Codes

One of the most common, and one of the most powerful, error-detecting codes is the cyclic redundancy check [43], which belongs to the shortened cyclic codes. In conjunction with ARQ protocol, CRC is used particularly in data communications. The basic idea can be described as follows.

Given a block of f bits, or message, the transmitter generates $(k - f)$-bit parity-check bits, known as a frame check sequence (FCS), such that the resulting frame, the coded message, consisting of k bits, is exactly divisible by some predetermined number. The receiver then divides the incoming frame by that number and, if there is no remainder, assumes there was no error during the transmission.

To clarify the CRC algorithm, a way of viewing the CRC process is to express all values as polynomials in a dummy variable X, with binary coefficients. The coefficients correspond to the bits in the binary number. Let us define

- $T(X)$ = k-bit frame to be transmitted,
- $D(X)$ = f-bit block of data, or message, the first f bits of $T(X)$,
- $P(X)$ = pattern of $k - f + 1$ bits; this is the predetermined divisor.

We would like $T(X)/P(X)$ to have no remainder; $T(X)$ can be exactly divisible by $P(X)$. It should be clear that

$$\frac{X^{k-f}D(X)}{P(X)} = Q(X) + \frac{R(X)}{P(X)}. \tag{2.4}$$

There is a quotient and a remainder. Then we can get

$$T(X) = X^{k-f}D(X) + R(X). \tag{2.5}$$

At the receiver, to verify that there is no error during the transmission, consider

$$\frac{T(X)}{P(X)} = \frac{X^{k-f}D(X)}{P(X)} + \frac{R(X)}{P(X)}. \tag{2.6}$$

Substituting Eq.(2.4) into Eq.(2.6), we have

$$\frac{T(X)}{P(X)} = Q(X) + \frac{R(X)}{P(X)} + \frac{R(X)}{P(X)}. \tag{2.7}$$

Since any binary number added to itself modulo 2 yields zero, thus

$$\frac{T(X)}{P(X)} = Q(X) + \frac{R(X) + R(X)}{P(X)} = Q(X). \tag{2.8}$$

There is no remainder, and therefore $T(X)$ is exactly divisible by $P(X)$. A CRC-16 = $X^{16} + X^{15} + X^2 + 1$ [44] was adopted as our error detection system in this thesis.

2.3.2 Convolutional Codes

A convolutional code is described by three integers, which are the number of input symbols, k, the total number of output symbols, n, and memory order, m. The n-tuple emitted by the convolutional encoding procedure is not only a function of an input k-tuple, but is also a function of the previous m k-input tuples. Fig.2.2 [45] shows a simple rate $1/2$ convolutional code encoder, which is a linear feedforward shift register. The connection between the shift register elements and the modulo 2 adders can be conveniently described by the following two generator sequences:

$$g^{(1)} = (g_0^{(1)} g_1^{(1)} g_2^{(1)}) = (101)$$
$$g^{(2)} = (g_0^{(2)} g_1^{(2)} g_2^{(2)}) = (111). \tag{2.9}$$

If $c = (\cdots, c_{-1}, c_0, c_1, \cdots, c_l, \cdots)$ is the input data stream, then the two output sequences, denoted by $v^{(1)} = (\cdots, v_{-1}^{(1)}, v_0^{(1)}, v_1^{(1)}, \cdots, v_l^{(1)}, \cdots)$ and $v^{(2)} = (\cdots, v_{-1}^{(2)}, v_0^{(2)}, v_1^{(2)}, \cdots, v_l^{(2)}, \cdots)$ can be obtained as

$$v^{(i)} = c * g^{(i)}, \quad i = 1, 2 \tag{2.10}$$

where $*$ denotes the convolutional operator.

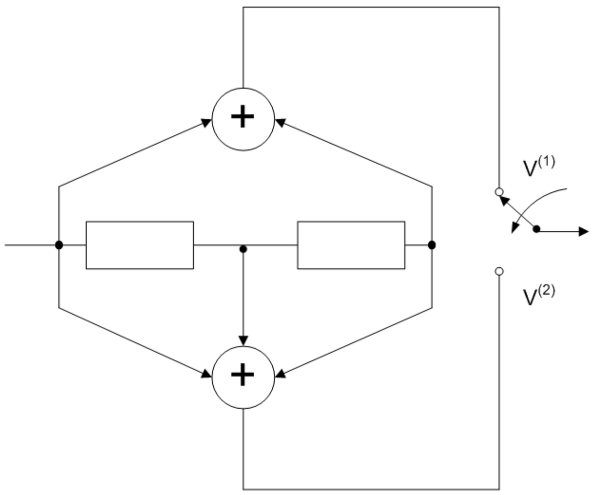

Figure 2.2: A (2,1,2) convolutional encoder

The performance of a convolutional code depends on the decoding algorithm and distance property. If a hard-decision decoding algorithm is used, the code performance is measured by Hamming distance. The minimum free distance of a convolutional code is defined as the minimum Hamming distance between any two code sequences, which is the minimum weight of all non-zero code sequences of any length. If a soft-decision decoding algorithm is used, the code performance is measured by Euclidean distance. The minimum free Euclidean distance is defined as the minimum Euclidean distance between any two code sequences. It depends on both the convolutional code trellis and modulation type.

2.3.3 Maximum Likelihood Decoding

The above section explains the operation of a convolutional encoder. To consider the decoding of a convolutional code, first, we describe the underlying theory of maximum likelihood decoding, and then present a practical implementation of the decoding algorithm, the Viterbi algorithm [7, 46].

In Fig.2.1, the task of the decoder is to produce an estimate $\hat{c}$ of the information sequence c based on the received sequence r. Since there is a one-to-one correspondence between the information sequence c and the codeword v, the decoder can produce an estimate $\hat{v}$ of the codeword v. A decoding rule is a strategy for choosing an estimated codeword $\hat{v}$ for each possible received sequence r. If the codeword v is transmitted, a decoding error occurs if and only if $\hat{v} \neq v$. Given that r is received, the conditional error probability of the decoder is defined as [7]

$$P(E|r) \triangleq P(\hat{v} \neq v|r). \tag{2.11}$$

Then the error probability of the decoder can be calculated as

$$P(E) = \sum_{r} P(E|r)P(r), \tag{2.12}$$

in which $P(r)$ is the probability of the received sequence r and is independent of the decoding rule employed. Hence, an optimum decoding rule, which minimizes $P(E)$, must minimize $P(E|r)$ in Eq.(2.11). Since minimizing $P(\hat{v} \neq v|r)$ is equivalent to maximizing $P(\hat{v} = v|r)$, therefore, by using Bayes' rule we get

$$P(v|r) = \frac{P(r|v)P(v)}{P(r)}. \tag{2.13}$$

So, $\hat{v}$ is chosen as the most likely codeword, given that r is received. In most cases, all information sequences, and hence all codewords, are equally likely in practice. Therefore, maximizing Eq.(2.13) is equivalent to maximizing the $P(r|v)$. A decoder that selects its estimate by maximizing $P(r|v)$ is called a maximum likelihood decoder.

2.3.4 Viterbi Algorithm

The Viterbi algorithm was proposed for the decoding of convolutional codes. It is applicable as a solution to various communication estimation problems, as long as the system can be modeled as a finite state machine (FSM) and represented by a time-invariant or time-varying trellis diagram [47]. We consider the system model using convolutional codes. The VA performs MLD by tracing through the trellis of the code. The trellis diagram of a (n,k,m) convolutional code has 2^{km} states and 2^k branches leaving and entering each state.

Assume an information sequence $c = (c_0, c_1, \cdots, c_{h-1})$ of length kh, is encoded into a codeword $v = (v_0, v_1, \cdots, v_{h+m-1})$ of length $N = n(h + m)$. The received sequence is denoted by $r = (r_0, r_1, \cdots, r_{h+m-1})$. As a result, the decoder needs to produce an estimate $\hat{v}$ of the codeword v based on the received sequence r. The criterion is to let the decoder choose $\hat{v}$ as the codeword v that maximizes the log-likelihood function $logP(r|v)$ [7]

$$P(r|v) = \prod_{i=0}^{h+m-1} P(r_i|v_i), \tag{2.14}$$

The equation is equivalent to

$$logP(r|v) = \prod_{i=0}^{h+m-1} logP(r_i|v_i), \tag{2.15}$$

where $logP(r|v)$ is called the path metric and $logP(r_i|v_i)$ is the branch metrics for branch i [45].

The VA traces through all possible paths in the trellis and at each step compares the metrics of all paths entering each state and stores the path with the largest metric, called the survivor. The survivor at each state is then stored along with its metric. The VA can be summarized as follows:

- Step 1. At time $t = 1$, compute the branch metric for the single branch entering each state. Store the branch (the survivor) and its metric for each state.
- Step 2. Increase t by 1, compute the partial metric for each path entering a state by adding the branch metric entering that state to the metric of the connecting survivor at the preceding trellis depth. For each state, store the path with the largest metric (the survivor) along with its metric, and eliminate all other paths.
- Step 3. Repeat Step 2 until the code sequence is decoded.

2.3.5 Turbo Codes

It is well known that using a serial concatenated two levels of coding, an inner and outer code linked by an interleaver, can achieve a low error rate and dramatically decrease the overall decoding complexity compared with that of a single code of the corresponding performance [48]. The low complexity is attained by decoding each component code separately. In decoding these concatenated codes, the inner decoder may use a soft-input soft-output (SISO) decoding algorithm to produce soft decisions for the outer decoder.

Turbo codes employ a similar idea of concatenating two codes and separating them by a random interleaver [6]. Rather than using a serial concatenation, two identical recursive systematic convolutional (RSC) codes in turbo codes are connected in parallel. A block diagram of a rate $1/3$ turbo encoder is shown in Fig.2.3 [45]. In the encoder, the same information sequence is encoded twice but in a different order. The first output sequence v_0 is equal to the input sequence c since the encoder is systematic; the first RSC encoder also produces the second parity sequence output, denoted by v_1. The third output sequence v_2 is produced by the second RSC encoder with an interleaved version of the input sequence, $\tilde{c}$, as the input.

Turbo and serial concatenated codes can be decoded by using a maximum a posteriori probability (MAP) algorithm and a soft output Viterbi algorithm. The MAP uses a decoding criterion that minimizes the symbol or bit error probability, whereas the SOVA minimizes the sequence error probability to generate soft output information [45].

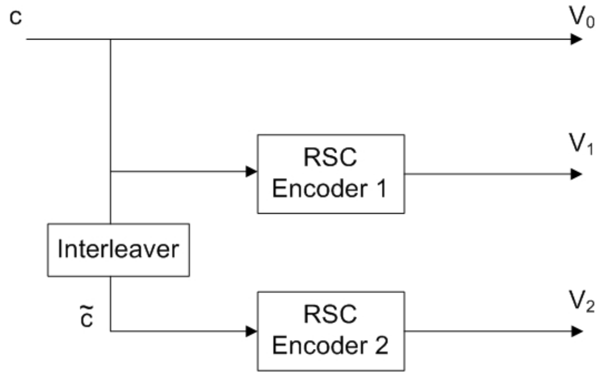

Figure 2.3: A rate 1/3 turbo encoder

2.3.6 Soft Output Viterbi Algorithm

The soft output Viterbi algorithm produces soft output to overcome the drawback of the VA, which generates hard symbol estimates [49].

The SOVA estimates the soft output information for each transmitted binary symbol in the form of the log-likelihood function $\Lambda(c_t)$, as follows:

$$\Lambda(c_t) = log\frac{P_r\{c_t = 1|r_1^\tau\}}{P_r\{c_t = 0|r_1^\tau\}}, \tag{2.16}$$

where r_1^τ is the received sequence and $P_r\{c_t = i|r_1^\tau\}$, i = 0, 1, is the a posteriori probability (APP) of the transmitted symbol, which is given by [45]

$$\begin{aligned} P_r\{c_t = 1|r_1^\tau\} &= \frac{e^{\Lambda_t}}{1+e^{\Lambda_t}} \\ P_r\{c_t = 0|r_1^\tau\} &= \frac{1}{1+e^{\Lambda_t}}. \end{aligned} \tag{2.17}$$

The SOVA decoder makes a hard decision by comparing $\Lambda(c_t)$ to a threshold value which is equal to zero

$$c_t = \begin{cases} 1 & \text{if}\Lambda(c_t) \geq 0 \\ 0 & \text{otherwise.} \end{cases} \tag{2.18}$$

The decoder selects the path $\hat{x}$ with the minimum path metric $\mu_{\tau,min}$ as the ML path in the same way as the VA. The probability of selecting this path is proportional to

$$P_r(c|r_1^\tau) = P_r(x|r_1^\tau) \sim e^{-\mu_\tau,min}. \tag{2.19}$$

Let us denote by $\mu_{t,c}$ the minimum path metric of the paths with the complementary symbol to the ML symbol at time t. If the ML symbol at time t is 1, then its complementary symbol is 0. Therefore, we can write

$$\begin{aligned} P_r(c_t = 1|r_1^\tau) &\sim e^{-\mu_\tau,min} \\ P_r(c_t = 0|r_1^\tau) &\sim e^{-\mu_{t,c}}. \end{aligned} \tag{2.20}$$

Let μ_t^1 represent the minimum path metric for all paths for which c_t is 1 and μ_t^0 the minimum path metric for all paths for which c_t is 0. There are two cases to be considered to calculate the logarithm ratio of each transmitted binary symbol.

1. If the ML estimate at time t is 1, its complementary symbol at time t is 0. Therefore, $\mu_t^1 = \mu_{\tau,min}$ and $\mu_t^0 = \mu_{t,c}$.
2. If the ML estimate at time t is 0, its complementary symbol at time t is 1, giving $\mu_t^1 = \mu_{t,c}$ and $\mu_t^0 = \mu_{\tau,min}$.

For both cases, the log-likelihood ratio (LLR) can be expressed as

$$log\frac{P_r\{c_t = 1|r_1^\tau\}}{P_r\{c_t = 0|r_1^\tau\}} \sim log\frac{e^{-\mu_{t,c}}}{e^{-\mu_{\tau,min}}} = \mu_{\tau,min} - \mu_{t,c} = \mu_t^0 - \mu_t^1, \tag{2.21}$$

$$\Lambda(c_t) \sim \mu_t^0 - \mu_t^1. \tag{2.22}$$

Then, the difference between the minimum path metric among all the paths with symbol 0 at time t and the minimum path metric among all the paths with symbol 1 at time t is the soft output of the decoder. The sign of $\Lambda(c_t)$ determines the hard estimate at time t and its absolute value represents the soft output information that can be used for decoding in the next stage.

2.3.7 Iterative SOVA Decoding of Turbo Codes

It is possible to use the SOVA algorithm to proceed with iterative decoding of turbo codes [50]. The iterative turbo decoding consists of two component decoders serially concatenated via an interleaver, identical to the one in the encoder. The block diagram of the iterative SOVA decoder is shown in Fig.2.4.

The first SOVA decoder takes as input the received information sequence r_0 and the received parity sequence generated by the first encoder r_1. The decoder then produces a soft output, which is interleaved and used to produce an improved estimate of the APPs of the information sequence, referred to as the extrinsic information (EI) for the second decoder.

The other two inputs to the second SOVA decoder are the interleaved received information sequence $\hat{r}_0$ and the received parity sequence produced by the second encoder r_2. The second SOVA decoder also produces a soft output which is used to improve the estimate of the APPs for the information sequence at the input of the first SOVA decoder in the next decoding operation.

The EIs for the first and second decoders can be respectively given by [45],

$$\begin{aligned} \Lambda_{2e}^{(r)}(c_t) &= \Lambda_2^{(r)}(c_t) - 4\tilde{r}_{t,0} - \tilde{\Lambda}_{1e}^{(r)}(c_t) \\ \Lambda_{1e}^{(r)}(c_t) &= \Lambda_1^{(r)}(c_t) - 4r_{t,0} - \tilde{\Lambda}_{2e}^{(r-1)}(c_t), \end{aligned} \tag{2.23}$$

where $\Lambda_1^{(r)}(c_t)$ and $\Lambda_2^{(r)}(c_t)$ are the LLRs of the first and second SOVA decoders; they can be calculated from Eqs.(2.21) and (2.22), respectively; $r_{t,0}$ and $\tilde{r}_{t,0}$ are the received information signal and interleaved version of that respectively; $\tilde{\Lambda}_{2e}^{(r-1)}(c_t)$ and $\tilde{\Lambda}_{1e}^{(r)}(c_t)$ are the interleaved version of the EIs for the first and second SOVA decoders respectively.

After a certain number of iterations, the decoders stop producing further performance improvements. At the last stage of decoding, the second decoder makes hard decisions based on Eq.(2.18).

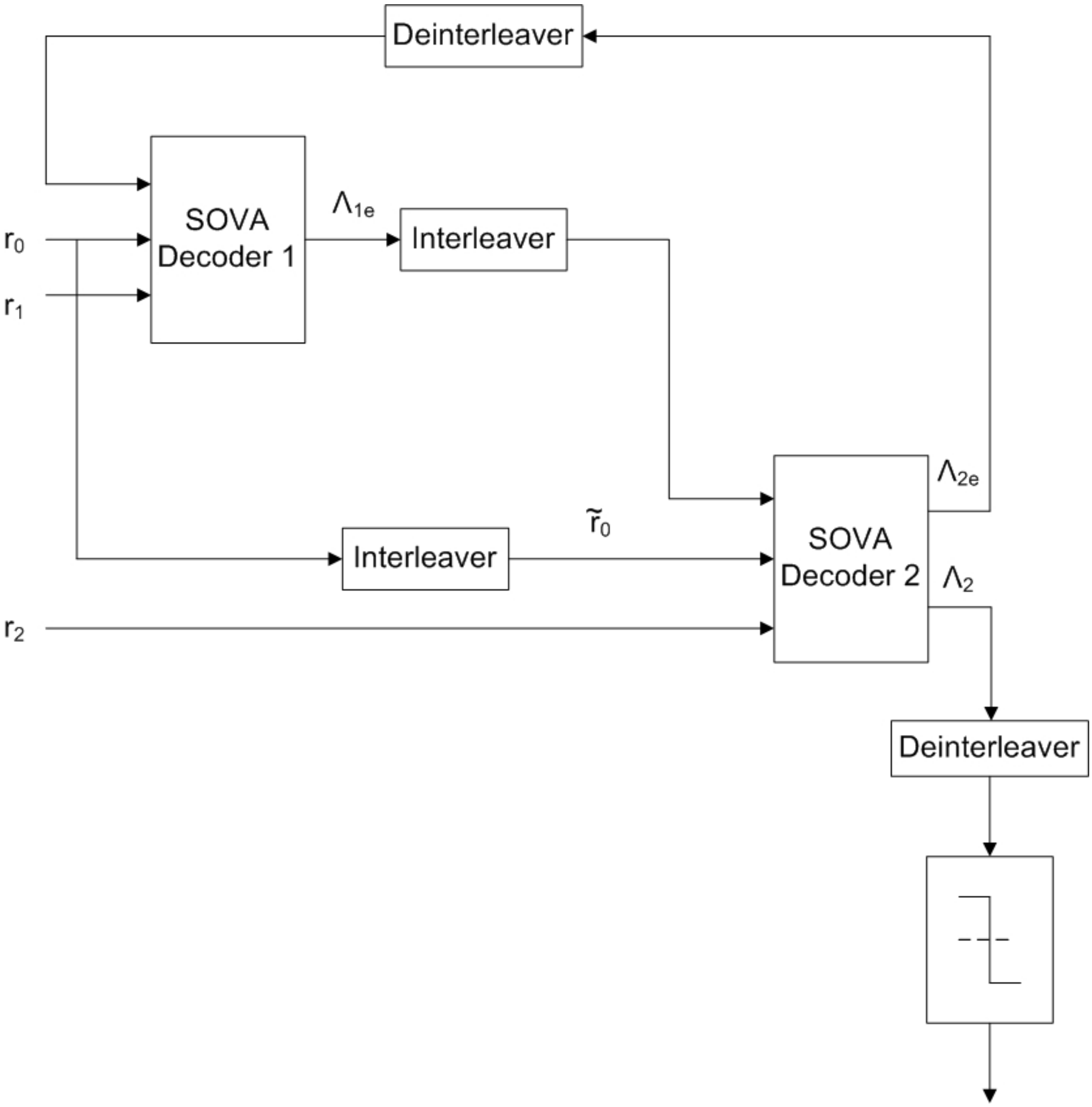

Figure 2.4: An iterative turbo code decoder based on the SOVA algorithm

2.4 Error Control Scheme - Automatic Repeat Request

In the previous section, we discussed some forms of FEC. This section is devoted to three standard versions of the ARQ scheme, which is referred to as an error detection strategy:

- Stop-and-wait (SW) ARQ
- Go-back-N (GBN) ARQ
- Selective-repeat (SR) ARQ

2.4.1 Stop-and-wait ARQ

The SW scheme represents the simplest ARQ procedure. In an SW ARQ error-control system, the transmitter sends a codeword to the receiver and waits for an acknowledgement, as shown in Fig.2.5 [8]. An ACK from the receiver indicates that the transmitted codeword has been successfully received, and the transmitter can send the next codeword. A NAK from the receiver indicates that the transmitted codeword has been detected in error, the transmitter needs to resend the previous codeword and waits for an acknowledgement. Retransmission continues until the transmitter receives an ACK or the maximum retransmission number is reached.

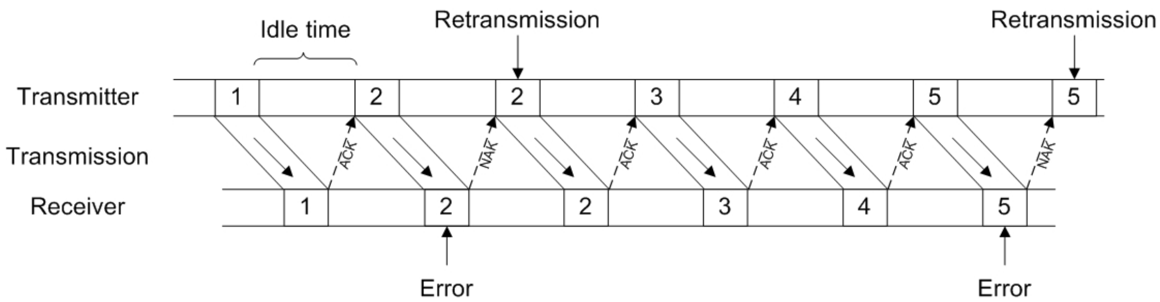

Figure 2.5: Stop-and-wait ARQ

The principal advantage of SW ARQ is its simplicity. But the inefficiency is its inevitable disadvantage because of the idle time spent waiting for an acknowledgement of each transmitted codeword. With the high data rates and utilization of satellite channels of long round-trip delays, the continuous ARQ is established to replace the SW procedure.

2.4.2 Go-back-N ARQ

In a go-back-N ARQ system [8], codewords are transmitted continuously. The transmitter keeps sending a new codeword, as soon as it has completed sending one, without waiting for an acknowledgment, as illustrated in Fig.2.6. The acknowledgment for a codeword arrives after a round-trip delay, which is defined as the time interval between the transmission of a codeword and the receipt of an acknowledgment for that codeword. The other $N-1$ codewords can be transmitted during this interval. The transmitter goes back to the codeword, which is negatively acknowledged, and resends that specified codeword, say codeword i, and subsequent $N-1$ codewords that were transmitted during the round-trip delay. At the receiver, the receiver discards the erroneously received codeword and the succeeding $N-1$ codewords, regardless of whether they are error-free or not. Retransmission continues until the codeword i is positively acknowledged, and then the transmitter proceeds to transmit new codewords.

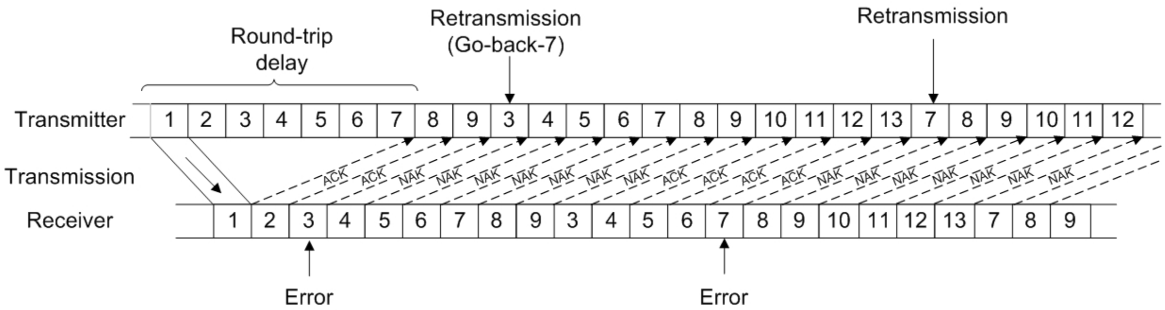

Figure 2.6: Go-back-N ARQ with N = 7

Because of the continuous transmission and retransmission of codewords, the go-back-N ARQ scheme is more effective than the SW ARQ. However, it becomes ineffective when the round-trip delay is large and the data transmission rate is high. This inefficiency is caused by the retransmission of next $N-1$ received codewords, even though many of them may be error-free. This defect can be overcome by using selective-repeat ARQ.

2.4.3 Selective-repeat ARQ

In a selective repeat ARQ system, codewords are also transmitted continuously, however, the transmitter only resends those codewords that are negatively acknowledged, as shown in Fig.2.7 [8]. Since the user needs to receive the codewords in the right order, a buffer has to be provided at the receiver to store the error-free received codewords. When the repeated transmitted codeword is successively received, the receiver can then release the error-free codewords in consecutive order.

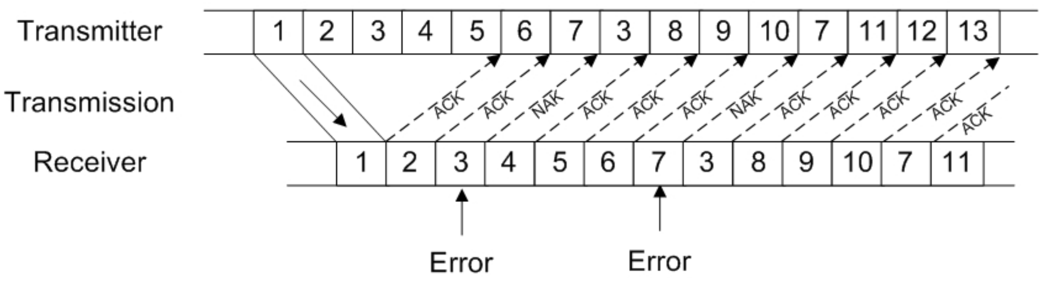

Figure 2.7: Selective-repeat ARQ

The selective-repeat ARQ is the most efficient scheme among the three basic ARQ schemes, but it is also the most complex one to implement.

2.5 Error Control Scheme - Hybrid Automatic Repeat Request

The two categories of techniques for controlling transmission errors in data communication systems have disadvantages. For example, in an FEC communication system,

- The received codeword needs to be decoded even it has been detected in error, and the decoded message has to be delivered to the user, regardless of whether it is correct or not.

- To obtain high system reliability, a long powerful code must be used and a large collection of error patterns must be corrected. This makes decoding difficult to implement and expensive [9].

In contrast, in an ARQ communication system, the throughput is not constant and it falls rapidly with increasing channel error rate, which is the primary weakness of the ARQ scheme.

Drawbacks of FEC and ARQ schemes can be overcome if the two basic error control schemes are properly combined. A HARQ scheme [9, 10, 51–53] is the approach to an error control strategy, which incorporates both FEC and ARQ. The HARQ system consists of a FEC subsystem contained in an ARQ system. By correcting the most frequently occurring error patterns, the FEC system can reduce the frequency of retransmission, increasing the system performance. However, when a less-frequent error pattern occurs and is detected, the receiver requests a retransmission rather than passing the unreliably decoded message to the user. This increases the system reliability. As a result, an appropriate combination of FEC and ARQ can offer better system performance.

The HARQ scheme can be classified into three categories, namely the type-I HARQ (repetition coding) scheme, which includes a pure type-I HARQ scheme and type-I HARQ scheme with Chase combining [32], the type-II HARQ (increment redundancy) scheme and type-III HARQ (a modified HARQII) scheme.

2.5.1 Type-I HARQ Scheme

The design of type-I HARQ scheme consists of selecting a fixed code with a certain rate and correction capability matched to the protection requirement of all the data to be transmitted and adapted to the average or worst channel conditions to be expected [54]. With or without using the buffer to store the previously received erroneous codeword, the type-I HARQ scheme can be divided into two subtypes, the pure type-I HARQ scheme and type-I HARQ scheme with Chase combining.

2.5.1.1 Pure Type-I HARQ Scheme

A pure type-I HARQ scheme uses a code which is designed for simultaneous error correction and error detection [8]. When a received codeword is detected in error, the receiver first attempts to correct the errors. As long as the number of errors are within the designed error-correcting capability of the code, the errors will be corrected and the decoded message will be delivered to the user. If the receiver detects an uncorrectable error pattern, it rejects the received codeword and asks for a retransmission. When the retransmitted same codeword is received, the receiver attempts to correct the errors (if any) again. The procedure will continue until the codeword is either successfully received or correctly decoded.

2.5.1.2 Type-I HARQ Scheme with Chase Combining

In 1977, Sindhu [55] discussed a scheme to use all the available packets at the receiver. The basic idea behind this technique is that a received packet which is determined to contain errors should not be discarded, because it contains useful information about the transmitted packet. If the transmission of this packet is also in error, the collective information present in the first packet and the retransmission can be used to correct certain errors in the two packets. Therefore, a single combined packet is more reliable than any of its constituent packets. Chase [32] further developed a practical and effective combining approach, known as Chase combining, for overcoming the problem of obtaining reliable communications when the actual channel capacity is unknown. When combining, packets should be weighted according to their relative reliability. Such a combining strategy can operate in a very high-error environment to achieve error-free results for all channels with finite capacity. These features were proved to be useful for a wide range of applications, as indicated in other literature [13, 29, 30, 56, 57].

2.5.2 Type-II HARQ Scheme

The type-I HARQ scheme is best suited for the communication systems wherein a fairly constant level of noise and interference is anticipated on the channel [8]. In this scenario, the error symbols of the received codewords can be corrected by the designed error-correcting code, thereby greatly reducing the number of retransmissions and enhancing the system performance. However, for a non-stationary channel in which the bit error rate changes, the type-I HARQ scheme has some shortcomings:

- When the channel condition is good, which means the channel BER is low, the transmission is smooth and no error correction is needed. Therefore, the extra designed error-correcting code is wasted during each transmission.
- When the channel is very noisy, the error-correcting capability may become inadequate. As a result, the frequency of retransmission increases and thus reduces the throughput.

The type-II HARQ scheme is proposed to overcome such drawbacks of the type-I HARQ scheme. The basic idea is to design an adaptive HARQ scheme. When the channel is quiet, the transmitted codeword includes the information symbols, none or a few parity symbols in each transmission, and the system behaves like the ARQ system. However, if the channel becomes noisy, as long as the receiver detects the errors in the received codeword, it saves the erroneously received codeword in the buffer, and at the same time requests a retransmission. During the retransmission procedure, the IR codes are transmitted to the receiver until a powerful enough codeword is formed to achieve error-free decoding. This scheme is indicated in Fig.2.8.

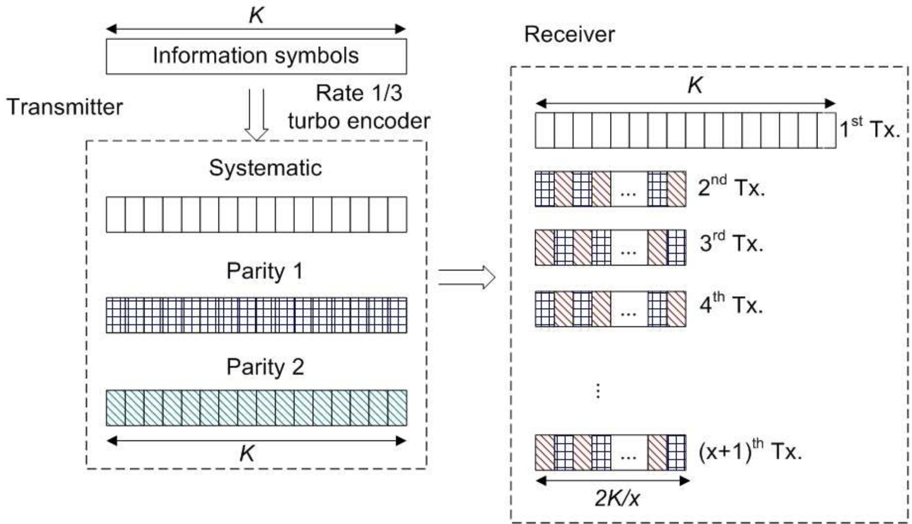

Figure 2.8: Type-II HARQ scheme

The type-II HARQ scheme can be incorporated with a rate $1/2$ convolutional code using Viterbi decoding [58–60]. Rate $1/N$ convolutional codes are punctured periodically with period P to obtain a family of codes with rate $P/(P + l)$, where l can be varied between 1 and $(N - 1)P$ [61, 62]. In [11], a concept of RCPC codes was proposed, where a rate compatibility restriction implies that all the code symbols of a high rate punctured code are used by lower rate codes. Since codes are compatible, rate variation within a data frame is possible to achieve unequal error protection (UEP). Fig.2.9 depicts an example of an RCPC-ARQ system.

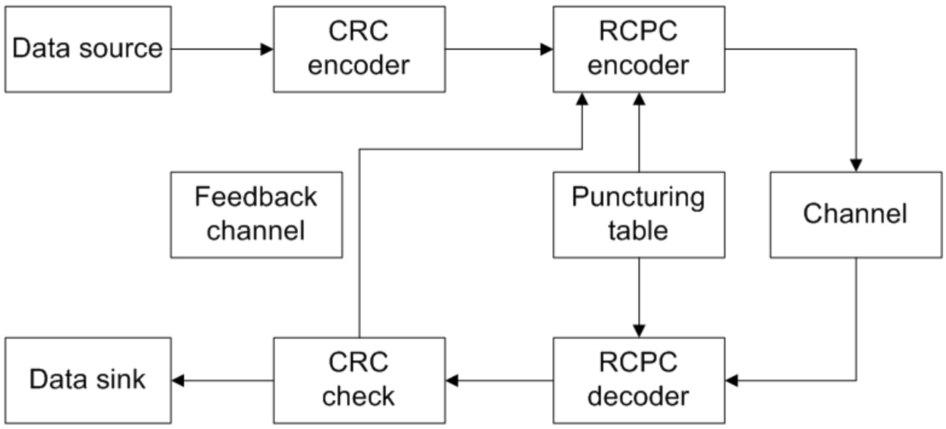

Figure 2.9: An example of an RCPC-ARQ system

An extension of the RCPC concept is applied to the HARQ system employing rate-compatible punctured turbo (RCPT) codes. The resulting system provides a powerful low-rate error correction capability at low SNR and outperforms the RCPC codes for sufficiently large block size [13].

2.5.3 Type-III HARQ Scheme

The main drawback of the type-II HARQ scheme is that additional incremental code symbols sent for a packet received with errors are not in general self-decodable [63]. The decoder has to rely on both the previous received packets as well as the current additional incremental code symbols for decoding, in situations wherein a current transmitted IR packet can be lost or severely damaged as a result of interference. Kallel [63] modified the type-II HARQ scheme by exploiting the complementary punctured convolutional (CPC) codes. Briefly, the CPC codes are composed of a set of equivalent codes with the same rate and distance spectrum from the same original mother code. Due to its self-decodable property, the type-III HARQ procedure has the choice of extracting the source information either from the last received packet or by combining all previous packets, as is generally the case with the type-II HARQ scheme.

Chapter 3

Introduction to Cooperative Diversity

In wireless communication systems, significant attenuation and fading limit the channel capacity of a specific communication link or an entire network, which could hamper the ability of a wireless network to deliver services with the same level of quality as that guaranteed by a wired system. This challenge is compounded by growing demands from services that require high data-rate transmission. One of the primary solutions is to create some form of diversity by transmitting and processing redundant signals over essentially independent channels. The multiple-input multiple-output and cooperative diversity techniques are two well-known techniques to achieve transmission diversity.

Section 3.1 gives a brief review of the MIMO and CD techniques. A variety of CD protocols, including fixed, selection, and incremental relaying, are described in section 3.2. Section 3.3 introduces two main coded cooperation schemes, the distributed rate compatible convolutional codes, and distributed turbo coding schemes, which are widely used in the CD communication systems.

3.1 Introduction

3.1.1 Multiple-input Multiple-output Technique

Multiple-input multiple-output communication techniques have been an important area of focus for the next-generation wireless systems because of their potential for high capacity, increased diversity, and interference suppression [18].

A MIMO channel with n_T transmitters and n_R receivers is typically represented as a matrix $\mathbf{H}$ of dimension $n_R \times n_T$, where each of the coefficients $[\mathbf{H}]_{i,j}$ represents the transfer function from the j^{th} transmitter to the i^{th} receiver, as shown below:

$$\mathbf{H} = \begin{bmatrix} h_{1,1} & h_{1,2} & \cdots & h_{1,nT} \\ h_{2,1} & h_{2,2} & \cdots & h_{2,nT} \\ \cdots & & & \\ h_{nR,1} & h_{nR,2} & \cdots & h_{nR,nT} \end{bmatrix}. \tag{3.1}$$

We denote the signal or symbol transmitted from the j^{th} transmitter as x_j, and collect all such symbols into an n_T-dimensional vector $\mathbf{x}$. With this notation, the matrix model of the channel is

$$\mathbf{y} = \mathbf{H}\mathbf{x} + \mathbf{n}, \tag{3.2}$$

where **n** is a vector of additive noise, and **y** is the vector of received data, with an element in **n** and **y** for each receive antenna.

This system, as shown in Fig.3.1, is implemented with an antenna array at both the transmitter and receiver in conjunction with using sophisticated coding, aiming to create separated transmission paths that experience independent fades. The MIMO channel can be provisioned for higher data rates, resistance to multipath fading, lower delays, and support for multiple users.

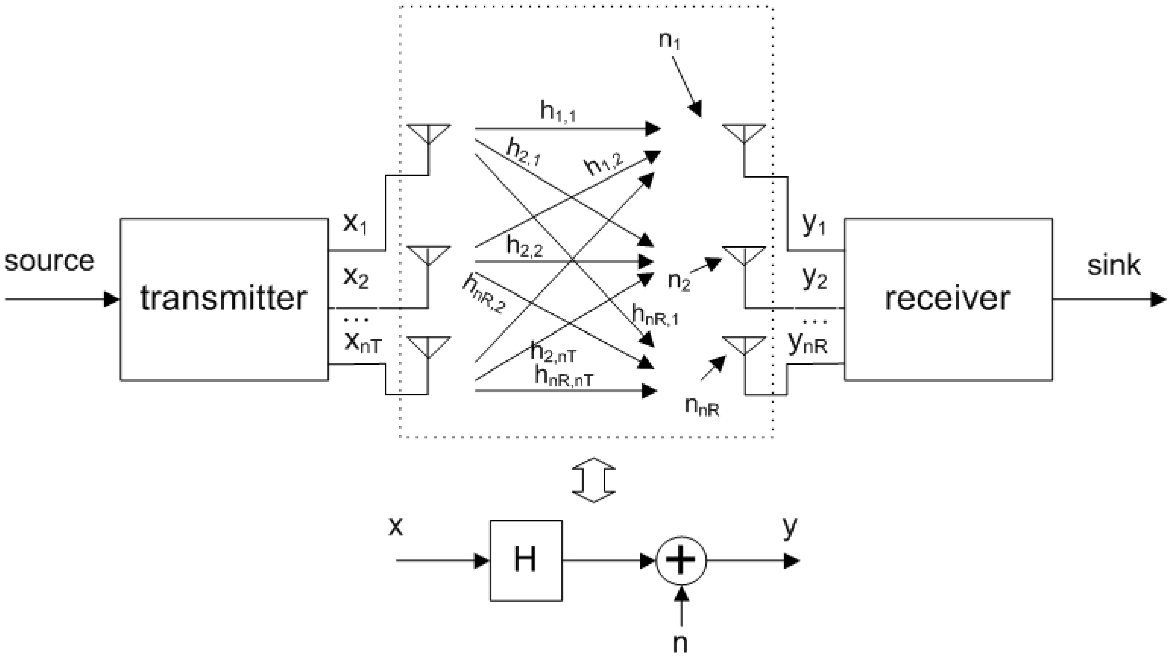

Figure 3.1: MIMO wireless link

3.1.2 Cooperative Diversity Technique

The MIMO technique is a successful method that can provide both spatial and temporal diversities to effectively mitigate the detrimental effects of fading for point-to-point channels [16, 17, 64, 65].

However, in reality, due to size, cost, or hardware limitations, the antenna array can hardly be implemented on a small platform such as a mobile handset (size) or wireless sensor nodes (size, power). Recently, a new form of spatial-temporal diversity has been created, namely cooperative diversity, which allows single-antenna mobiles to reap some of the benefits of MIMO systems [19–21]. The essence of the CD technique is that single-antenna mobiles in a multi-user scenario can “share” their antennas in a manner that creates a virtual MIMO system. Based on this virtual array, the devices are able to share their distributed resources such as bandwidth and power by relaying each other’s data in a cooperative way. Fig.3.2 shows the basic idea behind this concept, where by two users cooperate with each other to communicate with a remote base station. Since each user has an independent fading path to the destination, spatial diversity is generated by transmitting signals from different locations. When the direct path between the source and destination is in deep fade or blocked by an obstacle, due to the broadcast nature of the wireless medium, the relay likewise experiences better receive conditions than the destination and thus can offer the “overheard” information to the destination. Therefore, the combined signals at the destination can increase the overall received SNR.

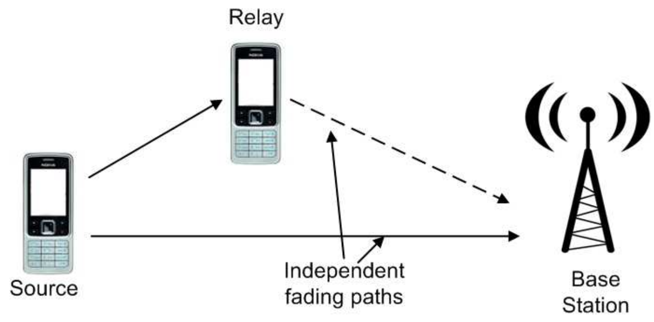

Figure 3.2: A model of a cooperative communication technique

Since the cooperating partners in the CD communication systems are geographically separated, the spacing between any pair of users, and hence their antennas, is wider than that of a conventional MIMO system. Therefore, the CD technique is immunized not only against small-scale channel fading but against large-scale channel fading.

3.2 Cooperative Diversity Protocols

In this section, we describe a variety of cooperative diversity protocols that can be utilized in the network of Fig.3.2, including fixed, selection, and incremental relaying protocols.

3.2.1 Fixed Relaying Protocols

For the fixed relaying, the relay either amplifies its received signals subject to its power constraint, or decodes, re-encodes and retransmits the message to the destination. The description for each of them is summarized as follows.

In a CD communication system, the source and relay transmit data through orthogonal channels [19, 20, 25, 39, 66, 67], which could be implemented through time, frequency or code-division multiple access. In a time division duplexing (TDD) scheme, time is slotted and the CD communication process needs two time slots. In the first time slot, the source first broadcasts its message to both the relay and destination. The received signals at the relay and destination, at time t, denoted by $y_{sr}(t)$ and $y_{sd}(t)$ can be expressed as

$$y_{sr}(t) = \sqrt{P_{sr}} h_{sr} x(t) + n_{sr}(t), \tag{3.3}$$

$$y_{sd}(t) = \sqrt{P_{sd}} h_{sd} x(t) + n_{sd}(t), \tag{3.4}$$

where P_{sr}, P_{sd} are the received signal power at the relay and destination respectively.

Upon receiving $y_{sr}(t)$, during the next time slot, the relay processes the received signal and sends it to the destination. Let $x_r(t)$ represent the signal transmitted from the relay at time t. It satisfies the following transmit power constraint,

$$E(|x_r(t)|^2) \leq P_r, \tag{3.5}$$

where P_r is the transmitted power limit at the relay.

For the second time slot, the destination receives the signal as

$$y_{rd}(t) = G_{rd} h_{rd} x_r(t) + n_{rd}(t). \tag{3.6}$$

3.2.1.1 Amplify and Forward

For amplify and forward transmission, the appropriate channel model is Eq.(3.3)-Eq.(3.6). Each user in this method receives a noisy version of the signal transmitted by its partner. The user then amplifies the received signal and then forwards it to the destination. The signal transmitted from the relay can be written as

$$x_r(t) = \mu y_{sr}(t), \tag{3.7}$$

where μ is an amplification factor that depends on the transmit signal power, noise spectral power density and fading attenuation. It can be calculated from Eqs.(3.3) and (3.5) as

$$\mu \leq \sqrt{\frac{P_r}{|h_{sr}|^2 P_{sr} + N_0}}. \tag{3.8}$$

The destination combines the signals sent by the source and relay as

$$\begin{aligned} & w_{sd} y_{sd}(t) + w_{rd} y_{rd}(t) \\ = \quad & w_{sd}\left[h_{sd} s(t) + n_{sd}(t)\right] + w_{rd}\left\{h_{rd}\mu\left[h_{sr} s(t) + n_{sr}(t)\right] + n_{rd}(t)\right\}, \end{aligned} \tag{3.9}$$

where the combining coefficients can be calculated as [68]

$$w_{sd} = \frac{\sqrt{P_{sd}} h_{sd}^*}{N_0}, \qquad w_{rd} = \frac{\mu\sqrt{P_{sr}} h_{rd}^* h_{sr}^*}{(\mu^2|h_{rd}|^2 + 1)N_0}. \tag{3.10}$$

Although noise is amplified by cooperation, the base station receives two independently faded versions of the signal and can make a better decision on the detection of information.

3.2.1.2 Decode and Forward

For decode and forward transmission, the appropriate channel is again Eq.(3.3)-Eq.(3.6). The relay decodes the received signal, then re-encodes and forwards it with power P_r to the destination.

$$x_r(t) = \sqrt{P_r} x(t), \tag{3.11}$$

The corresponding received signal at the destination can be expressed as

$$y_{rd}(t) = G_{rd} h_{rd} \sqrt{P_r} x(t) + n_{rd}(t). \tag{3.12}$$

3.2.2 Selection Relaying Protocol

In the fixed DAF protocol, the relay always forwards the decoded information to the destination. This protocol suffers performance loss if the relay cannot decode the transmitter's signal correctly. To overcome the drawbacks of the fixed relaying transmission, the author in [21] also proposed adaptive relaying protocols, including selection relaying and incremental relaying protocols.

The signal transmission in the selection relaying protocol still involves two time slots. In the first time slot, each source sends information to its destination, and the information is also received by other users in the network. In the second time slot, the relay decodes the received information and forwards the decoded information symbols only if the amplitude of the measured channel coefficient of source to the relay channel is larger than a certain threshold, otherwise, the relay does not send and remains idle [21].

The wireless link in the network is subject to fading and additive noise. Although the fading coefficients can be well estimated by the cooperation terminal, the additive noise is unknown to the receiver. To take into account both the effect of channel fading and the effect of additive noise, the author in [69] proposed an optimum threshold selection relaying scheme, which adapts based on the received signal power at the relay terminal. The relay decides whether, or not to forward the signal it receives from the source, by comparing the squared amplitude of the received signal, normalized by the average noise power, with a certain threshold.

3.2.3 Incremental Relaying Protocol

Since the relay repeats all the time, the fixed and selection relaying make inefficient use of the degrees of freedom of the channel. Using a single bit indicating the success or failure of the direct transmission, from the source to the destination, the relay retransmits in an attempt to exploit spatial diversity.

As an example, the source broadcasts its information to the relay and destination. After decoding the received signal, the destination indicates the received signal status (either correct or erroneous) by broadcasting a single bit of feedback to the source and relay. If the SNR between the source and destination channel is sufficiently high, the feedback indicates the correctly received signal from the source, and the relay remains idle. Otherwise, based on the received feedback from the destination, the relay needs to forward its received signal to the destination using either the AAF or DAF transmission protocols.

This incremental relaying protocol can be treated as extensions of incremental redundancy, or HARQ, to the relay context. Since the relay rarely repeats, the protocol of this form makes more efficient use of the degrees of freedom of the channel.

3.3 Coded Cooperation Schemes

Coded cooperation [23] is a method that integrates cooperation into channel-coding. Coded cooperation works by transmitting different portions of each user's codeword via two independent fading paths. The idea is that each user tries to send incremental redundancy to its partner. Whenever the case is impossible, the users automatically revert to a noncooperative mode. Since the coded cooperation is managed automatically through code design, there is no feedback involved between the users, thus allowing coded cooperation to achieve efficiency.

3.3.1 Distributed Rate Compatible Convolutional Codes

In the distributed rate compatible convolutional codes scheme (also termed as user cooperative coding [25]), each user encodes blocks of K source bits using a concatenation of a CRC code followed by a prescribed code from a family of RCPC codes [23]. The length of coded bits per block is N, so the overall code rate is $R = K/N$. The N coded bits are divided into two segments, with the first segment being a punctured rate R_1 codeword with $N_1 = K/R_1$, and the second being the N_2, which represents the remaining parity bits for the rate R codeword, where $N_1 + N_2 = N$.

The users cooperate by dividing the transmission of their N-bit code words into two frames. In the first frame, N_1 is broadcasted by each user, and is received by the base station as well as the partner. Each user thus receives a noisy version of the coded packet from its partner. If the partner's packet is successfully decoded based on the CRC check, the user calculates and transmits the partner's N_2 remaining parity bits in the second frame. Otherwise, the user's own parity bits are transmitted. The framework of this scheme is illustrated as in Fig.3.3 [70].

The users act independently in the second frame, with no knowledge of whether their own first frame was correctly decoded. As a result, there are four possible cooperative scenarios

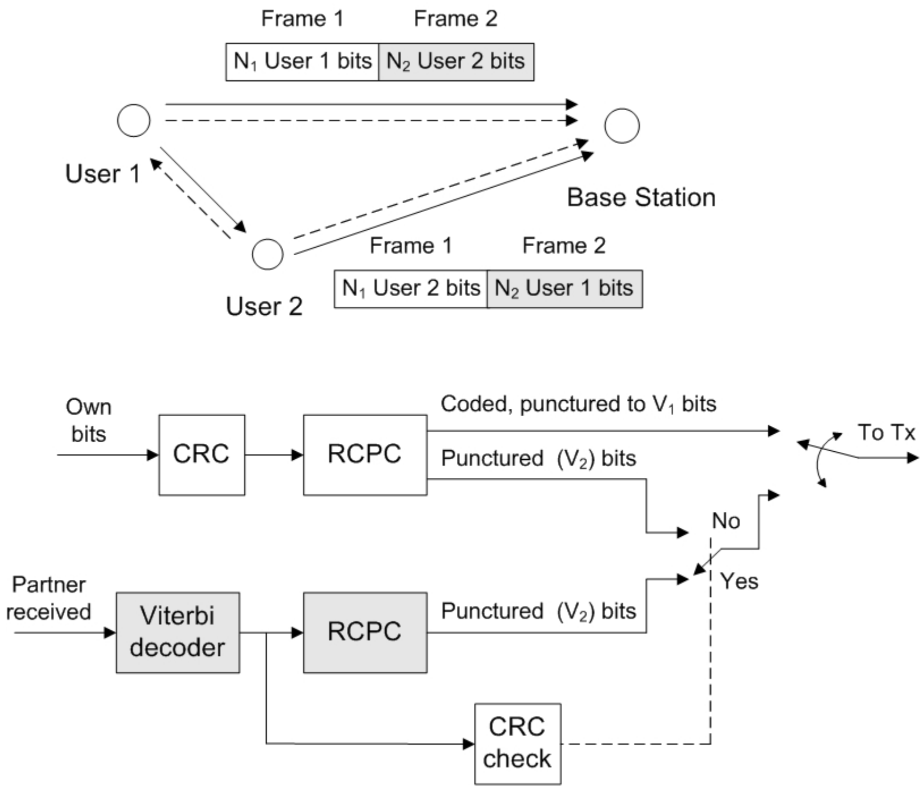

Figure 3.3: Block diagram of the distributed rate compatible puncture convolutional codes

for the second frame transmission: both users cooperate, neither user is cooperative, user 1 is cooperative but user 2 is not, and vice versa. A further comprehensive description and analysis can be referred to in the literature [71, 72].

3.3.2 Distributed Turbo Codes

Since cooperative coding contains two code components, it is natural to apply turbo codes [24]. For the sake of brevity, we omit the duplex transmission between the users, and only consider a half-duplex transmission channel (from user 1 to user 2) for this scheme's description. Recall that with turbo codes, the data is recursively encoded twice, first in its natural order and again after being interleaved [73]. In a DTC scheme, each user employs a very simple code, for example, a two-state rate $1/2$ RSC code, and an interleaver is added to user 2. Upon successful decoding of the partner (user 1), the user 2 interleaves user 1's source symbols and re-encodes them. Thus, the uninterleaved encoding is present in the user 1 to base station channel, while the interleaved encoding is present in the user 2 to base station channel. Then the users have cooperatively constructed a distributed turbo code [25, 39], the base station can detect the code iteratively by using a standard turbo decoder. The implementation of coded cooperation using distributed turbo codes is shown in Fig.3.4.

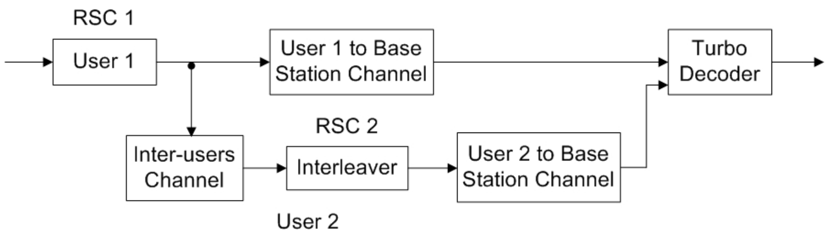

Figure 3.4: Block diagram of the distributed turbo coding technique

Chapter 4

Collaborative HARQ Schemes in Wireless Networks

4.1 Introduction

In conventional wireless networks, traditional multihop protocols were treated as a cascade of point-to-point links, with each radio directing its transmission to only a single receiver [74]. The quality of the signal received by the destination depends on various factors, for example, path loss, fading and noise. To guarantee reliable transmission over a radio channel, ARQ or HARQ protocols are usually adopted. The retransmission protocol specifies how data packets that are not correctly received and detected by the destination must be retransmitted from the source until they are successfully delivered. As other neighboring nodes do not take part in the packet retransmission process, these ARQ/HARQ protocols are referred to as non-collaborative.

However, one peculiar characteristic of the wireless networks is the inherent broadcast-oriented nature of radio. Beside the intended destination, a signal transmitted by a source may be also received by other neighboring nodes that are within earshot. In contrast with non-collaborative ARQ/HARQ, retransmitted packets do not need to come from the original source but could instead be sent by relays that overhear the transmission [26]. Therefore, the spatial diversity is provided by the relays and the protocol is referred to as collaborative HARQ.

As a consequence, when the relay retransmits, various relaying protocols can be utilized in cooperative wireless networks, including fixed, selection, incremental [21], compress and forward [75] protocols etc. Compared with other relaying protocols, AAF and DAF are two popular protocols and are frequently used in the cooperative systems [15, 76–78].

In an AAF scheme, each user receives a noise version of the signal transmitted by its partner. Then the parter amplifies and retransmits this noise version. The destination combines the packets sent by the user and partner, and makes a final decision on the transmitted packet.

In a DAF scheme, there are two possible policies to achieve transmission at the relay:

- Selection Transmission: The relay only retransmits if it decodes the received packet from the source correctly.
- Persistent Transmission: The relay always retransmits, even if the decoded packet is in error, which leads to severe error propagation.

Preliminary works on the HARQ schemes have focused on the AAF and DAF relaying protocols. In the HARQ I-DAF scheme, upon the reception of a negative acknowledgement (NAK), the relay retransmits a copy of the original packet, received from the source to the destination [27]. In the HARQ II-DAF scheme, the source broadcasts odd-numbered symbols of the codeword, the relay re-encodes the source message to obtain and transmit even-numbered symbols of the codeword to the destination when it decodes correctly [28]. In [29], HARQ I, HARQ II schemes with different cooperative strategies, including DAF and AAF protocols, have been further investigated.

Obviously, in wireless networks, an AAF protocol suffers from noise amplification and a DAF protocol leads to error propagation if the decoding packet is in error at the relay.

In this chapter, a HARQ scheme based on an adaptive relaying protocol (ARP) is proposed. It combines retransmission mechanisms, distributed turbo coding and an adaptive relaying protocol. When the relay is required for retransmission, it transmits either a repeated packet or a punctured packet to the destination, depending on the types of HARQ strategies used in the system. The relay switches the relaying protocol between AAF and DAF, depending on the decoding result. If the decoding result at the relay is correct, the relay decodes the received signal, and then interleaves, re-encodes and forwards it to the destination. The signals received at the destination consist of a coded signal transmitted from the source and a coded interleaved information transmitted from the relay. These two signals form a distributed turbo code [25, 39]. Otherwise the relay amplifies and forwards the originally received signal to the destination.

Moreover, we propose an efficient feedback strategy. Rather than using two feedback channels as in other HARQ relaying schemes, the proposed HARQ-ARP scheme has only one broadcast feedback channel, from the destination to both the relay and source. According to the number of received NAKs for each packet, the source and relay determine which one of them will be used in the next time slot for retransmission. This can reduce the feedback load and increase the transmission efficiency.

In this chapter, we mainly focus on the description of the proposed HARQ-ARP protocol. The performance analysis for the proposed HARQ-ARP scheme is carried out in the next chapter.

4.2 System Model

In this thesis, we study a typical two-hop relay network consisting of a source, a relay and a destination, as shown in Fig. 3.2.

We consider a quasi-static fading channel, in which the fading coefficients are constant within one transmission block, but change independently from one frame to another. Because of slow fading, accurate channel estimation is possible at the receiver [21, 79]. Therefore, we assume perfect CSI at the relay and destination. The source, on the other hand, only knows the statistics of fading, but not the current realization. Let "direct channel", "inter-user channel" and "relay channel" denote the channels between the source and destination, the source and relay and the relay and destination, respectively. The fading coefficients of the above three channels are independent. Similar to [21, 80], we assume that the source and relay transmit orthogonal signals in a TDD scheme. Furthermore, we assume an errorless, low-capacity feedback channel over which the receiving destination can transmit an ACK or a NAK signal to the source and relay.

The transmitted information binary stream with an appended CRC coded sequence, denoted by $\mathbf{U}$, is represented by

$$\mathbf{U} = (u(1), \cdots, u(t), \cdots, u(l)), \tag{4.1}$$

where $u(t)$ is a binary symbol transmitted at time t and l is the frame length.

The binary information sequence $\mathbf{U}$ is first encoded by a channel encoder. For simplicity, we consider a recursive systematic convolutional code with a code rate of $1/2$ [6]. Let $\mathbf{V}$ represent the corresponding codeword, given by

$$\mathbf{V} = (\mathbf{V}(1), \cdots, \mathbf{V}(t), \cdots, \mathbf{V}(l)), \tag{4.2}$$

where $\mathbf{V}(t) = (u(t), v(t))$ is the transmitted codeword, $u(t)$ is the information symbol and $v(t)$ is the corresponding parity symbol, $u(t), v(t) \in \{0, 1\}$.

The binary symbol stream $\mathbf{V}$ is then mapped into a modulated signal stream. For binary phase shift keying (BPSK) modulation, the modulated codeword, denoted by $\mathbf{X}$, is given by

$$\mathbf{X} = (\mathbf{X}(1), \cdots, \mathbf{X}(l)), \tag{4.3}$$

where $\mathbf{X} = (x(t,1), x(t,2)), x(t,j) \in \{-1, +1\}, t = 1, \cdots, l, j = 1, 2$ is the modulated signal transmitted by the source at time $2(t-1)+j$.

We impose a half-duplex constraint, so the cooperative communication process needs two time slots. In the first time slot, the source first broadcasts its message to both relay and destination.

The received signal at the relay, denoted by $y_{sr}(t)$, can be described as

$$y_{sr}(t) = \sqrt{P_{sr}} h_{sr} x(t) + n_{sr}(t), \tag{4.4}$$

where $P_{sr} = P_s \cdot (G_{sr})^2$ is the received signal power at the relay; P_s is the source transmit power, $G_{sr} = \left(\frac{\lambda_c}{4\pi d_0}\right)\left(\frac{d_{sr}}{d_0}\right)^{-\kappa/2}$ [81] is the gain of the inter-user channel, d_{sr} is the distance between the source and relay, d_0 is a reference distance; λ_c is the carrier wavelength and κ is a path-loss factor with values typically in the range $1 \leq \kappa \leq 4$. h_{sr} is the fading coefficient of the inter-user channel. It is modeled as a zero-mean, independent circular symmetric complex Gaussian random variable. Furthermore, $n_{sr}(t)$ is a zero mean complex Gaussian random variable with the two-sided spectral density of $N_0/2$.

Upon receiving $y_{sr}(t)$, in the next time slot, the relay processes the received signal and sends it to the destination. Let $x_r(t)$ represent the signal transmitted from the relay at time t. It satisfies the following transmit power constraint,

$$E(|x_r(t)|^2) \leq P_r, \tag{4.5}$$

where P_r is the transmitted power limit at the relay.

In the proposed scheme, the relay decodes the received signal from the source, interleaves the decoded information symbols, re-encodes and then forwards it with power P_r to the destination. Let $\widetilde{\mathbf{U}} = (\widetilde{u}(1), \cdots, \widetilde{u}(t), \cdots, \widetilde{u}(l))$ represent the interleaved version of $\mathbf{U}$. Let $\widetilde{\mathbf{V}} = (\widetilde{V}(1), \cdots, \widetilde{V}(l))$ denote the codeword of $\widetilde{\mathbf{U}}$, where $\widetilde{\mathbf{V}}(t) = (\widetilde{u}(t), \widetilde{v}(t))$ is the codeword of $\widetilde{u}(t)$. $\widetilde{\mathbf{V}}$ is then mapped into a modulated signal stream $\widetilde{\mathbf{X}} = \widetilde{X}(1), \cdots, \widetilde{X}(l)$, where $\widetilde{\mathbf{X}}(t) = (\widetilde{x}(t,1), \widetilde{x}(t,2))$, $\widetilde{x}_t(j)$ is the modulated signal transmitted by the relay using DTC, at time $2(t-1)+j$.

$$x_r(t) = \sqrt{P_r}\widetilde{x}(t). \tag{4.6}$$

The corresponding received signals at the destination at time $2t-1$ and $2t$, transmitted from the source and relay, can be expressed as

$$y_{sd}(t) = \sqrt{P_{sd}}h_{sd}x(t) + n_{sd}(t), \tag{4.7}$$

$$y_{rd}(t) = G_{rd}h_{rd}x_r(t) + n_{rd}(t), \tag{4.8}$$

where $P_{sd} = P_s \cdot (G_{sd})^2$ is the received signal power at the destination, P_s is the source transmit power; $G_{sd} = \left(\frac{\lambda_c}{4\pi d_0}\right)\left(\frac{d_{sd}}{d_0}\right)^{-\kappa/2}$ [81] is the gain of the direct channel, d_{sd} is the distance between the source and destination, d_0 is a reference distance; λ_c is the carrier wavelength and κ is a path-loss factor with values typically in the range $1 \leq \kappa \leq 4$. G_{rd} is the channel gain of the relay channel. h_{sd} and h_{rd} are the fading coefficients of the direct and relay channel, respectively. They are modeled as zero-mean, independent circular symmetric complex Gaussian random variables. Furthermore, $n_{sd}(t)$ and $n_{rd}(t)$ are zero mean complex Gaussian random variables with the two-sided spectral density of $N_0/2$.

The overall received signals at the destination are obtained as the sum of the coded source signal, and the coded interleaved source signal transmitted from the relay. These two signals form a standard distributed turbo code, as shown in Fig. 3.4. The received signals' expressions are given in Eqs.(4.7) and (4.8).

4.3 Collaborative HARQ with the ARP Scheme

In this section, we present the proposed three types of HARQ schemes with ARP, denoted by HARQ I-ARP, HARQ II-ARP and HARQ III-ARP, where ARP represents a combination of DAF and AAF. Since the HARQ II-ARP and HARQ III-ARP protocols follow the same retransmission rules of HARQ I-ARP, we present the HARQ I-ARP scheme in detail, and only the differences between the type I HARQ-ARP and type II/III HARQ-ARP schemes are described in sections 4.3.2 and 4.3.3.

For the proposed HARQ-ARP scheme, when the relay needs to retransmit, it checks the CRC of the broadcasted signals from the source. If the decoding result is correct, the relay adopts DAF, otherwise, it employs AAF.

In the conventional HARQ relaying schemes, two acknowledgement feedback channels are used, one from the destination to the source and the other from the relay to the source [28]. In contrast, the proposed scheme only needs one feedback channel. The destination examines the CRC check to determine the decoding result and then generates an ACK or a NAK signal, broadcasting to both the relay and source.

The proposed ARP scheme has two attractive features in a practical application, for example,

- The relay can automatically adapt to the channel quality by simply switching between the AAF and DAF without any need for the CSI to be fed back from the destination to the relay or source. This feature is especially desirable in practical relay networks, particularly in a large multi-hop network, where the feedback of CSI for adaptation is impractical.

- Compared with the non-adaptive relaying protocols: AAF and DAF, the ARP processing at the relay and destination is the same as those protocols, so it does not add any decoding complexity.

Without loss of generality, we consider a 1/2 RSC code as a mother code, C_m, in all the HARQ-ARP schemes.

4.3.1 Type I HARQ with the ARP Scheme

In the HARQ I-ARP scheme, the transmitter retransmits the packet when the destination receiver detects errors in the received packet. At the destination receiver, the previous erroneous packets are saved and combined with the current received packet, based on the respective SNR [32]. The retransmission can either come from the source or relay.

The transmission procedure of the HARQ I-ARP protocol is shown in Fig. 4.1 and it consists of the following steps:

1. The source broadcasts a packet to the relay and destination and also stores the transmitted packet in the buffer.
2. The reception procedure at the destination:
   a. If the packet is correctly decoded, an ACK will be broadcasted to the source and relay. The transmitter goes to step 1 to send a new packet. The relay is inactive in this case.
   b. Otherwise, a NAK will be returned to both the source and relay. The source stays in an idle state and the relay decodes the packet.
3. Actions at the relay: based on the CRC check value, the relay makes a decision to use either the DAF or AAF relaying protocol.
   a. If the CRC check is correct, the relay uses a DAF protocol. It interleaves the estimated information symbols, re-encodes and forwards them to the destination.
   b. If the relay cannot decode correctly, the relay uses an AAF protocol. It amplifies the received signal and forwards it to the destination.

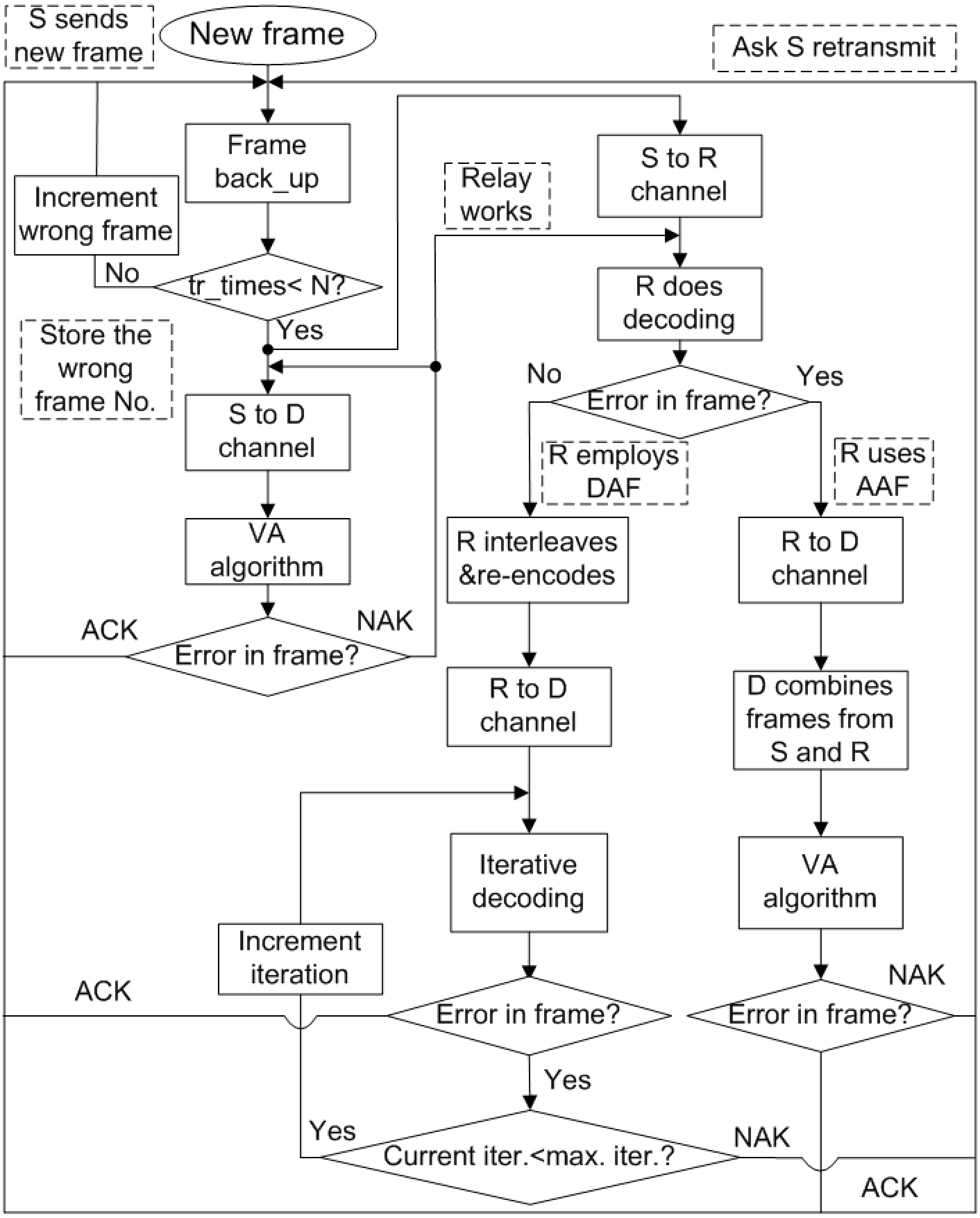

Figure 4.1: The proposed HARQ scheme with ARP where the relay uses either AAF or DAF

4. Actions at the destination:

    a. If the relay uses a DAF protocol, the destination receives two copies of the signal, the coded signal from the source and coded interleaved signal from the relay. Thus, a DTC codeword is formed. An iterative decoding method is used to recover the source information. Following are the operations, within the iterative decoding loop:

        * If the decoded packet is correct, exit the iterative decoding loop.
        * Otherwise, continue the decoding loop until the maximum number of iterations is reached. If the decoded packet is still erroneous, exit the iterative decoding loop.

    b. If the relay uses an AAF protocol, the destination combines the previous error packet from the source and current received packet from the relay.

    c. Based on the decoding result, an ACK/NAK will be broadcasted to the source and relay.

5. Actions at the source,

    a. If an ACK is received, the protocol goes to step 1 to send a new packet.

    b. Otherwise, if a NAK is received, a retransmission is required and the protocol proceeds to send the stored packet until the maximum retransmission number is reached. In that case, the source discards the packet and goes to step 1 to send another new packet.

From the above protocol procedure, we can see that the broadcast negative acknowledgement can determine whether the source or relay retransmits. Within the maximum number of transmissions, N, if the number of NAKs is odd, the source stays in an idle state and relay retransmits the received packet to the destination. If the number of NAKs is even, the source retransmits the original packet to the destination and relay waits for the further feedback from the destination. For example, at the first transmission, the source broadcasts a packet to the relay and destination, if the destination can not decode properly, the destination will send a NAK back to the source and relay. In this scenario, the received message at the relay will be transmitted to the destination and the source will stay in the idle state. At the destination, if the combined packets from the source and relay still can not be decoded

Transmission time	1^{st}	2^{nd}	3^{rd}
Information symbols	11111111	00000000	00000000
Parity symbols	01000010	10101000	00010101
Overall code rate	4/5	8/13	1/2

Table 4.1: Puncturing table for direct channel in a HARQ II-ARP scheme with a rate 4/5 code

correctly, the destination will broadcast the second NAK to the source and relay, then the source will retransmit the message to the destination and the relay will not transmit. So the source and relay alternatively transmit until the maximum retransmission number N is reached. The proposed broadcast feedback channel strategy can reduce the feedback load, increase the transmission efficiency and correspondingly decrease the transmission time.

4.3.2 Type II HARQ with the ARP Scheme

Unlike HARQ I-ARP, where the same data is repeated during each retransmission, HARQ II-ARP uses a high-rate error control code. The source transmits incremental redundancy symbols at each transmission step. According to the puncturing table, the relay and destination assemble previously received fragments of the packet to form a lower rate error control code to recover the information.

Two families of RCPT codes [13] with puncturing period $P = 8$ are used in the proposed HARQ II-ARP scheme, in which the optimal puncturing patterns are derived by using average distance spectra. Both code families employ a 4-state turbo code with generator polynomial $5/7$. To maximize capacity, the source and relay should transmit uncorrelated symbols, but related to the same message [25]. Therefore, the source and relay encode/re-encode the signal using different puncturing tables. The related puncturing matrices for the first to the third transmission are shown in Tables 4.1-4.4. A zero in the puncturing table means that the corresponding symbol of the packet is not transmitted.

Transmission time	1^{st}	2^{nd}	3^{rd}
Information symbols	11111111	00000000	00000000
Parity symbols	10000001	01010100	00101010
Overall code rate	4/5	8/13	1/2

Table 4.2: Puncturing table for relay channel in a HARQ II-ARP scheme with a rate 4/5 code

Transmission time	1^{st}	2^{nd}	3^{rd}
Systematic bits	11111111	00000000	00000000
Parity bit	11001010	00100001	00010100
Overall code rate	2/3	4/7	1/2

Table 4.3: Puncturing table for direct channel in a HARQ II-ARP scheme with a rate 2/3 code

4.3.3 Type III HARQ with the ARP Scheme

The type III HARQ with the ARP scheme follows the same retransmission rules of the type II HARQ with ARP scheme. The only difference is that the packet transmitted by the transmitter is self-decodable, so the relay and destination can use either the latest received packet or the combined received packets to recover the transmitted message [63]. In the proposed scheme, the relay and destination use the current received packets to decode. If the decoding result is not correct, the information symbols of all the received packets are combined by using Chase combining [32], while the parity symbols are combined by using code combining [29]. Since a set of punctured convolutional codes were derived from the same original low rate code with the same code rate and distance properties, when they are combined, the combined codes yield at least the mother code, C_m.

The corresponding puncturing tables for various code rates with the puncturing period $P = 8$ are shown in Tables 4.5-4.8.

Transmission time	1^{st}	2^{nd}	3^{rd}
Systematic bits	11111111	00000000	00000000
Parity bit	11010001	00100100	00001010
Overall code rate	2/3	4/7	1/2

Table 4.4: Puncturing table for relay channel in a HARQ II-ARP scheme with a rate 2/3 code

Transmission time	1^{st}	2^{nd}	3^{rd}
Information symbols	11111111	11111111	11111111
Parity symbols	01000010	10101000	00010101
Overall code rate	4/5	8/13	1/2

Table 4.5: Puncturing table for direct channel in a HARQ III-ARP scheme with a rate 4/5 code

Transmission time	1^{st}	2^{nd}	3^{rd}
Information symbols	11111111	11111111	11111111
Parity symbols	10000001	01010100	00101010
Overall code rate	4/5	8/13	1/2

Table 4.6: Puncturing table for relay channel in a HARQ III-ARP scheme with a rate 4/5 code

Transmission time	1^{st}	2^{nd}	3^{rd}
Systematic bits	11111111	11111111	11111111
Parity bit	11001010	00100001	00010100
Overall code rate	2/3	4/7	1/2

Table 4.7: Puncturing table for direct channel in a HARQ III-ARP scheme with a rate 2/3 code

Transmission time	1^{st}	2^{nd}	3^{rd}
Systematic bits	11111111	11111111	11111111
Parity bit	11010001	00100100	00001010
Overall code rate	2/3	4/7	1/2

Table 4.8: Puncturing table for relay channel in a HARQ III-ARP scheme with a rate 2/3 code

4.4 Simulation Results

In the previous section, we presented the proposed HARQ-ARP scheme for the CD communication systems in wireless networks. In this section, we illustrate the performance of the proposed scheme and compare it with another two reference schemes, HARQ with AAF (HARQ-AAF) and HARQ with DAF (HARQ-DAF).

In order to present equitable comparisons, we apply persistent transmission principles to all HARQ schemes, including HARQ-AAF, HARQ-ARP and HARQ-DAF schemes. The simulation conditions are summarized in Fig. 4.2, where the direct and relay channels have the same SNR, which varies from 0 to 14 dB, while the inter-user channel is variable from 0 to 24 dB.

Fig.4.3 shows the FER comparison of various HARQ I schemes. It is shown that HARQ I-DAF is superior to HARQ I-AAF by around 4 dB at FER $= 10^{-2}$, due to an additional coding gain with the DTC structure. We also observe that HARQ I-ARP outperforms HARQ I-DAF by around 1.5 dB at FER $= 10^{-2}$. When HARQ I-DAF cannot decode the received signal correctly, the relay switches to HARQ I-AAF to forward the amplified signal to the destination, which circumvents the error propagation introduced by HARQ I-DAF scheme in this scenario. The destination can thus successfully decode some combined signals from the relay and source. The limited gain comes from HARQ I-AAF's contribution.

Figs. 4.4 and 4.5 compare the FER performance of various HARQ II schemes. Unlike the HARQ I scheme, HARQ II-AAF's performance is significantly better than HARQ II-DAF's performance. In Fig. 4.4, at FER $= 10^{-1}$, the SNR gain is by around 2.5 dB, and in Fig. 4.5, at FER $= 10^{-1}$, the SNR gain is by around 2 dB, while the gain increases as the FER decreases. The reason behind this phenomenon is that when the relay receives the first transmitted high code rate punctured packet from the source, the relay cannot correctly decode it most of the time. The process of decoding, interleaving and re-encoding introduces serious propagation errors. Even after three transmissions, the decoding error can still occur at the relay.

On the same figures, it is also observed that HARQ II-ARP still has the best performance. For example, in Fig. 4.4, at FER $= 10^{-2}$ HARQ II-ARP can provide SNR gains of about 1 dB and 6.5 dB compared to HARQ II-AAF and HARQ II-DAF; in Fig. 4.5, at FER $= 10^{-2}$

Frame Length	130 symbols (2 tail symbols)	
Encoder	Source and Relay	½ code rate (HARQI scheme)
	Generator Polynomial	$(5,7)_8$ RSC
	Interleaver (relay channel for DAF-DTC scheme)	S-random
Propagation channel	Data Channel	Quasi-static fading
	Feedback Channel	Ideal (error free)
SNR range	Direct channel Relay channel	0-14dB
	Inter-user channel	0-24dB
Modulation	BPSK	
Decoder	Component decoder	SOVA/VA
	Number of iterations (for iterative decoding method)	8
	Decoding judgement	CRC-16
HARQ	Type	Type I Type II (puncturing period 8) Type III (puncturing period 8)
	Maximum number of transmission	3
Upper Bound for Frame Error	10000	
FER	10^{-1} and 10^{-2}	

Figure 4.2: Simulation conditions

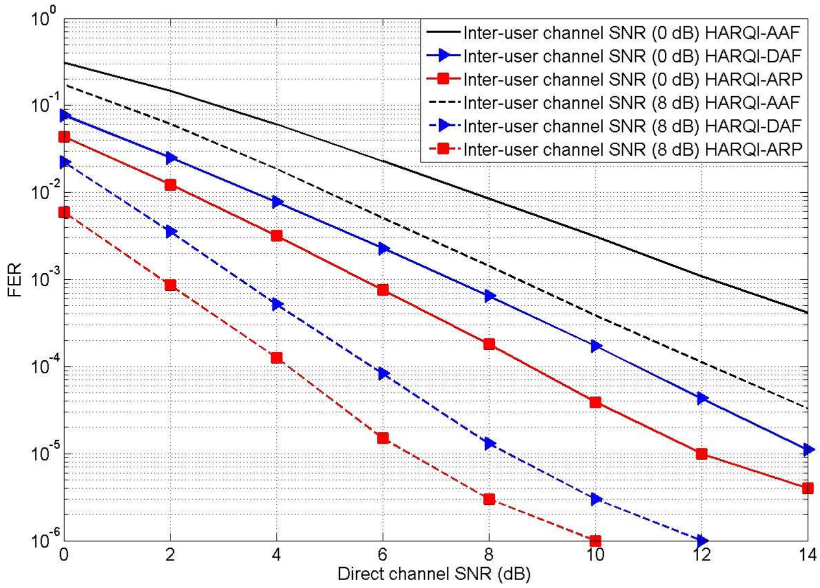

Figure 4.3: FER comparison of HARQ I-AAF, HARQ I-DAF and HARQ I-ARP schemes in a quasi-static fading channel with SNR 0-8 dB of the inter-user channel

HARQ II-ARP can provide SNR gains of about 0.8 dB and 5 dB compared to HARQ II-AAF and HARQ II-DAF. In the HARQ II-ARP scheme, when the decoding result at the relay is correct, the DTC scheme provides a high coding gain and this improves its performance significantly.

In Figs. 4.6 and 4.7, a similar FER performance comparison is shown for the HARQ III schemes. We observe that HARQ III-ARP is superior to HARQ III-AAF and HARQ III-DAF by around 1.5 dB and 4 dB at FER $= 10^{-2}$ in Fig. 4.6, and by around 1.5 dB and 2.8 dB at FER $= 10^{-2}$ in Fig. 4.7, respectively. In addition, HARQ III-AAF's performance is still better than HARQ III-DAF's performance, due to the same reason as for HARQ II.

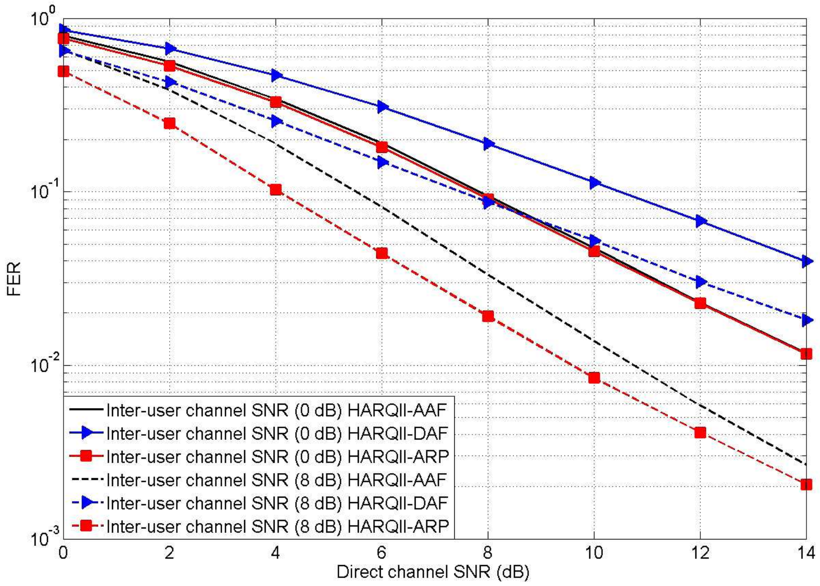

Figure 4.4: FER comparison of HARQ II-AAF, HARQ II-DAF and HARQ II-ARP schemes in a quasi-static fading channel with SNR 0-8 dB of the inter-user channel; the puncturing rates for the first transmission are 4/5

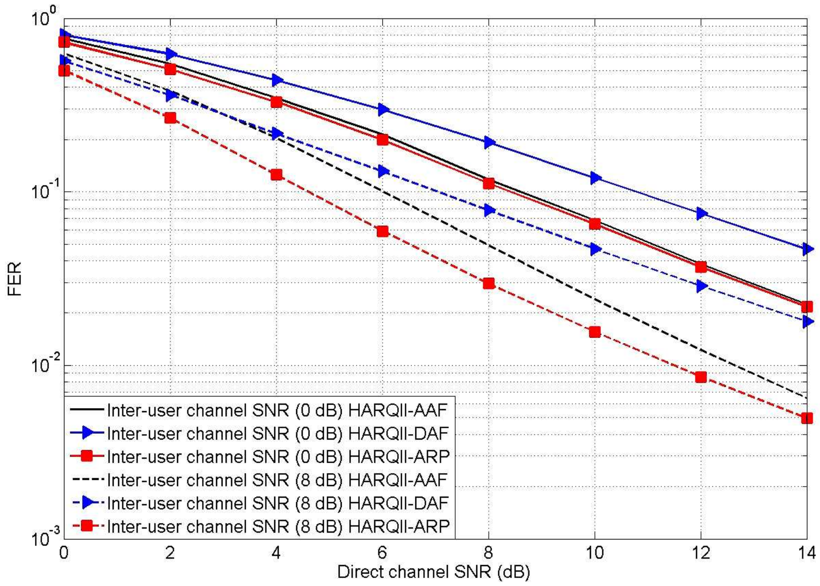

Figure 4.5: FER comparison of HARQ II-AAF, HARQ II-DAF and HARQ II-ARP schemes in a quasi-static fading channel with SNR 0-8 dB of the inter-user channel; the puncturing rates for the first transmission are 2/3

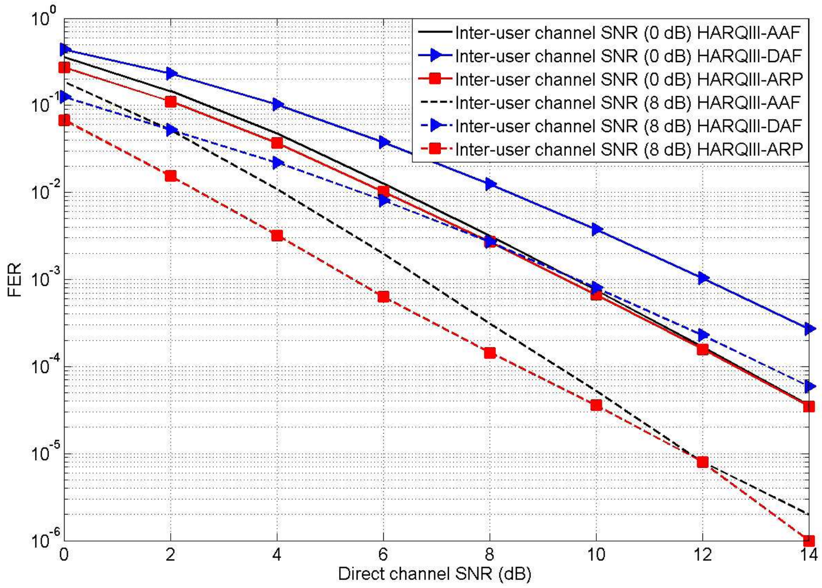

Figure 4.6: FER comparison of HARQ III-AAF, HARQ III-DAF and HARQ III-ARP schemes in a quasi-static fading channel with SNR 0-8 dB of the inter-user channel; the puncturing rates for the first transmission are 4/5

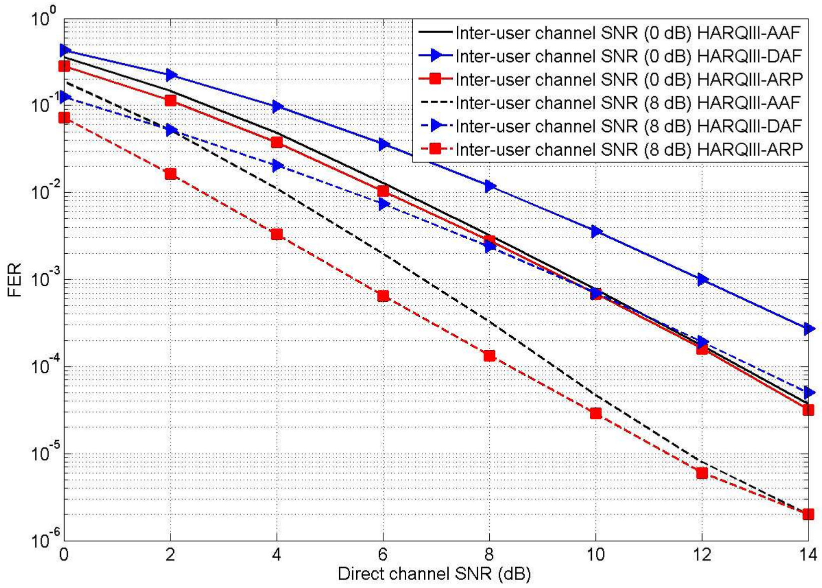

Figure 4.7: FER comparison of HARQ III-AAF, HARQ III-DAF and HARQ III-ARP schemes in a quasi-static fading channel with SNR 0-8 dB of the inter-user channel; the puncturing rates for the first transmission are 2/3

4.5 Conclusion

In this chapter, we studied the collaborative HARQ schemes for the cooperative diversity systems in wireless networks. According to the decoding result at the relay, we proposed an adaptive relaying protocol over the quasi-static fading channel. The proposed HARQ-ARP scheme combines the retransmission mechanisms (repetition coding and incremental redundancy), the distributed turbo coding and the proposed adaptive relaying strategy. Based on the different HARQ strategies, Chase combining and code combining were applied at the receiver. The simulation results indicated that the proposed HARQ-ARP scheme can achieve a superior FER compared to the reference schemes in all SNR regions.

Chapter 5

Performance Analysis of Collaborative HARQ Schemes in Wireless Networks

In the previous chapter, we proposed an adaptive relaying protocol and used it in the collaborative HARQ protocols for the CD communication systems. In this chapter, we present a theoretical analysis of the proposed HARQ-ARP scheme. The pairwise error probabilities (PEP) and word error probabilities (WEP) for all the HARQ schemes are derived for BPSK over a quasi-static fading channel. We then give a general throughput and FER expression for each HARQ scheme. Finally, we compare the performance with the reference systems, the HARQ-AAF scheme and HARQ-perfect DTC scheme, in which the received packet at the relay is assumed to be correct.

5.1 WEPs of the Relaying Protocols in the HARQ I Schemes

The calculation of the traditional union bound requires knowing the code distance spectrum, which necessitates an exhaustive search of the code trellis. Due to the high complexity of this search, similar to [82, 83], we consider an average upper bound. In order to calculate the WEP of HARQ-ARP, we need to know the probability of the erroneous or correctly received packet at the relay, that is, the probability of using either the HARQ-AAF or HARQ-perfect

DTC scheme at the relay. The corresponding WEPs of each type of relaying scheme are shown as follows.

5.1.1 WEP of HARQ I-AAF

Let $\gamma_{AAF}^{(n)}$ represent the instantaneous received SNR of the n^{th} combined packets from an AAF relay. Then we have

$$\gamma_{AAF}^{(n)} = \gamma_{sd} \sum_{i=1}^{n} |h_{sd}^{(i)}|^2 + \frac{1}{2} \sum_{i=1}^{n} H_2^{(i)}, \tag{5.1}$$

where $\gamma_{sd} = \frac{P_{sd}}{N_0}$, $H_2^{(i)} = (\frac{1}{2}\sum_{p=1}^{2} \frac{1}{\lambda_p})^{-1}$ is called the harmonic mean of variables [66] λ_p, $p = 1, 2$, $\lambda_1 = |h_{sr}^{(i)}|^2\gamma_{sr}$, and $\lambda_2 = |h_{rd}^{(i)}|^2\gamma_{rd}$, $\gamma_{sr} = \frac{P_{sr}}{N_0}$, $\gamma_{rd} = \frac{P_{rd}}{N_0}$. $h_{sd}^{(i)}, h_{sr}^{(i)}, h_{rd}^{(i)}$ are the fading coefficients of the direct, inter-user, and relay channel at the n^{th} transmission attempt, respectively. n is the maximum transmission attempt.

The PEP that the decoder makes a wrong decision by selecting an erroneous sequence with the Hamming distance d_1 of the combined packet at the receiver, for the HARQ I-AAF scenario, denoted by $P_n^{AAF-I}(d_1)$, is given by [24, 84]

$$\begin{aligned} P_n^{AAF-I}(d_1) &= E\left[Q\left(\sqrt{2d_1\gamma_{AAF}}\right)\right] \\ &\leq E\left[Q\left(\sqrt{2d_1\gamma_{sd}\sum_{i=1}^{n}|h_{sd}^{(i)}|^2}\right) Q\left(\sqrt{d_1\sum_{i=1}^{n} H_2^{(i)}}\right)\right]. \end{aligned} \tag{5.2}$$

The PEP of HARQ I-AAF scheme at the first transmission attempt, $P_1^{AAF-I}(d_1)$, can be calculated as

$$\begin{aligned} P_1^{AAF-I}(d_1) &\leq E\left[Q\left(\sqrt{2d_1\gamma_{sd}|h_{sd}^{(1)}|^2}\right) Q\left(\sqrt{d_1 H_2^{(1)}}\right)\right] \\ &\leq (d_1\gamma_{sd})^{-1} E\left[Q\left(\sqrt{d_1 H_2^{(1)}}\right)\right] \\ &= (d_1\gamma_{sd})^{-1} f_m(d_1), \end{aligned} \tag{5.3}$$

where

$$f_m(d_1) = E\left[Q\left(\sqrt{d_1 H_2^{(1)}}\right)\right]. \tag{5.4}$$

The function $Q(x)$ is defined as

$$Q(x) = \frac{1}{\sqrt{2\pi}} \int_x^\infty e^{-y^2/2} dy. \tag{5.5}$$

The closed form expression of $f_m(d_1)$ can be calculated by using the moment generating function (MDF) of the Harmonic mean of two exponential random variables [66]. At high SNR, $f_m(d_1)$ can be approximated as [39, 68]

$$f_m(d_1) = \left(\frac{1}{\gamma_{sr}} + \frac{1}{\gamma_{rd}}\right)(d_1)^{-1}. \tag{5.6}$$

By substituting Eq.(5.6) into Eq.(5.3), we have

$$P_1^{AAF-I}(d_1) \leq \frac{1}{(\gamma_{sd})}\left(\frac{1}{\gamma_{sr}} + \frac{1}{\gamma_{rd}}\right)(d_1)^{-2}. \tag{5.7}$$

Following similar calculations as in Eq.(5.3) to Eq.(5.7), Eq.(5.2) can be further generally written as

$$P_n^{AAF-I}(d_1) \leq \frac{1}{(\gamma_{sd})^n}\left(\frac{1}{\gamma_{sr}} + \frac{1}{\gamma_{rd}}\right)^n (d_1)^{-2n}. \tag{5.8}$$

Let $P_{F,AAF-I}$ represent the average WEP upper bound for combined packets from the direct and relay channel at the destination in the HARQ I-AAF scheme. Then we have

$$\begin{aligned} P_{F,AAF-I} &= \sum_{d_1=d_{1,min}}^{4l} \overline{A}(d_1) P_n^{AAF}(d_1) \\ &= \frac{1}{(\gamma_{sd})^n}\left(\frac{1}{\gamma_{sr}} + \frac{1}{\gamma_{rd}}\right)^n \sum_{d_1=d_{1,min}}^{4l} \overline{A}(d_1)\frac{1}{(d_1)^{2n}}, \end{aligned} \tag{5.9}$$

where $\overline{A}(d_1) = \sum_{j=1}^{l} \binom{l}{j} p(d_1|j)$, $\binom{l}{j}$ is the number of words with Hamming weight j and $p(d_1|j)$ is the probability that an input word with Hamming weight j produces a codeword with Hamming weight d_1.

5.1.2 WEP of HARQ I-perfect DTC

Let $\gamma_{DTC}^{(n)}$ represent the instantaneous received SNR of the n^{th} combined packets from a DAF relay using perfect DTC, then we have

$$\gamma_{DTC}^{(n)} = \gamma_{sd} \sum_{i=1}^{n} |h_{sd}^{(i)}|^2 + \gamma_{rd} \sum_{i=1}^{n} |h_{rd}^{(i)}|^2. \tag{5.10}$$

The average PEP of incorrectly decoding a combined packet using DTC at the n^{th} transmission attempt with Hamming weight d, denoted by $P_n^{(Perfect-I)}(d)$, can be calculated as

$$\begin{aligned} P_n^{(Perfect-I)}(d) &= E\left[Q\left(\sqrt{2d_1\gamma_{sd}\sum_{i=1}^{n}|h_{sd}^{(i)}|^2 + 2d_2\gamma_{rd}\sum_{i=1}^{n}|h_{rd}^{(i)}|^2}\right)\right] \\ &= E\left[Q\left(\sqrt{2d_1\gamma_{sd}\sum_{i=1}^{n}|h_{sd}^{(i)}|^2}\right) Q\left(\sqrt{2d_2\gamma_{rd}\sum_{i=1}^{n}|h_{rd}^{(i)}|^2}\right)\right] \\ &\leq \frac{1}{(\gamma_{sd})^n}\frac{1}{(\gamma_{rd})^n}(d_1)^{-n}(d_2)^{-n}, \end{aligned} \tag{5.11}$$

where d_1 and d_2 are the Hamming weights of the erroneous packets with Hamming weight d, transmitted from the source and relay respectively, such that $d = d_1 + d_2$.

Let $P_{F,Perfect-I}$ represent the average WEP upper bound for combined packets from the direct and relay channels at the destination in the HARQ I-perfect DTC scheme. Then we have

$$\begin{aligned} P_{F,Perfect-I} &= \sum_{d=d,min}^{4l} \overline{A}(d) P_n^{(Perfect)}(d) \\ &= \frac{1}{(\gamma_{sd})^n}\frac{1}{(\gamma_{rd})^n} \sum_{d=d,min}^{4l} \overline{A}(d) \frac{1}{(d_1)^n (d_2)^n}. \end{aligned} \tag{5.12}$$

5.1.3 WEP of HARQ I-ARP

The ARP is the combination of AAF and DAF. Therefore, to obtain the WEP of ARP, at first, we need to calculate the probability of a scenario using either AAF or perfect DTC at the relay.

Let $P_{F,sr-I}(d_{sr}, \gamma_{sr}|h_{sr}^{(n)})$ be the conditional PEP of incorrectly decoding a packet into another packet with Hamming distance of d_{sr} in the inter-user channel at the n^{th} transmission attempt. Then we have

$$P_{F,sr-I}(d_{sr}, \gamma_{sr}|h_{sr}^{(n)}) = Q\left(\sqrt{2d_{sr}\gamma_{sr}\sum_{i=1}^{n}|h_{sr}^{(i)}|^2}\right). \tag{5.13}$$

Let $P_{F,sr-I}(\gamma_{sr}|h_{sr})$ represent the conditional WEP in the inter-user channel for the HARQ I scheme, so

$$\begin{aligned} P_{F,sr-I}(\gamma_{sr}|h_{sr}^{(n)}) &= \sum_{d_{sr}=d_{sr,min}}^{2l} \overline{A}(d_{sr})P_{F,sr-I}(d_{sr}, \gamma_{sr}|h_{sr}^{(n)}) \\ &= \frac{1}{(\gamma_{sr})^n}\sum_{d_{sr}=d_{sr,min}}^{2l} \overline{A}(d_{sr})(d_{sr})^{-n}, \end{aligned} \tag{5.14}$$

where $d_{sr,min}$ is the minimum code Hamming distance.

Let $P^{(ARP-I)}$ represent the average PEP at high SNR for the ARP scheme at the relay. At the first transmission attempt, it can be approximated as

$$\begin{aligned}
& P_1^{(ARP-I)}(d) \\
\leq \;& E\left\{P_{F,sr-I}(d_{sr},\gamma_{sr}|h_{sr}^{(1)})P_1^{AAF}(d_1) + \left[1 - P_{F,sr-I}(d_{sr},\gamma_{sr}|h_{sr}^{(1)})\right] P_1^{(Perfect)}(d)\right\} \\
\leq \;& E\left\{P_{F,sr-I}(d_{sr},\gamma_{sr}|h_{sr}^{(1)})Q\left(\sqrt{2d_1\gamma_{sd}|h_{sd}^{(1)}|^2}\right)Q\left(\sqrt{d_1H_2^{(1)}}\right)\right. \\
+ \;& \left.\left[1 - P_{F,sr-I}(d_{sr},\gamma_{sr}|h_{sr}^{(1)})\right] Q\left(\sqrt{2d_1\gamma_{sd}|h_{sd}^{(1)}|^2}\right)Q\left(\sqrt{2d_2\gamma_{rd}|h_{rd}^{(1)}|^2}\right)\right\} \\
\leq \;& \frac{1}{d_1\gamma_{sd}}E\left\{P_{F,sr-I}(d_{sr},\gamma_{sr}|h_{sr}^{(1)})Q\left(\sqrt{d_1H_2^{(1)}}\right)\right. \\
+ \;& \left.\left[1 - P_{F,sr-I}(d_{sr},\gamma_{sr}|h_{sr}^{(1)})\right] Q\left(\sqrt{2d_2\gamma_{rd}|h_{rd}^{(1)}|^2}\right)\right\} \\
\leq \;& \frac{1}{d_1\gamma_{sd}}\left\{f(d_1) + \left(\frac{1 - \frac{1}{\gamma_{sr}}\sum_{d_{sr}=d_{sr,min}}^{2l}\overline{A}(d_{sr})\frac{1}{d_{sr}}}{d_2\gamma_{rd}}\right)\right\},
\end{aligned} \tag{5.15}$$

where

$$f(d_1) = E\left[\sum_{d_{sr}=d_{sr,min}}^{2l}\overline{A}(d_{sr})Q\left(\sqrt{2d_{sr}\gamma_{sr}|h_{sr}^{(1)}|^2}\right)Q\left(\sqrt{d_1H_2^{(1)}}\right)\right]. \tag{5.16}$$

The calculation of $f(d_1)$ in Eq.(5.16) depends on the product of two Q-functions, and both of them have the relationship with γ_{sr}. Therefore, we cannot approximate $Q\left(\sqrt{d_1H_2^{(1)}}\right)$ as we did in Eq.(5.6). By observing the expression of $H_2^{(1)}$, we can see that

$$H_2^{(1)} = \left[\frac{1}{2}\left(|h_{sr}^{(1)}|^2\gamma_{sr}\right)^{-1} + \left(|h_{rd}^{(1)}|^2\gamma_{rd}\right)^{-1}\right]^{-1} \approx 2\text{min}\left\{|h_{sr}^{(1)}|^2\gamma_{sr}, \gamma_{rd}h_{rd}^{(1)}|^2\right\}. \tag{5.17}$$

By using the above approximation, we can get

$$\begin{aligned}
& f(d_1) \\
\approx\ & E\left[\sum_{d_{sr}=d_{sr,min}}^{2l} \overline{A}(d_{sr}) Q\left(\sqrt{2d_{sr}\gamma_{sr}|h_{sr}^{(1)}|^2}\right) Q\left(\sqrt{2d_1 \min\left\{|h_{sr}^{(1)}|^2\gamma_{sr}, \gamma_{rd} h_{rd}^{(1)}|^2\right\}}\right)\right] \\
\leq\ & \sum_{d_{sr}=d_{sr,min}}^{2l} \overline{A}(d_{sr}) \frac{1}{\gamma_{sr}} \left[\frac{1}{(d_{sr}+d_1)} + \frac{1}{\gamma_{rd} d_{sr}(d_{sr}+d_1)}\right]. \qquad (5.18)
\end{aligned}$$

By substituting Eq.(5.18) into Eq.(5.15), we have Eq.(5.19):

$$\begin{aligned}
P_1^{(ARP-I)}(d) \leq\ & \frac{1}{d_1\gamma_{sd}} \left\{ \sum_{d_{sr}=d_{sr,min}}^{2l} \overline{A}(d_{sr}) \frac{1}{\gamma_{sr}} \left[\frac{1}{(d_{sr}+d_1)} + \frac{1}{\gamma_{rd} d_{sr}(d_{sr}+d_1)}\right] \right. \\
+\ & \left. \frac{1 - \frac{1}{\gamma_{sr}} \sum_{d_{sr}=d_{sr,min}}^{2l} \overline{A}(d_{sr}) \frac{1}{d_{sr}}}{d_2\gamma_{rd}} \right\} \\
=\ & \frac{1}{d_1\gamma_{sd}} \frac{1}{d_2\gamma_{rd}} \left(1 + \frac{\gamma_{rd}}{\gamma_{sr}} \sum_{d_{sr}=d_{sr,min}}^{2l} \overline{A}(d_{sr}) \frac{d_2}{d_{sr}+d_1}\right). \qquad (5.19)
\end{aligned}$$

Since the fading coefficients are independent during the retransmission attempts, the PEP of HARQ I-ARP of the n^{th} transmission can be generally written as

$$P_n^{(ARP-I)}(d) = \frac{1}{(\gamma_{sd}\gamma_{rd}d_1d_2)^n} \left[1 + \left(\frac{\gamma_{rd}}{\gamma_{sr}}\right)^n \sum_{d_{sr}=d_{sr,min}}^{2l} \overline{A}(d_{sr}) \frac{(d_2)^n}{(d_{sr}+d_1)^n}\right]. \qquad (5.20)$$

Then the average WEP upper bound in the HARQ I-ARP scheme can be expressed as

$$\begin{aligned}
P_{F,ARP-I} \leq\ & \sum_{d=d,min}^{4l} \overline{A}(d) P_n^{(ARP-I)}(d) \\
\leq\ & \frac{1}{(\gamma_{sd})^n(\gamma_{rd})^n} \sum_{d=d,min}^{4l} \overline{A}(d) \times \\
& \left\{ \frac{1}{(d_1)^n(d_2)^n} \left[1 + \left(\frac{\gamma_{rd}}{\gamma_{sr}}\right)^n \sum_{d_{sr}=d_{sr,min}}^{2l} \overline{A}(d_{sr}) \frac{(d_2)^n}{(d_{sr}+d_1)^n}\right] \right\}. \qquad (5.21)
\end{aligned}$$

5.2 WEPs of the Relaying Protocols in the HARQ II Schemes

Compared with the HARQ I schemes, the only difference when calculating the WEP of the relaying protocols in the HARQ II schemes is the n^{th} combined packet Hamming weight. In the HARQ I relaying schemes, the transmitter sends the repetition code in each transmission, so the n^{th} combined codeword Hamming weight is fixed. However, in the HARQ II relaying schemes, with the transmitted incremental redundancy during the retransmission attempts, the n^{th} combined packet Hamming weight changes.

5.2.1 WEP of HARQ II-AAF

Following a similar calculation to that in Eq.(5.2), the PEP for HARQ II-AAF, denoted by $P_n^{AAF-II}(d_1^{(n)})$, is given by

$$\begin{aligned}
P_n^{AAF-II}(d_1^{(n)}) &= E\left[Q\left(\sqrt{2d_1^{(n)}\gamma_{AAF}}\right)\right] \\
&\leq E\left[Q\left(\sqrt{2\gamma_{sd}\sum_{i=1}^{n} d_1^{(i)}|h_{sd}^{(i)}|^2}\right)Q\left(\sqrt{\sum_{i=1}^{n} d_1^{(i)}H_2^{(i)}}\right)\right] \\
&\leq E\left[Q\left(\sqrt{2\gamma_{sd}\left((d_1^{(0)}+d_1^{(1)})|h_{sd}^{(1)}|^2+\cdots+d_1^{(n)}|h_{sd}^{(n)}|^2\right)}\right)\times\right. \\
&\qquad \left. Q\left(\sqrt{(d_1^{(0)}+d_1^{(1)})H_2^{(1)}+\cdots+d_1^{(n)}H_2^{(n)}}\right)\right] \\
&\leq \frac{1}{(\gamma_{sd})^n}\left(\frac{1}{\gamma_{sr}}+\frac{1}{\gamma_{rd}}\right)^n (d_1^{(0)}+d_1^{(1)})^{-2}\cdots(d_1^{(n)})^{-2}, \qquad (5.22)
\end{aligned}$$

where $d_1^{(0)}$ and $d_1^{(1)}$ are the Hamming weight of the information symbols and parity symbols of the first transmitted code C_1, $d_1^{(i)}$ is the Hamming weight of incremental redundancy in the i^{th} transmission. If we assume that the maximum transmission attempts are n, $d_1 = d_1^{(0)} + d_1^{(1)} + \cdots + d_1^{(n)}$, which is the Hamming weight of the mother code C_m.

The average WEP upper bound for the HARQ II-AAF scheme can be expressed as

$$\begin{aligned} P_{F,AAF-II} &= \sum_{d_1^{(n)}=d_{1,min}^{(n)}}^{4l} \overline{A}(d_1^{(n)}) P_n^{AAF-II}(d_1^{(n)}) \\ &= \frac{1}{(\gamma_{sd})^n} \left(\frac{1}{\gamma_{sr}} + \frac{1}{\gamma_{rd}} \right)^n \sum_{d_1^{(n)}=d_{1,min}^{(n)}}^{4l} \overline{A}(d_1^{(n)}) \frac{1}{(d_1^{(0)} + d_1^{(1)})^2 \cdots (d_1^{(n)})^2}, \end{aligned} \quad (5.23)$$

where $\overline{A}(d_1^{(n)})$ is the number of codewords with Hamming weight $d_1^{(n)}$, $d_{1,min}^{(n)}$ is the combined packet for the n^{th} previous retransmission attempts.

5.2.2 WEP of HARQ II-perfect DTC

Similarly to that in Eq.(5.11), for perfect DTC used in the HARQ II scheme, the PEP can be calculated as

$$\begin{aligned} & P_n^{(Perfect-II)}(d^{(n)}) \\ &= E\left[Q\left(\sqrt{2\gamma_{sd} \sum_{i=1}^{n} d_1^{(i)} |h_{sd}^{(i)}|^2 + 2\gamma_{rd} \sum_{i=1}^{n} d_2^{(i)} |h_{rd}^{(i)}|^2} \right) \right] \\ &\leq \frac{1}{(\gamma_{sd})^n (\gamma_{rd})^n} (d_1^{(0)} + d_1^{(1)})^{-1} \cdots (d_1^{(n)})^{-1} (d_2^{(0)} + d_2^{(1)})^{-1} \cdots (d_2^{(n)})^{-1}. \end{aligned} \quad (5.24)$$

The WEP of perfect DTC used in the HARQ II scheme is given by

$$\begin{aligned} & P_{F,Perfect-II} \quad (5.25) \\ &= \sum_{d^{(n)}=d^{(n)},min}^{4l} \overline{A}(d^{(n)}) P_n^{(Perfect-II)}(d^{(n)}) \\ &= \frac{1}{(\gamma_{sd})^n (\gamma_{rd})^n} \sum_{d^{(n)}=d^{(n)},min}^{4l} \overline{A}(d^{(n)}) \frac{1}{\left[(d_1^{(0)} + d_1^{(1)}) \cdots d_1^{(n)}\right] \left[(d_2^{(0)} + d_2^{(1)}) \cdots d_2^{(n)}\right]}. \end{aligned}$$

5.2.3 WEP of HARQ II-ARP

In the HARQ II-ARP scheme, the conditional PEP of incorrectly decoding a codeword into another codeword in the inter-user channel at the n^{th} transmission attempt can be computed in a similar way to that in Eq.(5.13), so

$$P_{F,sr-II}(d_{sr}^{(n)},\gamma_{sr}|h_{sr}^{(n)}) = Q\left(\sqrt{2\gamma_{sr}\sum_{i=1}^{n} d_{sr}^{(i)}|h_{sr}^{(i)}|^2}\right). \tag{5.26}$$

Furthermore, the average PEP of the HARQ II-ARP scheme, similarly to Eq.(5.19) at high SNR, is given as

$$\begin{aligned}
&P_n^{(ARP-II)}(d^{(n)}) \qquad\qquad (5.27)\\
\leq\ & E\left\{P_{F,sr-II}(d_{sr}^{(n)},\gamma_{sr}|h_{sr}^{(n)})P_n^{AAF-II}(d_1^{(n)})\right.\\
+\ & \left.\left[1-P_{F,sr-II}(d_{sr}^{(n)},\gamma_{sr}|h_{sr}^{(n)})\right]P_n^{(Perfect-II)}(d^{(n)})\right\}\\
\leq\ & \frac{1}{\left[(\gamma_{sd})^n\left(d_1^{(0)}+d_1^{(1)}\right)\cdots d_1^{(n)}\right]\left[(\gamma_{rd})^n\left(d_2^{(0)}+d_2^{(1)}\right)\cdots d_2^{(n)}\right]}\times\\
& \left\{1+\left(\frac{\gamma_{rd}}{\gamma_{sd}}\right)^n\sum_{d_{sr}^{(n)}=d_{sr}^{(n)},min}^{2l}\overline{A}(d_{sr}^{(n)})\frac{\left(d_2^{(0)}+d_2^{(1)}\right)\cdots d_2^{(n)}}{\left[\left(d_{sr}^{(0)}+d_{sr}^{(1)}\right)+\left(d_1^{(0)}+d_1^{(1)}\right)\right]\cdots\left(d_{sr}^{(n)}+d_1^{(n)}\right)}\right\}.
\end{aligned}$$

Then, the average WEP upper bound for the HARQ II-ARP scheme can be written as

$$\begin{aligned}
&P_{F,ARP-II} \qquad\qquad (5.28)\\
\leq\ & \sum_{d^{(n)}=d^{(n)},min}^{4l}\overline{A}(d^{(n)})P_n^{(ARP-II)}(d^{(n)})\\
\leq\ & \frac{1}{(\gamma_{sd})^n(\gamma_{rd})^n}\sum_{d^{(n)}=d^{(n)},min}^{4l}\overline{A}(d^{(n)})\left\{\frac{1}{\left[\left(d_1^{(0)}+d_1^{(1)}\right)\cdots d_1^{(n)}\right]\left[\left(d_2^{(0)}+d_2^{(1)}\right)\cdots d_2^{(n)}\right]}\times\right.\\
& \left.\left[1+\left(\frac{\gamma_{rd}}{\gamma_{sr}}\right)^n\sum_{d_{sr}=d_{sr}^{(n)},min}^{2l}\overline{A}(d_{sr}^{(n)})\frac{\left(d_2^{(0)}+d_2^{(1)}\right)\cdots d_2^{(n)}}{\left[\left(d_{sr}^{(0)}+d_{sr}^{(1)}\right)+\left(d_1^{(0)}+d_1^{(1)}\right)\right]\cdots\left(d_{sr}^{(n)}+d_1^{(n)}\right)}\right]\right\}.
\end{aligned}$$

5.3 WEPs of the Relaying Protocols in the HARQ III Schemes

As we presented in section 4.3.3, the HARQ III scheme uses the current received packet to decode, because of its self-decodable property. The receiver only combines all the received packets together for decoding if the decoding result is not correct. Since each transmitted packet contains the information symbols and different parity symbols, the Hamming weight calculation of the combined codes, at the n^{th} transmission attempt, is thus different.

5.3.1 WEP of HARQ III-AAF

Similarly to Eq.(5.22), the PEP for the HARQ III-AAF scheme, denoted by $P_n^{AAF-III}(d_1^{(n)})$, can be expressed as

$$\begin{aligned}
& P_n^{AAF-III}(d_1^{(n)}) \\
& = E\left[Q\left(\sqrt{2d_1^{(n)}\gamma_{AAF}}\right)\right] \\
& \leq E\left[Q\left(\sqrt{2\gamma_{sd}\left(\sum_{i=1}^{n}(d_1^{(0)}+d_1^{(i)})|h_{sd}^{(i)}|^2\right)}\right)Q\left(\sqrt{\left(\sum_{i=1}^{n}(d_1^{(0)}+d_1^{(i)})H_2^i\right)}\right)\right] \\
& \leq \frac{1}{(\gamma_{sd})^n}\left(\frac{1}{\gamma_{sr}}+\frac{1}{\gamma_{rd}}\right)^n\left[\prod_{i=1}^{n}(d_1^{(0)}+d_1^{(i)})\right]^{-2}. \qquad (5.29)
\end{aligned}$$

The WEP of the HARQ III-AAF is given by

$$\begin{aligned}
& P_{F,AAF-III} \\
& = \sum_{d_1^{(n)}=d_1^{(n)},min}^{4l}\overline{A}(d_1^{(n)})P_n^{AAF-III}(d_1^{(n)}) \\
& = \frac{1}{(\gamma_{sd})^n}\left(\frac{1}{\gamma_{sr}}+\frac{1}{\gamma_{rd}}\right)^n\sum_{d_1^{(n)}=d_1^{(n)},min}^{4l}\overline{A}(d_1^{(n)})\frac{1}{\left[\prod_{i=1}^{n}(d_1^{(0)}+d_1^{(i)})\right]^2}. \qquad (5.30)
\end{aligned}$$

5.3.2 WEP of HARQ III-perfect DTC

Following a similar analysis to that in Eq.(5.24), for the perfect DTC in the HARQ III scheme, the PEP can be calculated as

$$\begin{aligned} & P_n^{(Perfect-III)}(d^{(n)}) \\ = \quad & E\left[Q\left(\sqrt{2\gamma_{sd}\sum_{i=1}^{n} d_1^{(i)}|h_{sd}^{(i)}|^2 + 2\gamma_{rd}\sum_{i=1}^{n} d_2^{(i)}|h_{rd}^{(i)}|^2}\right)\right] \\ \leq \quad & \frac{1}{(\gamma_{sd})^n(\gamma_{rd})^n}\frac{1}{\prod_{i=1}^{n}(d_1^{(0)}+d_1^{(i)})\prod_{i=1}^{n}(d_2^{(0)}+d_2^{(i)})}. \end{aligned} \tag{5.31}$$

The WEP of the perfect-DTC used in the HARQ III scheme, is given by

$$\begin{aligned} & P_{F,Perfect-III} \\ = \quad & \sum_{d^{(n)}=d^{(n)},min}^{4l} \overline{A}(d^{(n)})P_n^{(Perfect-III)}(d^{(n)}) \\ = \quad & \frac{1}{(\gamma_{sd})^n(\gamma_{rd})^n}\sum_{d^{(n)}=d^{(n)},min}^{4l}\overline{A}(d^{(n)})\frac{1}{\prod_{i=1}^{n}(d_1^{(0)}+d_1^{(i)})\prod_{i=1}^{n}(d_2^{(0)}+d_2^{(i)})}. \end{aligned} \tag{5.32}$$

5.3.3 WEP of HARQ III-ARP

Following a similar analysis to that in Eq.(5.27), the average PEP of the HARQ III-ARP scheme at high SNR is given as

$$\begin{aligned} & P_n^{(ARP-III)}(d^{(n)}) \\ \leq \quad & E\Big\{P_{F,sr-III}(d_{sr}^{(n)},\gamma_{sr}|h_{sr}^{(n)})P_n^{AAF-III}(d_1^{(n)}) \\ + \quad & \left[1-P_{F,sr-III}(d_{sr}^{(n)},\gamma_{sr}|h_{sr}^{(n)})\right]P_n^{(Perfect-III)}(d^{(n)})\Big\} \\ \leq \quad & \frac{1}{\left[(\gamma_{sd})^n\prod_{i=1}^{n}\left(d_1^{(0)}+d_1^{(i)}\right)\right]\left[(\gamma_{rd})^n\prod_{i=1}^{n}\left(d_2^{(0)}+d_2^{(i)}\right)\right]}\times \\ & \left\{1+\left(\frac{\gamma_{rd}}{\gamma_{sd}}\right)^n\sum_{d_{sr}^{(n)}=d_{sr}^{(n)},min}^{2l}\overline{A}(d_{sr}^{(n)})\frac{\prod_{i=1}^{n}\left(d_2^{(0)}+d_2^{(i)}\right)}{\prod_{i=1}^{n}\left[\left(d_{sr}^{(0)}+d_{sr}^{(i)}\right)+\left(d_1^{(0)}+d_1^{(i)}\right)\right]}\right\} \end{aligned} \tag{5.33}$$

Then, the average WEP upper bound for the HARQ III-ARP scheme can be written as

$$\begin{aligned}
&P_{F,ARP-III} \\
&\leq \sum_{d^{(n)}=d^{(n)},min}^{4l} \overline{A}(d^{(n)}) P_n^{(ARP-III)}(d^{(n)}) \\
&\leq \frac{1}{(\gamma_{sd})^n(\gamma_{rd})^n} \sum_{d^{(n)}=d^{(n)},min}^{4l} \overline{A}(d^{(n)}) \left\{ \frac{1}{\left[\prod_{i=1}^n \left(d_1^{(0)}+d_1^{(i)}\right)\right]\left[\prod_{i=1}^n \left(d_2^{(0)}+d_2^{(i)}\right)\right]} \times \right. \\
&\left. \left[1+\left(\frac{\gamma_{rd}}{\gamma_{sr}}\right)^n \sum_{d_{sr}=d_{sr}^{(n)},min}^{2l} \overline{A}(d_{sr}^{(n)}) \frac{\prod_{i=1}^n \left(d_2^{(0)}+d_2^{(i)}\right)}{\prod_{i=1}^n \left[\left(d_{sr}^{(0)}+d_{sr}^{(i)}\right)+\left(d_1^{(0)}+d_1^{(i)}\right)\right]}\right]\right\}. \quad (5.34)
\end{aligned}$$

5.4 Throughput Analysis

In this section, we analyze the throughput of various HARQ schemes. The average throughput, denoted by R_{AV} can be calculated as in [11, 13],

$$R_{AV} = \frac{P}{P+l_{av}} \frac{k}{k+n_c+m} \quad (5.35)$$

where l_{av} is the average number of additional transmitted symbols per P information symbols, k is the number of information symbols, n_c is the length of CRC check symbols for error detection, and m is the number of tail symbols. The factor $k/(k+n_c+m)$ is the loss in the throughput due to the added parity symbols for error detection and the tail symbols to each transmitted packet.

Eq.(5.35) applies to all the HARQ schemes, including the HARQI (repetition coding), HARQII and HARQIII (incremental redundancy) schemes. In this thesis, the source and relay use same code rate during each transmission. Therefore, the length of each transmitted packet is the same. Let P_i denote the average number of additional transmitted symbols for each transmission attempt, and let $P_{F_{SD}^{(i)}}$, $P_{F_{com}^{(i)}}$ denote the probability of the decoded combined packets from direct channel containing errors, and the decoded combined packets from direct and relay channels containing errors, at the i^{th} transmission attempt, $i = 1, 2, \cdots, n$, respectively. Clearly, with the HARQ relaying schemes, in general, we have l_{av}

$$l_{av} = P_1 + P_1 P_{F_{SD}^{(1)}} + P_2 P_{F_{com}^{(1)}} + \cdots + P_i P_{F_{SD}^{(i)}} + P_{i+1} P_{F_{com}^{(i)}} + \cdots + P_{n-1} P_{F_{SD}^{(n-1)}} + P_n P_{F_{com}^{(n-1)}}, \tag{5.36}$$

where $P_{F_{com}^{(i)}}$ can be expressed for the three types of HARQ schemes, given by Eqs.(5.9), (5.23) and (5.30) for the HARQ-AAF scheme, by Eqs.(5.12), (5.25) and (5.32) for the HARQ-perfect DTC scheme and by Eqs.(5.21), (5.28) and (5.34) for the HARQ-ARP scheme, respectively.

The term $P_{F_{SD}^{(i)}}$ can be calculated similarly to that in Eq.(5.2) or in Eq.(5.11), depending on the relaying protocols. For example, in the HARQI schemes, the general expression of $P_{F_{SD}^{(n)}}$ for the HARQ-AAF scheme is given by

$$\begin{aligned} & P_{F_{SD}^{(n)}, AAF-I} \\ \leq \;& E\left[Q\left(\sqrt{2d_1 \gamma_{sd} \sum_{i=1}^{n} |h_{sd}^{(i)}|^2} \right) Q\left(\sqrt{d_1 H_2^{(n-1)}} \right) \right] \\ \leq \;& \frac{1}{(\gamma_{sd})^n (\gamma_{rd})^{n-1}} \sum_{d_1 = d_{1,min}}^{4l} \overline{A}(d_1) \frac{1}{(d_1)^{(2n-1)}} \left\{ 1 + (\frac{\gamma_{rd}}{\gamma_{sr}})^{n-1} \right\}. \end{aligned} \tag{5.37}$$

In the HARQI perfect-DTC scheme, $P_{F_{SD}^{(n)}}$ is given by

$$\begin{aligned} & P_{F_{SD}^{(n)}, Perfect-I} \\ \leq \;& E\left(Q\left(\sqrt{2d_1 \gamma_{sd} \sum_{i=1}^{n} |h_{sd}^{(i)}|^2} \right) Q\left(\sqrt{2d_2 \gamma_{rd} \sum_{i=1}^{n-1} |h_{rd}^{(i)}|^2} \right) \right] \\ \leq \;& \frac{1}{(\gamma_{sd})^n (\gamma_{rd})^{(n-1)}} \sum_{d = d,min}^{4l} \overline{A}(d) \frac{1}{(d_1)^n (d_2)^{(n-1)}}. \end{aligned} \tag{5.38}$$

Following a similar calculation to that in Eq.(5.15), the $P_{F_{SD}^{(n)}, ARP-I}$ in HARQI-ARP can be expressed as

$$\begin{aligned}
&P_{F_{SD}^{(n)},ARP-I} \\
&\leq \; E\left\{P_{F,sr-I}(d_{sr},\gamma_{sr}|h_{sr}^{(i)})Q\left(\sqrt{2d_1\gamma_{sd}\sum_{i=1}^{n}|h_{sd}^{(i)}|^2+d_1H_2^{(n-1)}}\right)\right. \\
&+ \; \left.\left[1-P_{F,sr-I}(d_{sr},\gamma_{sr}|h_{sr}^{(i)})\right]Q\left(\sqrt{2d_1\gamma_{sd}\sum_{i=1}^{n}|h_{sd}^{(i)}|^2+2d_2\gamma_{rd}\sum_{i=1}^{n-1}|h_{rd}^{(i)}|^2}\right)\right\} \\
&\leq \; \frac{1}{(\gamma_{sd})^n(\gamma_{rd})^{n-1}}\sum_{d=d,min}^{4l}\overline{A}(d)\frac{1}{(d_1)^n(d_2)^{n-1}}\left\{1+\left(\frac{\gamma_{rd}}{\gamma_{sr}}\right)^{n-1}\times\right. \\
&\quad \left.\sum_{d_{sr}=d_{sr,min}}^{2l}\overline{A}(d_{sr})\frac{(d_2)^{n-1}}{(d_{sr}+d_1)^{n-1}}\right\}. \qquad (5.39)
\end{aligned}$$

Therefore, the $P_{F_{SD}^{(i)}}$ for three types of HARQ schemes, can be generally expressed as follows:

$$P_{F_{SD}^{(n)},AAF-p}=\frac{1}{(\gamma_{sd})^n(\gamma_{rd})^{n-1}}\sum_{d_1=d_{1,min}^{(n)}}^{4l}\overline{A}(d_1)\left\{\theta(\gamma)D_p^{'}\right\}, \qquad (5.40)$$

$$P_{F_{SD}^{(n)},Perfect-p}=\frac{1}{(\gamma_{sd})^n(\gamma_{rd})^{n-1}}\sum_{d=d^{(n)},min}^{4l}\overline{A}(d)D_p, \qquad (5.41)$$

$$P_{F_{SD}^{(n)},ARP-p}=\frac{1}{(\gamma_{sd})^n(\gamma_{rd})^{n-1}}\sum_{d=d^{(n)},min}^{4l}\overline{A}(d)\left\{\theta(\gamma)D_pS_p\right\}, \qquad (5.42)$$

where $D_p^{'}$ in Eq.(5.44), D_p in Eq.(5.45), and S_p in Eq.(5.46) represent the coefficients for the above $P_{F_{SD}^{(i)}}$ expressions, p represents the protocol type number, denoted by I, II and III, respectively; and

$$\theta(\gamma)=1+\left(\frac{\gamma_{rd}}{\gamma_{sr}}\right)^{n-1}, \qquad (5.43)$$

$$\begin{aligned}
D_I^{'} &= \frac{1}{(d_1)^{2n-1}} \\
D_{II}^{'} &= \frac{1}{\left[\left(d_1^{(0)} + d_1^{(1)} \cdots d_1^{(n)}\right)\right]\left[\left(d_1^{(0)} + d_1^{(1)} \cdots d_1^{(n-1)}\right)\right]} \\
D_{III}^{'} &= \frac{1}{\left[\prod_{i=1}^{n}\left(d_1^{(0)} + d_1^{(i)}\right)\right]\left[\prod_{i=1}^{n-1}\left(d_1^{(0)} + d_1^{(i)}\right)\right]},
\end{aligned} \tag{5.44}$$

$$\begin{aligned}
D_I &= \frac{1}{(d_1)^n (d_2)^{n-1}} \\
D_{II} &= \frac{1}{\left[\left(d_1^{(0)} + d_1^{(1)} \cdots d_1^{(n)}\right)\right]\left[\left(d_2^{(0)} + d_2^{(1)} \cdots d_2^{(n-1)}\right)\right]} \\
D_{III} &= \frac{1}{\left[\prod_{i=1}^{n}\left(d_1^{(0)} + d_1^{(i)}\right)\right]\left[\prod_{i=1}^{n-1}\left(d_2^{(0)} + d_2^{(i)}\right)\right]},
\end{aligned} \tag{5.45}$$

$$\begin{aligned}
S_I &= \left\{\sum_{d_{sr}=d_{sr,min}}^{2l} \overline{A}(d_{sr}) \frac{(d_2)^{n-1}}{(d_{sr}+d_1)^{n-1}}\right\} \\
S_{II} &= \left\{\sum_{d_{sr}=d_{sr,min}^{(n-1)}}^{2l} \overline{A}(d_{sr}^{(n-1)}) \frac{\left(d_2^{(0)} + d_2^{(1)}\right) \cdots d_2^{(n-1)}}{\left[\left(d_{sr}^{(0)} + d_{sr}^{(1)}\right) + \left(d_1^{(0)} + d_1^{(1)}\right)\right] \cdots \left(d_{sr}^{(n-1)} + d_1^{(n-1)}\right)}\right\} \\
S_{III} &= \left\{\sum_{d_{sr}=d_{sr,min}^{(n-1)}}^{2l} \overline{A}(d_{sr}^{(n-1)}) \frac{\prod_{i=1}^{n-1}\left(d_2^{(0)} + d_2^{(i)}\right)}{\prod_{i=1}^{n-1}\left[\left(d_{sr}^{(0)} + d_{sr}^{(i)}\right) + \left(d_1^{(0)} + d_1^{(i)}\right)\right]}\right\}.
\end{aligned} \tag{5.46}$$

Substituting Eq.(5.36) into Eq.(5.35), we can obtain the average throughput expression.

5.5 Performance Comparison between HARQ-ARP and HARQ-perfect DTC

In this thesis, we apply the same code to all HARQ relaying schemes, so the code distance spectrum is fixed during the comparison.

To evaluate the performance between HARQ-ARP and HARQ-perfect DTC, we need to investigate the expression $P_{F^{(i)}_{com}}$ and $P_{F^{(i)}_{SD}}$ for various HARQ schemes. To make this comparison clear and easy to follow, first we give the general expression of WEPs for three types of HARQ-ARP by Eqs.(5.21), (5.28) and (5.34) and HARQ-perfect DTC by Eqs.(5.12), (5.25) and (5.32)

$$P_{F,ARP-p} = \frac{1}{(\gamma_{sd})^n(\gamma_{rd})^n} \sum_{d^{(n)}=d^{(n)},min}^{4l} \overline{A}(d^{(n)})C_pR_p, \tag{5.47}$$

$$P_{F,Perfect-p} = \frac{1}{(\gamma_{sd})^n(\gamma_{rd})^n} \sum_{d^{(n)}=d^{(n)},min}^{4l} \overline{A}(d^{(n)})C_p, \tag{5.48}$$

where C_p in Eq.(5.49) and R_p in Eq.(5.50) represent the coefficients for the above WEP expressions.

$$\begin{aligned} C_I &= \frac{1}{(d_1)^n(d_2)^n} \\ C_{II} &= \frac{1}{\left[(d_1^{(0)}+d_1^{(1)})\cdots(d_1^{(n)})\right]\left[(d_2^{(0)}+d_2^{(1)})\cdots(d_2^{(n)})\right]} \\ C_{III} &= \frac{1}{\left[\prod_{i=1}^n(d_1^{(0)}+d_1^{(i)})\right]\left[\prod_{i=1}^n(d_2^{(0)}+d_2^{(i)})\right]}, \end{aligned} \tag{5.49}$$

$$\begin{aligned} R_I &= \left\{1+\left(\frac{\gamma_{rd}}{\gamma_{sr}}\right)^n \sum_{d_{sr}=d_{sr,min}}^{2l} \overline{A}(d_{sr})\frac{(d_2)^n}{(d_{sr}+d_1)^n}\right\} \\ R_{II} &= \left\{1+\left(\frac{\gamma_{rd}}{\gamma_{sr}}\right)^n \sum_{d_{sr}=d_{sr}^{(n)},min}^{2l} \overline{A}(d_{sr}^{(n)})\frac{\left(d_2^{(0)}+d_2^{(1)}\right)\cdots d_2^{(n)}}{\left[\left(d_{sr}^{(0)}+d_{sr}^{(1)}\right)+\left(d_1^{(0)}+d_1^{(1)}\right)\right]\cdots\left(d_{sr}^{(n)}+d_1^{(n)}\right)}\right\} \\ R_{III} &= \left\{1+\left(\frac{\gamma_{rd}}{\gamma_{sr}}\right)^n \sum_{d_{sr}=d_{sr}^{(n)},min}^{2l} \overline{A}(d_{sr}^{(n)})\frac{\prod_{i=1}^n\left(d_2^{(0)}+d_2^{(i)}\right)}{\prod_{i=1}^n\left[\left(d_{sr}^{(0)}+d_{sr}^{(i)}\right)+\left(d_1^{(0)}+d_1^{(i)}\right)\right]}\right\}. \end{aligned} \tag{5.50}$$

Clearly, for the HARQ-ARP scheme, the FER expression of the unsuccessful decoding at the destination at the n^{th} transmission attempt is

$$\begin{aligned} P^{(n)}_{FER,ARP} &= P^{(1)}_{SD,ARP-p} P^{(1)}_{F,ARP-p} \cdots P^{(n)}_{SD,ARP-p} P^{(n)}_{F,ARP-p} \\ &= \prod_{i=1}^{n} P^{(i)}_{SD,ARP-p} \prod_{i=1}^{n} P^{(i)}_{F,ARP-p}. \end{aligned} \tag{5.51}$$

For the HARQ-perfect DTC scheme, the FER expression can be similarly expressed as

$$P^{(n)}_{FER,Perfect} = \prod_{i=1}^{n} P^{(i)}_{SD,Perfect-p} \prod_{i=1}^{n} P^{(i)}_{F,Perfect-p}. \tag{5.52}$$

Since the additional transmitted symbols for each transmission attempt P_i, and the probability of decoding a packet in the first transmission from the direct channel $P_{F^{(1)}_{SD}}$ in Eq.(5.36) are the same for both schemes, we only need to study the expression $P^{(i)}_{F,ARP-p}$ and $P^{(i)}_{F,Perfect-p}$, $i = 1, 2, \cdots, n$ for the HARQ-ARP by Eq.(5.47) and HARQ-perfect DTC schemes by Eq.(5.48), respectively; and $P^{(i)}_{SD}, i = 2, \cdots, n$ given by Eq.(5.41) for the HARQ-perfect DTC scheme and Eq.(5.42) for the HARQ-ARP, respectively.

It can be noted that as γ_{sr} increases, $(\frac{\gamma_{rd}}{\gamma_{sr}}) \to 0$, S_p in Eq.(5.46) and R_p in Eq.(5.50) approach one, so the value of the $P^{(i)}_{F,ARP-p}$ in Eq.(5.47) and $P_{F^{(i)}_{SD,ARP-p}}$ in Eq.(5.42) approach the value of the $P^{(i)}_{F,Perfect-p}$ in Eq.(5.48) and $P_{F^{(i)}_{SD,Perfect-p}}$ in Eq.(5.41), respectively.

As each item in Eqs.(5.51) and (5.52) has the same relationship, we conclude that the performance of HARQ-ARP approaches HARQ-perfect DTC in the high γ_{sr} region.

5.6 Performance Comparison between HARQ-ARP and HARQ-AAF

In this section, we compare the performance between HARQ-ARP and HARQ-AAF. In order to analyze their performance, we can follow a similar analysis in section 5.5. The HARQ-ARP is based on a DTC scheme, and we need to evaluate its distance spectrum, which makes the analysis and comparison complicated and more difficult. However, it is well known that the turbo codes outperform convolutional codes at medium to high SNRs [45]. We also observed that both HARQ-AAF and HARQ-ARP with a DAF scheme (the relay fully decodes the received signal, re-encodes and forwards it to the destination) are based on convolutional codes, and it is easy to compare the performance of these two schemes. Motivated by this, in this thesis, we compare the performance of the HARQ-ARP with a DAF scheme and the HARQ-AAF scheme.

We will demonstrate that the performance of HARQ-ARP with a DAF scheme is superior to the HARQ-AAF scheme. As a result, the HARQ-ARP with a DTC scheme exhibits better performance than the HARQ-AAF scheme. For the HARQ-ARP with a DAF scheme, the signals transmitted from the source and relay are the same codes, so their Hamming weights of the erroneous packets are the same. Therefore, following the similar WEP calculations that we used for the ARP-DTC scheme in Eqs.(5.21), (5.28), and (5.34), the WEP of ARP-DAF for each type of HARQ scheme can be generally calculated as

$$P_{F,_{DAF}^{ARP}-p} = \frac{1}{(\gamma_{sd})^n(\gamma_{rd})^n} \sum_{d_1=d_1^{(n)},min}^{4l} \overline{A}(d_1)C_p^{'} \left\{1+\left(\frac{\gamma_{rd}}{\gamma_{sr}}\right)^n R_p^{'}\right\}, \tag{5.53}$$

where $R_p^{'}$ in Eq.(5.53) can be expressed as in Eq.(5.54)

$$\begin{aligned}
R_I^{'} &= \left\{\sum_{d_{sr}=d_{sr,min}}^{2l} \overline{A}(d_{sr})\frac{(d_1)^n}{(d_{sr}+d_1)^n}\right\} \\
R_{II}^{'} &= \left\{\sum_{d_{sr}=d_{sr}^{(n)},min}^{2l} \overline{A}(d_{sr}^{(n)})\frac{\left(d_1^{(0)}+d_1^{(1)}\right)\cdots d_1^{(n)}}{\left[\left(d_{sr}^{(0)}+d_{sr}^{(1)}\right)+\left(d_1^{(0)}+d_1^{(1)}\right)\right]\cdots\left(d_{sr}^{(n)}+d_1^{(n)}\right)}\right\} \\
R_{III}^{'} &= \left\{\sum_{d_{sr}=d_{sr}^{(n)},min}^{2l} \overline{A}(d_{sr}^{(n)})\frac{\prod_{i=1}^{n}\left(d_1^{(0)}+d_1^{(i)}\right)}{\prod_{i=1}^{n}\left[\left(d_{sr}^{(0)}+d_{sr}^{(i)}\right)+\left(d_1^{(0)}+d_1^{(i)}\right)\right]}\right\},
\end{aligned} \tag{5.54}$$

and the $P_{F_{SD}^{(i)}}$ for the HARQ ARP-DAF schemes is given by

$$P_{F_{SD}^{(n)},_{DAF}^{ARP}-p} = \frac{1}{(\gamma_{sd})^n(\gamma_{rd})^{n-1}} \sum_{d_1=d_1^{(n)},min}^{4l} \overline{A}(d)\left\{\theta(\gamma)D_p^{'}S_p^{'}\right\}, \tag{5.55}$$

where

$$\begin{aligned}
S_I^{'} &= \left\{ \sum_{d_{sr}=d_{sr,min}}^{2l} \overline{A}(d_{sr}) \frac{(d_1)^{n-1}}{(d_{sr}+d_1)^{n-1}} \right\} \\
S_{II}^{'} &= \left\{ \sum_{d_{sr}=d_{sr,min}^{(n-1)}}^{2l} \overline{A}(d_{sr}^{(n-1)}) \frac{\left(d_1^{(0)}+d_1^{(1)}\right)\cdots d_1^{(n-1)}}{\left[\left(d_{sr}^{(0)}+d_{sr}^{(1)}\right)+\left(d_1^{(0)}+d_1^{(1)}\right)\right]\cdots\left(d_{sr}^{(n-1)}+d_1^{(n-1)}\right)} \right\} \\
S_{III}^{'} &= \left\{ \sum_{d_{sr}=d_{sr,min}^{(n-1)}}^{2l} \overline{A}(d_{sr}^{(n-1)}) \frac{\prod_{i=1}^{n-1}\left(d_1^{(0)}+d_1^{(i)}\right)}{\prod_{i=1}^{n-1}\left[\left(d_{sr}^{(0)}+d_{sr}^{(i)}\right)+\left(d_1^{(0)}+d_1^{(i)}\right)\right]} \right\}. \qquad (5.56)
\end{aligned}$$

In addition, the general expression of the HARQ-AAF schemes by Eqs.(5.9), (5.23) and (5.30), can be written as

$$P_{F,_{AAF}-p} = \frac{1}{(\gamma_{sd})^n(\gamma_{rd})^n} \sum_{d_1=d_1^{(n)},min}^{4l} \overline{A}(d_1)C_p^{'}.\left\{1+\left(\frac{\gamma_{rd}}{\gamma_{sr}}\right)^n\right\}. \qquad (5.57)$$

Following a similar analysis to that in section 5.5, the FER expressions for the HARQ ARP-DAF and HARQ-AAF schemes are given by

$$P_{FER,_{DAF}^{ARP}}^{(n)} = \prod_{i=1}^{n} P_{F_{SD}^{(i)},_{DAF}^{ARP}-p} \prod_{i=1}^{n} P_{F,_{DAF}^{ARP}-p}^{(i)}. \qquad (5.58)$$

$$P_{FER,AAF}^{(n)} = \prod_{i=1}^{n} P_{SD,AAF-p}^{(i)} \prod_{i=1}^{n} P_{F,AAF-p}^{(i)}. \qquad (5.59)$$

Since $R_p^{'} \leq 1$ and $S_p^{'} \leq 1$, which have been proved in Appendix B, by comparing $P_{F,_{DAF}^{ARP}-p}$ in Eq.(5.53) with $P_{F,_{AAF}-p}$ in Eq.(5.57) and $P_{F_{SD}^{(n)},AAF-p}$ in Eq.(5.40) with $P_{F_{SD}^{(n)},_{DAF}^{ARP}-p}$ in Eq.(5.55), it is obvious that $P_{F,_{DAF}^{ARP}-p} \leq P_{F,_{AAF}-p}$ and $P_{F_{SD}^{(n)},AAF-p} \leq P_{F_{SD}^{(n)},_{DAF}^{ARP}-p}$ for each item in the above two FER expressions. So the ARP-DAF can achieve a smaller error rate compared to AAF under the same γ_{sr} and γ_{rd} values, such as

$$P^{(n)}_{FER,{}^{ARP}_{DAF}} \leq P^{(n)}_{FER,AAF}. \tag{5.60}$$

Furthermore, the $P_{F,{}^{ARP}_{DTC}}$ and $P_{F,_{AAF}}$ have a similar relationship:

$$P^{(n)}_{FER,{}^{ARP}_{DTC}} \leq P^{(n)}_{FER,AAF}. \tag{5.61}$$

From the above analysis, it has been shown that the HARQ-ARP with DTC performs better than the HARQ-AAF.

5.7 Simulation Results

In this section, we present the performance comparison for various HARQ schemes. In all simulations, data are grouped into a frame of length 130 symbols, including 16 CRC check symbols and 2 tail symbols. A four-state rate 1/2 RSC code with generators $(5,7)_8$ is used with BPSK modulation. At the receiver, a soft output Viterbi algorithm (SOVA)/Viterbi algorithm (VA) decoder is used. The maximum retransmission number is set to $n = 3$ and the maximum number of iterations is set to 8. The direct and relay channels have the same SNR, which varies from 0 to 14 dB, while the inter-user channel is variable from 0 to 24 dB.

In order to present equitable comparisons, we apply the persistent transmission principle at the relay. Since the relay always transmits in the HARQ AAF and HARQ ARP-DAF/DTC schemes, so the relay in the HARQ DAF-DTC scheme transmits even though the received packet from the source is decoded in error.

Fig. 5.1 compares the FER results obtained by the analysis developed in this thesis and simulations. The FER results are presented for the HARQ I-perfect DTC scheme, which is used as the performance lower bound for the proposed HARQ I-ARP scheme. The analytical FER results are calculated for the first term in the bound sum in Eq.(5.48), and for asymptotic values of γ_{sd}, γ_{rd}, for which the number of paths $\sum_{d^{(n)}=d^{(n)},min}^{4l} \overline{A}(d^{(n)})$ has not a big effect on the performance and thus was assumed to be 1. The two curves have the same slope and the match could be improved by calculating the coefficient $\sum_{d^{(n)}=d^{(n)},min}^{4l} \overline{A}(d^{(n)})$ and including more levels from the expression in Eq.(5.48).

Fig. 5.2 shows the FER comparison of various HARQ I schemes. As an example, it is shown that the ARP-DAF outperforms the AAF in the whole γ_{rd} region, and the performance of the ARP-DTC is much better than the ARP-DAF due to an addtional coding gain with the DTC structure. The simulation results also prove our analysis in section 5.5, that is, the ARP-DTC scheme can achieve a high error rate reduction compared to AAF under the same γ_{sr} and γ_{rd} values. In addition, we can see from Eqs.(5.42) and (5.47) that as γ_{sr} increases, $\theta\gamma S_p$ and C_p come close to one. So $P_{FER,ARP}^{(n)} \rightarrow P_{FER,Perfect}^{(n)}$, thus the ARP-DTC approaches the perfect DTC scheme. This trend can be noticed from the simulation results shown in these two figures.

Figs. 5.3, 5.4, 5.5 and 5.6 compare the FER performance of various HARQ II and HARQ III schemes with the two RCPT families we introduced in sections 4.3.2 and 4.3.3. Similar to the HARQ I scheme, the ARP-DTC's performance is significantly better than the AAF's performance in the whole γ_{rd} region. When the γ_{sr} increases, the FER of ARP-DTC comes close to the perfect DTC as well.

Figs. 5.7, 5.8, 5.9 and 5.10 show the throughput performance of the HARQ I, HARQ II and HARQ III schemes with 8 and 24 dB of the inter-user channel. The simulation results confirm our performance analysis in sections 5.5 and 5.6. We can see that the ARP with DTC scheme provides a better throughput than the AAF scheme. The throughput increases as the γ_{sd} increases, and as the γ_{sr} increases to 24 dB, the performance of ARP-DTC tends to reach the perfect-DTC.

Fig. 5.11 compares the throughput of various HARQ ARP-DTC schemes. In a relatively high SNR range, the destination can correctly decode the highest rate packets at the first transmission and thus reduce the number of retransmissions. Therefore, the HARQ II ARP-DTC achieves a higher throughput efficiency than the other HARQ ARP-DTC schemes. Also, the throughput of the HARQ III ARP-DTC scheme is better than that of the HARQ I ARP-DTC scheme. This is because in the type III scheme the packet with punctured symbols is transmitted, which reduces the total symbols of transmission. However, at low SNRs, in the range of 0-4 dB, the HARQ I ARP-DTC scheme outperforms the other two schemes, due to its high SNR gain.

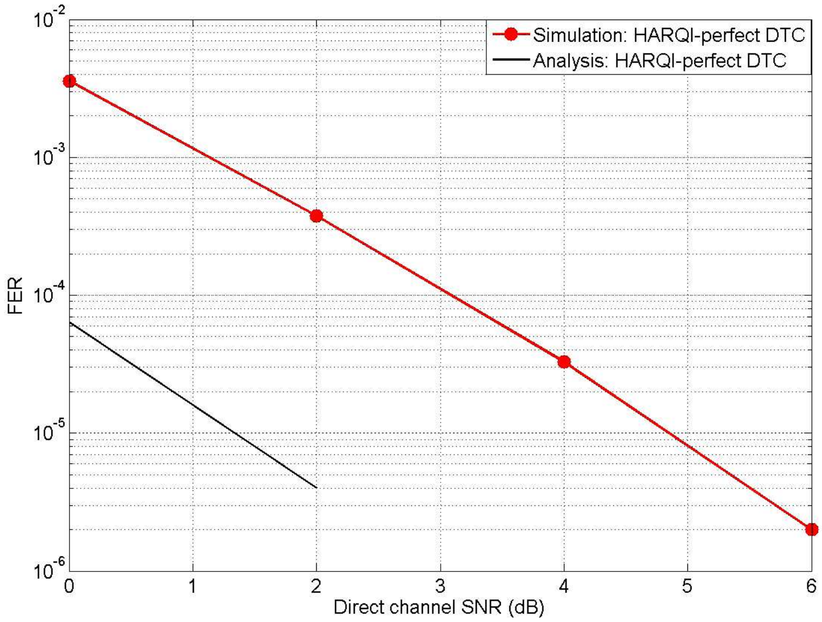

Figure 5.1: Comparisons between FER performance based on analysis and simulations for HARQ I-perfect DTC, which is used as the performance lower bound for the proposed HARQ I-ARP scheme

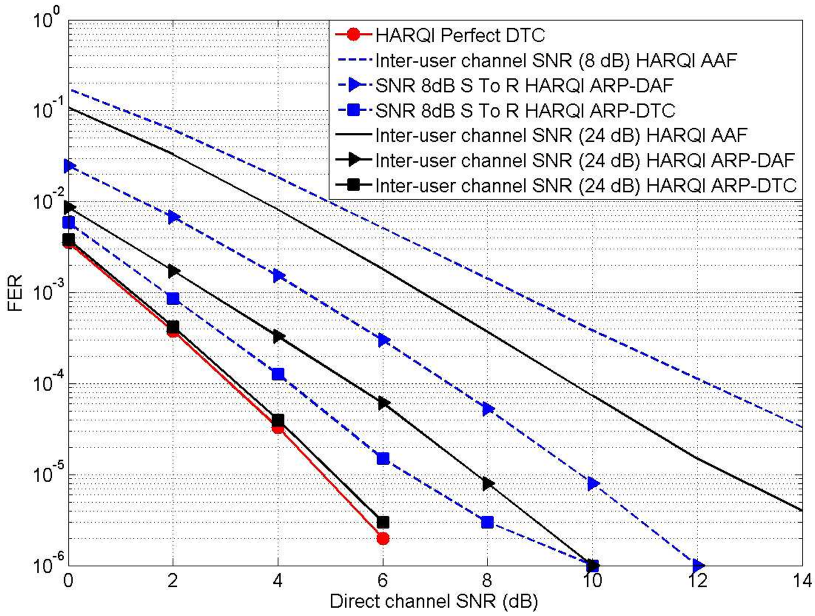

Figure 5.2: FER comparison of HARQ I AAF, HARQ I perfect-DTC, HARQ I ARP-DAF and HARQ I ARP-DTC schemes in a quasi-static fading channel with SNR 8 dB and 24 dB of the inter-user channel

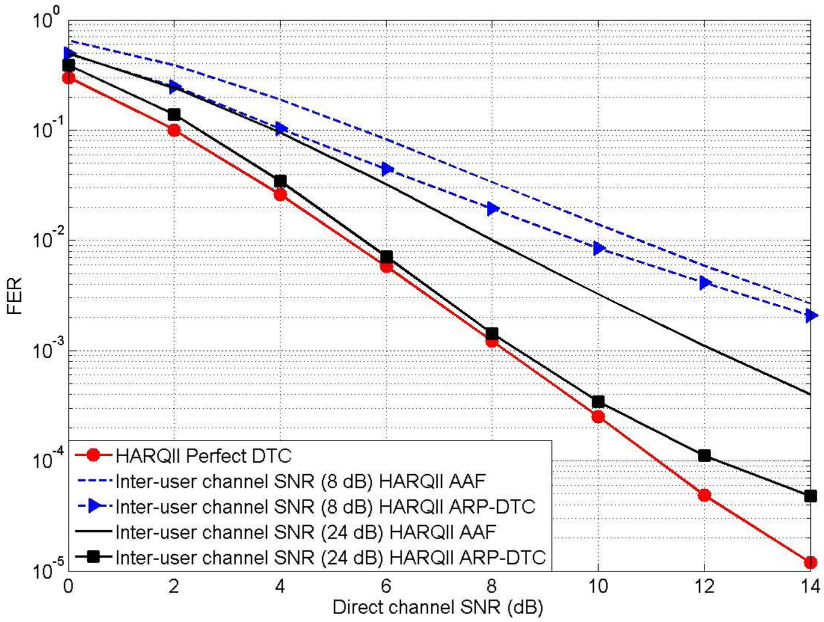

Figure 5.3: FER comparison of HARQ II AAF, HARQ II perfect-DTC, and HARQ II ARP-DTC schemes in a quasi-static fading channel with SNR 8 dB and 24 dB of the inter-user channel; the puncturing rates for the first transmission are 4/5

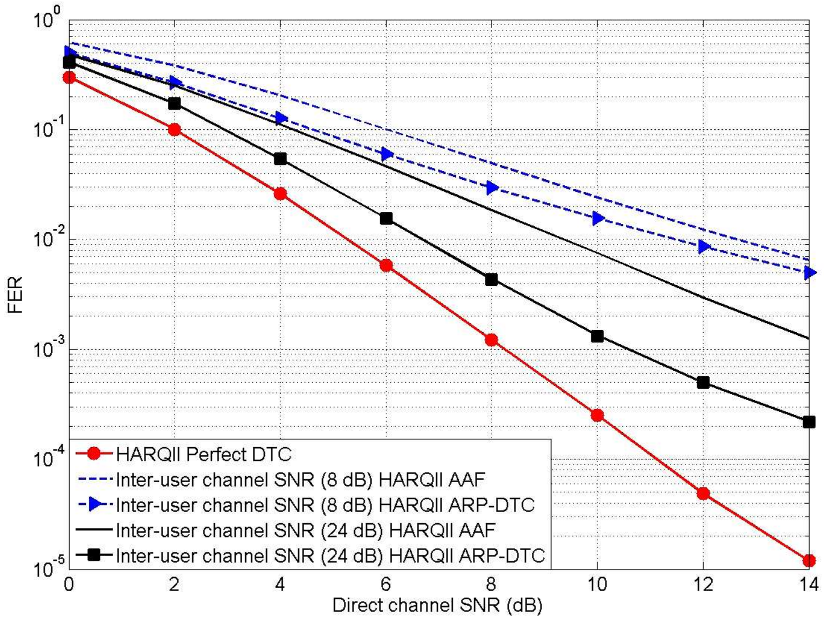

Figure 5.4: FER comparison of HARQ II AAF, HARQ II perfect-DTC, and HARQ II ARP-DTC schemes in a quasi-static fading channel with SNR 8 dB and 24 dB of the inter-user channel; the puncturing rates for the first transmission are 2/3

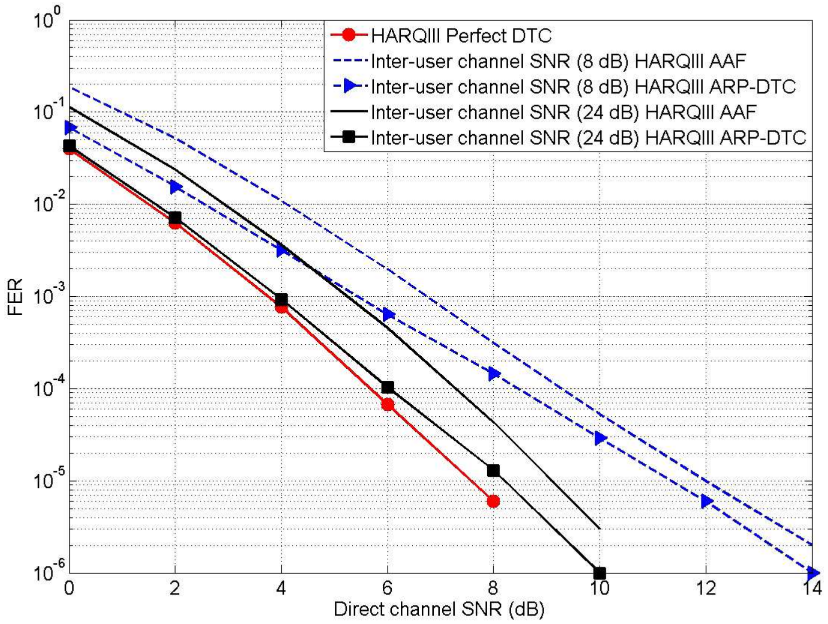

Figure 5.5: FER comparison of HARQ III AAF, HARQ III perfect-DTC, and HARQ III ARP-DTC schemes in a quasi-static fading channel with SNR 8 dB and 24 dB of the inter-user channel; the puncturing rates for the first transmission are 4/5

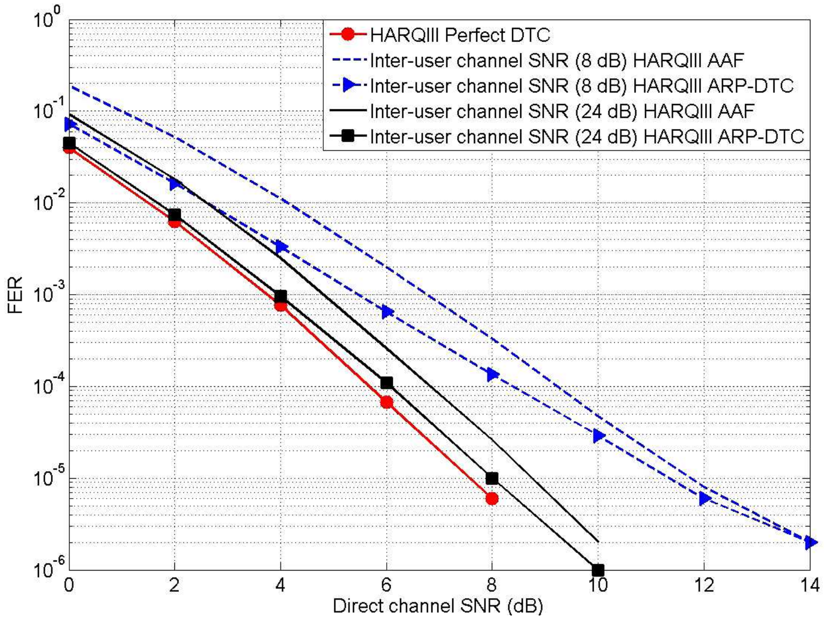

Figure 5.6: FER comparison of HARQ III AAF, HARQ III perfect-DTC, and HARQ III ARP-DTC schemes in a quasi-static fading channel with SNR 8 dB and 24 dB of the inter-user channel; the puncturing rates for the first transmission are 2/3

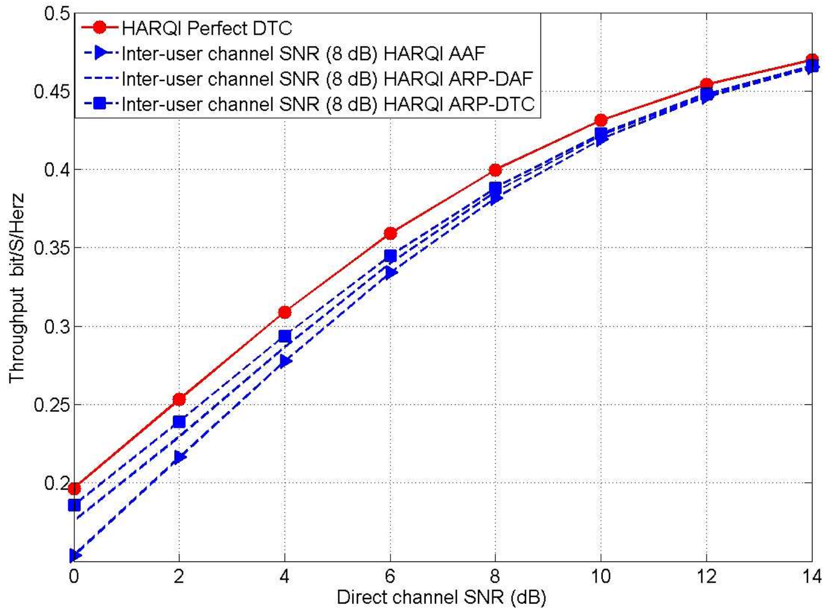

Figure 5.7: Throughput comparison of HARQ I AAF, HARQ I perfect-DTC, HARQ I ARP-DAF and HARQ I ARP-DTC schemes in a quasi-static fading channel with SNR 8 dB of the inter-user channel

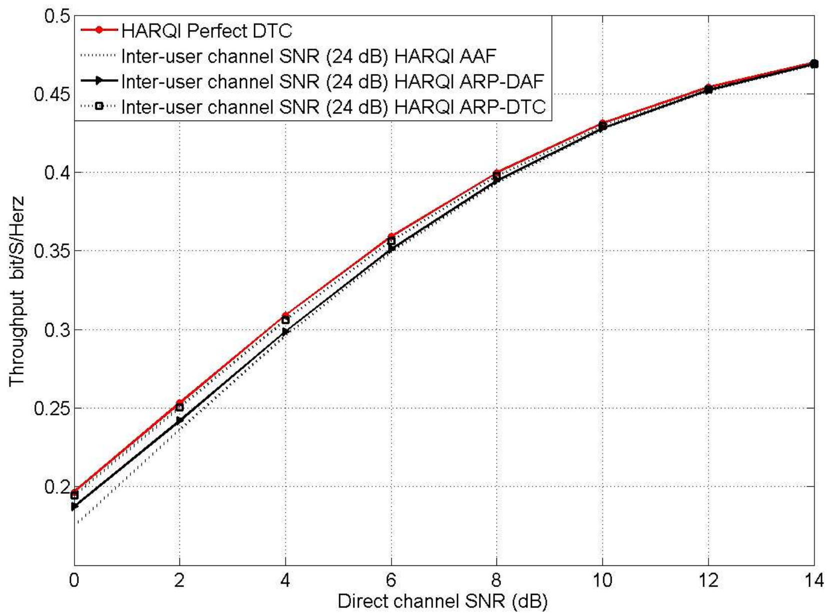

Figure 5.8: Throughput comparison of HARQ I AAF, HARQ I perfect-DTC, HARQ I ARP-DAF and HARQ I ARP-DTC schemes in a quasi-static fading channel with SNR 24 dB of the inter-user channel

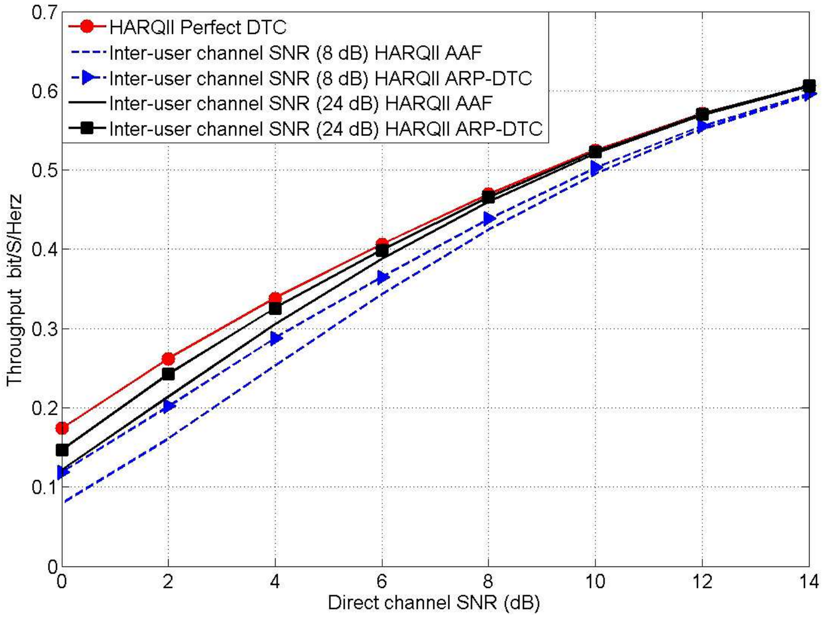

Figure 5.9: Throughput comparison of HARQ II AAF, HARQ II perfect-DTC, and HARQ II ARP-DTC schemes in a quasi-static fading channel with SNR 8 dB and 24 dB of the inter-user channel; the puncturing rates for the first transmission are 4/5

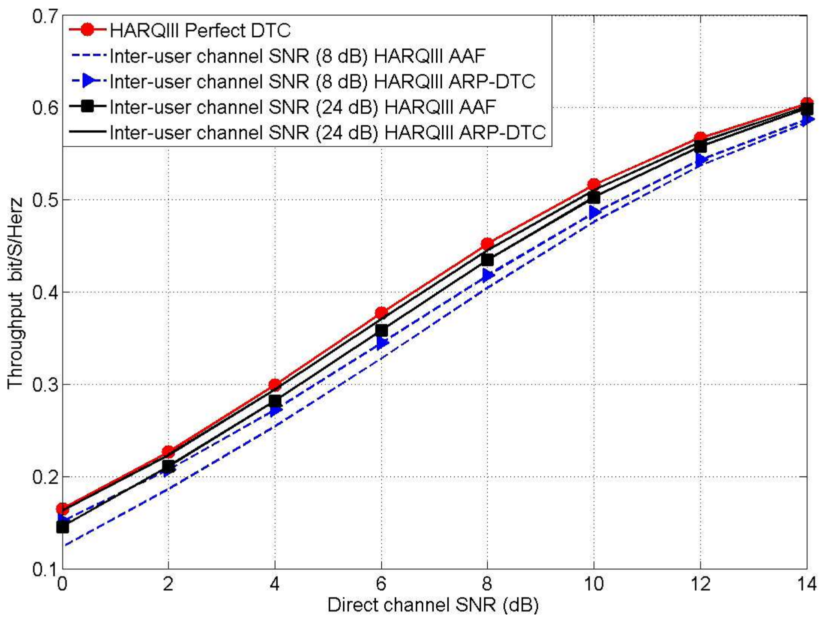

Figure 5.10: Throughput comparison of HARQ III AAF, HARQ III perfect-DTC, and HARQ III ARP-DTC schemes in a quasi-static fading channel with SNR 8 dB and 24 dB of the inter-user channel; the puncturing rates for the first transmission are 4/5

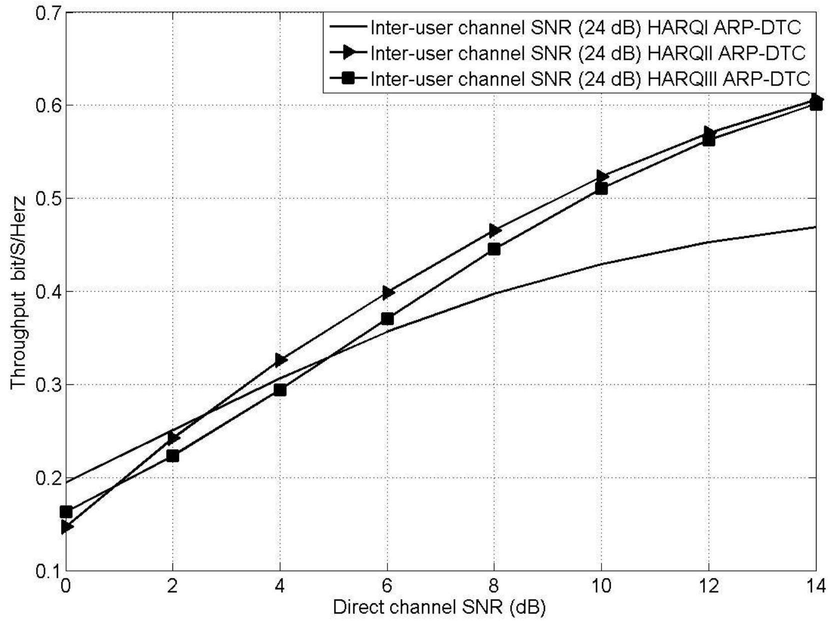

Figure 5.11: Throughput comparison of HARQ ARP schemes in a quasi-static fading channel with SNR 24 dB of the inter-user channel; the puncturing rates for the first transmission are 4/5

5.8 Conclusion

In this chapter, we investigated the performance of three types of HARQ ARP schemes. The exact WEPs for HARQ AAF, HARQ perfect-DTC and HARQ ARP were derived and the general throughput expression of each relaying scheme was developed. Based on the analysis, it was shown that each type of HARQ ARP outperforms the HARQ AAF in the whole γ_{rd} region, due to the contribution of the turbo coding gain. As the γ_{sr} increases with the inter-user channel, the performance of the HARQ ARP approaches the HARQ perfect-DTC scheme. In addition, we showed, from the throughput comparison of the three types of HARQ ARP schemes, that the HARQ II ARP achieved higher throughput in a relatively higher SNR range.

Chapter 6

Conclusions

This thesis has studied collaborative HARQ schemes for the cooperative wireless networks. The main goals were to provide a better understanding of HARQ protocols used in the cooperative communication systems and develop a robust HARQ scheme by using an adaptive relaying protocol and distributed coding scheme.

Chapters 1, 2 and 3 provided the background knowledge. Specifically, Chapter 1 gave a broad overview of the research field, aiming to explain the goals of this work and outline the remaining chapters. Chapter 2 covered the basic elements of a digital communication system, fading channels, classic error control coding schemes as well as associated decoding algorithms. Chapter 3 introduced the knowledge of the cooperative communications. We reviewed the various relaying protocols used in the cooperative diversity (CD) communication systems, including amplify and forward (AAF), decode and forward (DAF), selection relaying and incremental relaying protocols.

Chapter 4 presented our proposed collaborative HARQ strategies for the CD communication systems. An adaptive relaying protocol (ARP) was proposed and was used together with type I, II, and III HARQs. The proposed ARP takes merits of the fixed relaying protocols and avoids their disadvantages. By using distributed coding scheme in the proposed scheme, we improved the reliability of the systems through not only diversity benefit but also a coding advantage. The simulation results indicated that the proposed HARQ with ARP (HARQ-ARP) scheme outperforms the reference HARQ schemes in all SNR regions by $1 \sim 6$ dB.

Chapter 5 provided the performance analysis of the proposed HARQ-ARP scheme. We analyzed the performance of the HARQ relaying schemes by two figures of merit: the average throughput R_{AV} and frame error rate (FER). We derived the pairwise error probabilities (PEP) and word error probabilities (WEP) of the HARQ with AAF (HARQ-AAF), HARQ with perfect distributed turbo codes (HARQ-perfect DTC) and HARQ-ARP schemes. Based on the derived PEPs and WEPs for each scheme, we obtained a general throughput expression, which can be used for any of these schemes. In addition, we derived a general FER expression for the collaborative HARQ systems and used it to compare between the proposed HARQ-ARP scheme and the reference schemes. The theoretical analysis is validated through simulations. The analytical and simulation results show that the HARQ-ARP scheme achieves better performance than HARQ-AAF scheme. As the quality of the inter-user channel is improved, the performance of HARQ-ARP approaches the HARQ-perfect DTC scheme.

6.1 Future Work

In this thesis, we only consider a single relay network. In practical systems, there might be a number of relays between the source and destination. To the best of our knowledge, no general framework for designing collaborative HARQ protocols for multi-hop relay networks has been developed. The design of collaborative HARQ schemes includes several elements, such as, component codes design, optimal puncturing pattern, path selection and power allocation. They can be optimized by using union bound and code distance spectra. This presents a big challenge and will be a promising topic to be considered in future research.

In addition, most existing research in HARQ has considered an incremental redundancy (IR) scheme to provide a lower code rate and a higher throughput. Although the use of an IR scheme is a promising solution to achieve a higher throughput, its decoding is very complex. The complexity of such a decoding grows at least as $O(k/R)$, where k is the information bits and R is the rate of the low rate code [85]. Luby [86] circumvented this problem by designing rateless codes, also known as Luby Transform (LT) codes, which are not obtained by puncturing standard block codes. Unlike conventional codes, LT codes encode and transmit the source information in an infinitely long codestream. The codes have the special property that a receiver can recover the original information from unordered subsets of the codestream, once the total obtained mutual information from multiple sources marginally exceeds the entropy of the source information. LT codes have been suggested for use in single relay links [87, 88], and broadcast and multicast applications [89] in wireless networks, however, their use in cooperative multi-relay wireless networks has not been analyzed yet. Therefore, it would also be worth investigating the performance of the LT codes on a multi-hop CD communication system.

Appendix A

The Derivation of Eq. (5.18)

To simplify the calculation, we use x to denote $\gamma_{sr}|h_{sr}|^2$, y to denote $\gamma_{rd}|h_{rd}|^2$, respectively. The pdf of x and y can be calculated as [33]

$$p(x) = \frac{1}{\gamma_{sr}} e^{\frac{-x}{\gamma_{sr}}}, \tag{A.1}$$

$$p(y) = \frac{1}{\gamma_{rd}} e^{\frac{-y}{\gamma_{rd}}}, \tag{A.2}$$

$$\begin{aligned} f(d_1) &\approx E\left[\sum_{d_s r = d_{sr,min}}^{2l} \overline{A}(d_{sr}) Q\left(\sqrt{2d_{sr}\gamma_{sr}|h_{sr}|^2}\right) Q\left(\sqrt{2d_1 min\{|h_{sr}|^2\gamma_{sr}, \gamma_{rd}|h_{rd}|^2\}}\right)\right] \\ &= E\left(\sum_{d_s r = d_{sr,min}}^{2l} \overline{A}(d_{sr})\varphi(d_1)\right), \end{aligned} \tag{A.3}$$

where

$$
\begin{aligned}
\varphi(d_1) &\leq \int_0^\infty \int_0^y Q\left(\sqrt{2d_{sr}x}\right) Q\left(\sqrt{2d_1 x}\right) p(x)d(x)p(y)dy \\
&+ \int_0^\infty \int_0^x Q\left(\sqrt{2d_{sr}x}\right) Q\left(\sqrt{2d_1 y}\right) p(y)d(y)p(x)dx \\
&\leq \int_0^\infty \int_0^y e^{(-d_{sr}x)} e^{(-d_1 x)} \frac{1}{\gamma_{sr}} e^{\frac{-x}{\gamma_{sr}}} dxp(y)dy + \int_0^\infty \int_0^x e^{(-d_{sr}x)} e^{(-d_1 y)} \frac{1}{\gamma_{rd}} e^{\frac{-y}{\gamma_{rd}}} dyp(x)dx \\
&\leq \frac{1}{\gamma_{sr}} \int_0^\infty \int_0^y e^{\left[-\left(d_{sr}+d_1+\frac{1}{\gamma_{rd}}\right)x\right]} dxp(y)dy + \frac{1}{\gamma_{rd}} \int_0^\infty e^{(-d_{sr}x)} \int_0^x e^{\left[-\left(d_1+\frac{1}{\gamma_{rd}}\right)y\right]} dyp(x)dx \\
&\leq \frac{1}{\gamma_{sr}} \int_0^\infty \frac{-1}{d_{sr}+d_1+\frac{1}{\gamma_{rd}}} e^{\left[-\left(d_{sr}+d_1+\frac{1}{\gamma_{rd}}\right)x\right]_0^y} p(y)dy \\
&+ \frac{1}{\gamma_{rd}} \int_0^\infty e^{-ax} \left[\frac{-\gamma_{rd} e^{-\left(d_1+\frac{1}{\gamma_{rd}}\right)y}}{1+d_1\gamma_{rd}}\right]_0^x p(x)dx \\
&\leq \frac{1}{\gamma_{sr}} \int_0^\infty \frac{-1}{d_{sr}+d_1+\frac{1}{\gamma_{sr}}} \left[\left(e^{-\left(d_{sr}+d_1+\frac{1}{\gamma_{sr}}\right)y} - 1\right) \frac{1}{\gamma_{rd}} e^{\frac{-y}{\gamma_{rd}}} dy\right] \\
&+ \frac{1}{\gamma_{rd}} \int_0^\infty e^{-ax} \left(\frac{-\gamma_{rd}}{1+d_1\gamma_{rd}}\right) \left[e^{-\left(d_1+\frac{1}{\gamma_{rd}}\right)x} - 1\right] \frac{1}{\gamma_{rd}} e^{\frac{-x}{\gamma_{sr}}} dx \\
&\leq \frac{1}{\gamma_{sr}\gamma_{rd}} \left[\frac{-1}{d_{sr}+d_1+\frac{1}{\gamma_{sr}}} \left(\frac{-1}{d_{sr}+d_1+\frac{1}{\gamma_{sr}}+\frac{1}{\gamma_{rd}}} \frac{1}{e^{\left(d_{sr}+d_1+\frac{1}{\gamma_{sr}}+\frac{1}{\gamma_{rd}}\right)y}} + \gamma_{rd} e^{\frac{-y}{\gamma_{rd}}}\right)\right]_0^\infty \\
&+ \frac{1}{\gamma_{sr}\gamma_{rd}} \left[\frac{-\gamma_{rd}}{1+d_1\gamma_{rd}} \left(\frac{-1}{d_{sr}+d_1+\frac{1}{\gamma_{sr}}+\frac{1}{\gamma_{rd}}}\right) \frac{1}{e^{\left(d_{sr}+d_1+\frac{1}{\gamma_{sr}}+\frac{1}{\gamma_{rd}}\right)x}}\right. \\
&+ \left.\frac{1}{d_{sr}+\frac{1}{\gamma_{sr}}} \frac{1}{e^{\left(d_{sr}+\frac{1}{\gamma_{sr}}\right)}}\right]_0^\infty \\
&\leq \frac{1}{\gamma_{sr}\gamma_{rd}} \left[\frac{\left(d_{sr}+d_1+\frac{1}{\gamma_{sr}}\right)\gamma_{rd}}{\left(d_{sr}+d_1+\frac{1}{\gamma_{sr}}\right)\left(d_{sr}+d_1+\frac{1}{\gamma_{sr}}+\frac{1}{\gamma_{rd}}\right)}\right] \\
&+ \frac{1}{\gamma_{sr}\gamma_{rd}} \left[\frac{-\gamma_{rd}}{1+d_1\gamma_{rd}} \frac{-\left(d_1+\frac{1}{\gamma_{rd}}\right)}{\left(d_{sr}+d_1+\frac{1}{\gamma_{sr}}+\frac{1}{\gamma_{rd}}\right)\left(d_{sr}+\frac{1}{\gamma_{sr}}\right)}\right] \\
&\leq \frac{1}{\gamma_{sr}}\left(\frac{1}{d_{sr}+d_1}\right) + \frac{1}{\gamma_{sr}\gamma_{rd}d_{sr}}\left(\frac{1}{d_{sr}+d_1}\right).
\end{aligned}
\tag{A.4}
$$

By substituting the result of Eq.(A.4) into Eq.(A.3), we can obtain the final result as in

Eq.(5.18),

$$f(d_1) \approx E\left\{ \sum_{d_s r = d_{sr,min}}^{2l} \overline{A}(d_{sr}) \frac{1}{\gamma_{sr}} \left[\frac{1}{(d_{sr}+d_1)} + \frac{1}{d_{sr}\gamma_{rd}\,(d_1+d_{sr})} \right] \right\}. \qquad \text{(A.5)}$$

Appendix B

Proof of Inequality (5.54)

In this section, we only show the proof of inequality of $R^{'}_{1-I}$ at the first transmission attempt in the HARQI scheme. The n^{th} transmission expressions of $R^{'}_{1}, R^{'}_{2}$ and $R^{'}_{3}$ in HARQI, HARQII and HARQIII schemes can be proved in a similar way.

As shown in Eq.(5.14), $P^{(1)}_{F,sr-I}$ represents the conditional WEP in the inter-user channel at the first transmission attempt, therefore, we have

$$P^{(1)}_{F,sr-I} = \sum_{d_{sr}=d_{sr,min}}^{2l} \overline{A}(d_{sr}) E\left[Q\left(\sqrt{2d_{sr}\gamma_{sr}|h^{(1)}_{sr}|^2}\right)\right] \leq 1. \tag{B.1}$$

Then we arrive at

$$P^{(1)}_{F,sr-I} Q\left(\sqrt{2d_1\gamma_{sr}|h^{(1)}_{sr}|^2}\right) \leq Q\left(\sqrt{2d_1\gamma_{sr}|h^{(1)}_{sr}|^2}\right). \tag{B.2}$$

To simplify the calculation, we still use x denote $\gamma_{sr}|h_{sr}|^2$. By averaging the above inequality with regard to $|h_{sr}|^2$, we get

$$\begin{aligned}
\int_0^\infty \sum_{d_{sr}=d_{sr,min}}^{2l} \overline{A}(d_{sr}) Q\left(\sqrt{2d_{sr}x}\right) Q\left(\sqrt{2d_1 x}\right) &\leq \int_0^\infty Q\left(\sqrt{2d_1 x}\right) \\
\Rightarrow \int_0^\infty \sum_{d_{sr}=d_{sr,min}}^{2l} \overline{A}(d_{sr})(e^{-d_{sr}x})(e^{-d_1 x})\frac{1}{\gamma_{sr}}(e^{-\frac{x}{\gamma_{sr}}})dx &\leq \int_0^\infty (e^{-d_1 x})\frac{1}{\gamma_{sr}}(e^{-\frac{x}{\gamma_{sr}}})dx \\
\Rightarrow \sum_{d_{sr}=d_{sr,min}}^{2l} \overline{A}(d_{sr})\frac{1}{d_{sr}+d_1} &\leq \frac{1}{d_1} \\
\Rightarrow \sum_{d_{sr}=d_{sr,min}}^{2l} \overline{A}(d_{sr})\frac{d_1}{d_{sr}+d_1} &\leq 1.
\end{aligned} \tag{B.3}$$

Therefore, we have $R'_{1-I} \leq 1$.

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