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Comparison of Coded Diversity Combining Techniques in Packetized Digital Wireless Systems

By

Richard O. Farley

A thesis submitted to
the Faculty of Graduate Studies and Research
in partial fulfillment of
the requirements for the degree of

Master of Engineering

Ottawa-Carleton Institute for Electrical Engineering
Department of Systems and Computer Engineering
Carleton University
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*Comparison of Coded Diversity Combining Techniques in
Packetized Digital Wireless Systems*

submitted by Richard O. Farley
in partial fulfillment of the requirements for
the degree of Master of Engineering

Thesis Supervisor

Chair, Department of Systems and Computer Engineering

Carleton University
April 1, 1996

Abstract

ATM (Asynchronous Transfer Mode, a form of fast packet switching based on constant-length packets 53 bytes long) networks are being introduced as the vehicle for new telecommunications applications, both in the private and public areas, to support fixed and mobile, narrowband and broadband communications. When internetworking is examined the wireless connections become the bottleneck because of unreliability. Antenna diversity techniques operating on a packet basis can be used as a means to improve the reliability of wireless connections and make feasible the extension of ATM-related protocols to the wireless domain.

This thesis compares the performance of a number of packet based diversity techniques employing punctured convolutional coding and block interleaving. Results were obtained through simulations for coding without diversity, post-Viterbi-decoding selection diversity and code-diversity combining for AWGN (additive white Gaussian noise) and Rayleigh fading radio channel models. The code combining was done with and without reliability information. Code combining was also examined using a type-II hybrid ARQ protocol on a Rayleigh fading channel. The criteria for comparison of the diversity schemes was the cell loss probability (at least one uncorrectable error in an ATM cell) and the probability of bit error.

It was found that interleaving of at least 100 coded ATM cells would be required for the code combining approach to outperform the post-Viterbi decoding selection scheme in a slow fading environment in terms of cell-loss probability. This could be improved upon if more reliable weights are used in the code-combining and is suggested as future work.

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I would like to thank Professor David D. Falconer and Dr. G. Stamatelos for all their patient guidance, supervision, and ideas (without which none of this would have been possible) during the course of this thesis. I benefited greatly from our meetings and realized during the writing of this thesis the full extent of how much I had learned throughout this process.

Special thanks goes to my colleagues at BNR for their constant prodding and encouragement to complete this thesis, most notably, Kai Ko and David Dockendorff, as well as Art Dunham and Jane Stephenson.

I must also thank my parents for giving me all the opportunities in my younger years to enable me to pursue a career in the most exciting field of telecommunications. It would always boggle my mind when I would talk with my grandfather who had witnessed the advent of commercial flight, satellite communications, the landing of man on the moon, and many other marvels of technological innovation. I know that I'll be able to witness the same kind of accomplishments in the next sixty years in telecommunications.

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Introduction

1.1 Thesis Objectives

The exponential growth in cellular radio in recent years and the emergence of personal communications networks (PCNs) bears witness to the growing need to combine mobility with communications accessibility. Some estimates place the number of worldwide PCN subscribers at 100 million by the year 2000 and at 450 million by 2010. This need for mobile access extends to non-voice services such as paging and fax, and will no doubt also extend to many if not all telecommunications services made possible through ISDN and other technologies. The trend towards non-voice wireless services will be accelerated by the growing popularity and use of small but powerful portable information terminals such as notebook computers, which will be tetherless, and whose applications could include teleconferencing and other multimedia communications services. The use of such devices for these applications suggests that portability and broad-band service capabilities will be necessary. It is expected that PCNs will have their strongest initial appeal in office-type environments, but that as costs diminish, they will also be heavily used in factory, hospital, school, entertainment and home environments. [14]

The full-scale development of high-speed asynchronous transfer mode (ATM)-switched networks and the evolution towards “third-generation” portables will make the integration of wireless and optical systems not only desirable but necessary. ATM [44]-[45], a form of fast packet switching based on constant-length packets 53 bytes long (cells), is the ITU-T preferred transport technique for a public, switched, high-speed fibre-optic network capable of handling constant-rate, highly bursty or smoothly variable traffic.

The wireless link is the most expensive and vulnerable part of the path through the public network; therefore maximization of its utilization is an important issue. Traditional reliability-improving techniques that employ retransmissions on a per-link basis might induce intolerable delay or storage requirements when used over a high-speed wireless link. Instead, other techniques involving FEC (forward error correction) block/convolutional code should be used to improve the error performance and keep the processing delay to a minimum. Antenna diversity techniques can also be employed in order to

improve the reliability of service from the user's point of view by combatting multipath propagation effects and fading.

When the integration of wireless systems with existing networks is considered, a number of processes like FEC need to be streamlined and constant-length packets might be necessary. It has also been noted that a scheme that maintains the decoded packet size constant, that of the ATM cell, avoids protocol conversions, preserves bandwidth and facilitates the interconnection with the public switched ATM network,^[14] or the backbone ATM switch in a local ATM application.

This thesis explores the code combining approach^[23] in combination with the use of punctured convolutional codes in a mobile environment. The code combining approach combines the benefits of diversity with FEC. The packets used here are fixed-length ATM cells. The main advantage of this scheme is that if multiple diversity branches are available the coding overhead can be effectively distributed amongst these branches by employing this technique^[23,24]. Several authors^[26,43] have done a number of studies on the use of rate-compatible punctured codes in a fading environment with the assumption of the availability of channel-state information to estimate the metrics used in soft-decision Viterbi decoding. The code symbol errors were also assumed independent. The work done here will use limited amounts of interleaving. The use of rate-compatible punctured convolutional codes is appropriate in this environment since the coding overhead can be adapted to the channel conditions and the required quality of service. Hard-decision decoding is used throughout the simulations since the transmission rates considered are OC-3 (155 Mb/s) or greater and soft-decision decoding would not be feasible.

The objectives of this thesis are:

- to develop a simulation tool to analyse the performance (cell loss probability) of a number of packet based diversity techniques employing punctured convolutional coding and block interleaving.
- to use this tool to compare the cell loss probability obtained for post-detection selection diversity and code-diversity combining for :
  - (i) a two- and three-branch additive white Gaussian noise radio channel model,

(ii) a two- and three-branch Rayleigh fading radio channel model under the following conditions: slow fading without interleaving, slow fading with limited interleaving, fast fading, slow fading with correlated diversity branches

(iii) a lognormal shadowing channel with Rayleigh fading superimposed

(iv) a hybrid ARQ/FEC protocol on a Rayleigh fading channel using rate compatible punctured convolutional codes.

The superiority of the code combining approach over a post detection approach was demonstrated over an AWGN channel.

When the diversity schemes were compared over fading conditions, it was seen that interleaving of at least 100 ATM cells would be required to obtain a substantial coding gain for the code-combining approach. Depending on the application this may introduce excessive delay ^[14], and a simple ARQ scheme might be preferable. When the diversity schemes were examined over correlated diversity branches the coding gain derived from the equal weights code-combining approach was superior as the correlation coefficient was increased. This suggested a method to accurately determine the reliability weights to be used when code-combining the diversity branches. This led to the examination of a hybrid-ARQ scheme in which the weights for each transmitted packet were derived using the "SOVA"^[19,20] approach.

1.2 Thesis Outline

This chapter concludes with the outlook of current wireless systems, and is followed by the four main chapters and a concluding chapter. Chapter two addresses channel coding applied for wireless channels including background, diversity combining techniques, and radio channel characteristics.

Chapter three examines the performance of packet-diversity techniques in additive white Gaussian noise. Chapter four examines packet-diversity techniques in fading conditions with and without interleaving. The effect of correlated diversity branches is also examined. The performances is also briefly examined for shadowing with superimposed Rayleigh fading; and chapter five examines some hybrid-ARQ schemes.

The last chapter deals with the concluding remarks and suggested future work.

1.3 Wireless Data Communications

Wireless local-area data networks (WLANs) can be categorized as providing low-mobility high-speed data communications within a confined region, e.g., a campus or a large building. Coverage range from a wireless data terminal is short, tens to hundreds of feet, like cordless telephones. Coverage is limited to within a room or to several rooms in a building. WLANs have been evolving for a few years, but overall, the situation is chaotic, with many different products being offered by many different vendors. There is no stable definition of the needs or design objectives for WLANs, with data rates ranging from hundreds of kb/s to more than 10 MB/s, and with several products providing one or two MB/s wireless link rates. A partial list of some advertised products is given in Table 1.1. Users of WLANs are not nearly as numerous as users of more voice-oriented wireless systems. Part of the difficulty stems from these systems being driven by the computer industry that views the wireless system as just another plug-in interface card, without giving sufficient consideration to the vagaries and needs of a reliable radio system. [29]

Table 1.2 lists some existing wireless system standards and technologies. Important differences between these present systems and future ones is the integration of voice, data, and video services and higher data rates, up to 155.52 Mbps (OC-3 SONET rate).

Universal access to "broad-band" services, requiring bit rates above about 1 Mb/s, will be provided to customers through broadband ISDN (B-ISDN) and broadband LANs. Wireless access to B-ISDN services will require much wider RF bandwidths than those available below 3 GHz. The Broadband Indoor Wireless Communications Project of CITR (Canadian Institute for Telecommunications Research) is investigating the 20-60 GHz range of frequencies (millimetre waves) for indoor wireless systems that could support a variety of broad-band and narrow-band ISDN. It is to be capable of supporting ATM at transport bit rates up to about 160 Mb/s for broadband LANs. The overall system is shown in Figure 1.1.

Table 1.3 shows some possible broad-band wireless services ^{[46]-[49]} with estimated bit rate (in Mb/s) and performance requirements.

TABLE 1.1. Partial List of WLAN Products [29]

Product Company Location	Freq. (MHz)	Link rate	User rate	Protocol(s)	Access	No. of chan. or spread factor	Mod./coding	Power	Network Topology
Altair Plus II Motorola Arlington Hts., IL	18-19 GHz	15 Mb/s	5.7 Mb/s	Ethernet			4-level FSK	25 mW peak	Eight devices/ radio; radio to base to Ethernet
WaveLAN NCR/AT&T Dayton, OH	902-928	2 Mb/s	1.6 Mb/s	Ethernet-like	DS SS		DQPSK	250 mW	Peer-to-peer
AirLAN Soletek San Diego, CA	902-928		2 Mb/s	Ethernet	DS SS		DQPSK	250 mW	PC/MCA w/ ant; radio to hub
Freeport Windata Inc. Northboro, MA	902-928	16 Mb/s	5.7 Mb/s	Ethernet	DS SS	32 chips/bit	16 PSK trellis coding	650 mW	Hub
Intersect Perisoft Inc. Madison, WI	902-928		2 Mb/s	Ethernet, token ring	DS SS		DQPSK	250 mW	Hub
LAWN O'Neill Comm. Horsham, PA	902-928		38.4 kb/s	AX.25	SS	20 users/chan., max. 4 chan.		20 mW	Peer-to-peer
WiLAN Wi-LAN Inc. Calgary, Alberta	902-928	20 Mb/s	1.5 Mb/s/ chan.	Ethernet, token ring	CDMA/ TDMA	3 chan. 10-15 links each	"unconven- tional"	30 mW	Peer-to-peer
RadioPort ALPS Electric USA	902-928		242 kb/s	Ethernet	SS	7/3 channels		100 mW	Peer-to-peer
AirLAN 600 Telesys SLW Don Mills, Ont	902- 928, 2.4 GHz		1.35 Mb/s	Ethernet	SS			1 W max	PC's with ant.; radio to hub
RadioLink Cal. Microwave Sunnyvale, CA	902- 928, 2.4 GHz	250 kb/s	64 kb/s		FH SS	250 ms/hop 500 kHz space			Hub
Range LAN Proxim, Inc Mountainview, CA	902-928		242 kb/s	Ethernet, token ring	DS SS	3 chan.		160 mW	
RangeLAN2 Proxim, Inc Mountainview, CA	2.4 GHz	1.6 Mb/s	50 kb/s max.	Ethernet, token ring	FH SS	10 chan. (w/ 5 kb/s; 15 sub- ch. each		100 mW	Peer-to-peer bridge
Netwave Xircom Calais, ME, CA	2.4 GHz	1 Mb/s/ adaptor		Ethernet, token ring	FH SS	82 1-MHz chan. or "hops"			Hub
Freelink Cabletron Sys. Rochester, NH	2.4 and 5.8 GHz		5.7 Mb/s	Ethernet	DS SS	32 chips/bit	16 PSK trellis coding	100 mW	Hub

TABLE 1.2. Wireless PCS technologies

	High Power Systems					Low Power Systems			
	Digital Cellular (High Tier PCS)					Low Tier PCS		Digital Cordless	
	IS-54	IS-95 (DS)	GSM	DCS-1800	WACS/ PACS	Handi-Phone	DECT		
Multiple access	TDMA/ FDMA	CDMA/ FDMA	TDMA/ FDMA	TDMA/ FDMA	TDMA/ FDMA	TDMA/ FDMA	TDMA/ FDMA	FDMA	CT-2
Freq. band (MHz)	869-894	869-894	935-960	1710-1785	Emerg.	1895-	1880-	864-868	
Uplink (MHz)	824-849	824-849	890-915	1805-1880	Tech.	1907	1900	(Eur. and Asia)	
Downlink (MHz)	(USA)	(USA)	(EUR.)	(UK)	(USA)	(Japan)	(Eur.)		
RF ch. spacing						300	1728	100	
Downlink (KHz)	30	1250	200	200	300				
Uplink (KHz)	30	1250	200	200	300				
Modulation	$\pi/4$ DQPSK	BPSK/QPSK	GMSK	GMSK	$\pi/4$ DQPSK	$\pi/4$ DQPSK	GFSK	GFSK	
Portable Txmit Power, max./avg.	600 mW/ 200 mW	600 mW	1 W/125 mW	1 W/125 mW	200 mW/25 mW	80 mW/10 mW	250 mW/ 10 mW	10 mW/5 mW	
Speech Coding	VSELP	QCELP	PRE-LTP	PRE-TLP	ADPCM	ADPCM	ADPCM	ADPCM	
Speech Rate (kb/s)	7.95	8 (var.)	13	13	32/16/8	32	32	32	
Speech ch./RF ch.	3	-	8	8	8/16/32	4	12	1	
Ch. bit rate (kb/s)						384	1152	72	
Uplink (kb/s)	48.6		270.833	270.833	384				
Downlink (kb/s)	48.6		270.833	270.833	384				
Ch. coding	1/2 rate conv.	1/2 rate fwd 1/3 rate rev.	1/2 rate conv.	1/2 rate conv.	CRC	CRC	CRC (control)	None	
Frame (ms)	40	20	4.615	4.615	2.5	5	10	2	

Many of these proposed broadband services have variable bit rates and present bursty traffic loads to the network. The estimated rates shown in the table are after compression signal processing has been applied. High-resolution still images and documents, after compression, may have a data volume on the order of 8 to 10 Mb per image, and will present very bursty loads to the network in many applications. ATM is likely to be the preferred transport approach for broadband wireless systems accessing all these services, just as it is for fibre-based transport. ATM permits a wide range of services with different

bit rates to share wide-band transmission and switching facilities. In particular, narrow-band services would coexist with the broadband services of Table 1.1 in an ATM-based system.^[14]

There are also real-time applications requiring immediate delivery and non-real-time applications such as file transfer, whose delivery can be delayed. A "raw" packet loss rate

of around 10^{-3} may be acceptable in most applications^[47], given that the actual loss rate will be reduced by high-level protocols.

TABLE 1.3. Broadband wireless services and their estimated performance requirements^[14]

Service	Average rate (Mb/s)	Peak rate (Mb/s)	Typical session duration	Max. packet loss rate
Video teleconferencing*	0.3-2	10	30 min	10^{-5}
Video telephony*	0.06-2	6	3 min	10^{-5}
High-volume transfer (e.g., CAD)	1-20	1-20	10-100 Mb	10^{-9}
High-resolution image retrieval (e.g. medical applications)	4-45	4-45	8 Mb	10^{-9}
Video entertainment* (e.g., conventional and HDTV)	3-32	150	15 min-2h	10^{-7}
Portable multimedia	1-10	10-100	30 min	10^{-9}
Manufacturing (robotics)	0.01-3	0.01-3	continuous	10^{-9}
* indicates service with variable bit rate				

If the wireless link is going to support broadband services then its reliability and performance should be compatible (from a user's perspective) with that of the fiber link.

The following chapter will provide the needed background on channel coding as applied to wireless channels.

TABLE 1.4. Partial List of WLAN Products [29]

Product Company Location	Freq. (MHz)	Link rate	User rate	Protocol(s)	Access	No. of chan. or spread factor	Mod./coding	Power	Network Topology
Altair Plus II Motorola Arlington Hts., IL	18-19 GHz	15 Mb/s	5.7 Mb/s	Ethernet			4-level FSK	25 mW peak	Eight devices/ radio; radio to base to Ethernet
WaveLAN NCR/AT&T Dayton, OH	902-928	2 Mb/s	1.6 Mb/s	Ethernet-like	DS SS		DQPSK	250 mW	Peer-to-peer
AirLAN Solitec San Diego, CA	902-928		2 Mb/s	Ethernet	DS SS		DQPSK	250 mW	PC/MCA w/ ant; radio to hub
Freeport Windata Inc. Northboro, MA	902-928	16 Mb/s	5.7 Mb/s	Ethernet	DS SS	32 chips/bt	16 PSK trellis coding	650 mW	Hub
Intersect Persoft Inc. Madison, WI	902-928		2 Mb/s	Ethernet, token-ring	DS SS		DQPSK	250 mW	Hub
LAWN O'Neill Comm Horsham, PA	902-928		38.4 kb/s	AX.25	SS	20 users/chan., max. 4 chan.		20 mW	Peer-to-peer
WiLAN Wi-LAN Inc. Calgary, Alberta	902-928	20 Mb/s	1.5 Mb/s/ chan.	Ethernet, token ring	CDMA/ TDMA	3 chan. 10-15 links each	"unconven- tional"	30 mW	Peer-to-peer
RadioPort ALPS Electric USA	902-928		242 kb/s	Ethernet	SS	1/3 channels		100 mW	Peer-to-peer
AirLAN 600 Telesys. SLW Don Mills, Ont.	902- 928, 2.4 GHz		1.35 Mb/ s	Ethernet	SS			1 W max	PC's with ant., radio to hub
RadioLink Cal. Microwave Sunnyvale, CA	902- 928; 2.4 GHz	250 kb/s	64 kb/s		FH SS	250 ms/hop 500 kHz space			Hub
Range LAN Proxim, Inc. Mountainview, CA	902-928		242 kb/s	Ethernet, token ring	DS SS	3 chan.		100 mW	
RangeLAN2 Proxim, Inc. Mountainview, CA	2.4 GHz	1.6 Mb/s	50 kb/s max.	Ethernet, token ring	FH SS	10 chan. @ 5 kb/s; 15 sub- ch. each		100 mW	Peer-to-peer bridge
Netwave Xircom Calabasas, CA	2.4 GHz	1 Mb/s/ adaptor		Ethernet, token ring	FH SS	82 1-MHz chan. or "hops"			Hub
FreeLink Cabletron Sys. Rochester, NH	2.4 and 5.8 GHz		5.7 Mb/s	Ethernet	DS SS	32 chips/bt	16 PSK trellis coding	100 mW	Hub

Chapter 2

Channel Coding for Wireless Communications

2.1 Introduction

This chapter will give a description of convolutional coding, hard- and soft-decision Viterbi decoding, and diversity combining techniques. It will conclude with an example of the coding utilized in present standards, specifically the GSM standard.

2.2 Convolutional Codes^[27]

A convolutional code (CC) is a sequence of encoded symbols which is generated by passing the information sequence through a binary shift register as shown in Figure 2.1. At each symbol instant, a k -bit information symbol is inserted into the input stages of the shift register. The register consists of Kk binary stages that constitute the present k -bit information symbol and the $(K-1)$ previous k -bit input symbols. The parameter K is known as the

FIGURE 2.1. Convolutional encoder for the $CC(n, k, K)$ code

constraint length and it determines the memory length of the shift register. The n linear algebraic function generators, $g_1, g_2, \dots, g_n$, defined by their connections to various register

stages yield an n -bit convolutional coded symbol via the set of modulo-2 adders. Each generator is described as a vector with a dimension of full register length of Kk bits. The i -th element of the vector specifies the existence or otherwise of a connection from the i -th bit position of the register to the modulo-2 adder. The value of the vector is defined by assigning a logical 1 where the connections are made, and all other positions in the register where there are no connections are assigned to be logical zeros. At each symbol instant, a new k -bit information symbol enters the register, the other symbols move to the next symbol location and one symbol leaves the register. A new encoded symbol is then generated as previously described. The process continues, encoding the sequence of information symbols into a new sequence of output symbols.

The convolutional code is basically described by three parameters n , k and K and is denoted by $CC(n, k, K)$. The coding rate, defined by

$$R = \frac{k}{n} \quad (\text{EQ 2.1})$$

represents the amount of information per encoded bit.

2.3 State and Trellis Diagrams

For the sake of simplicity, we demonstrate the encoding process by considering the half-rate ($R = k/n = 1/2$), convolutional code, $CC(2, 1, 3)$ with constraint length of three binary stages ($K = 3$). The generators in Figure 2.2 are described by the vectors,

$$g_1 = [1 \ 0 \ 1] \quad \text{and} \quad g_2 = [1 \ 1 \ 1] \quad (\text{EQ 2.2})$$

or are written in equivalent form of polynomials as

$$g_1 = 1 + z^2 \quad \text{and} \quad g_2 = 1 + z + z^2 \quad (\text{EQ 2.3})$$

Figure 2.2 shows the action of the encoding process from an information sequence to a chain of state transitions and finally to a sequence of encoded symbols. As the information sequence is shifted into the register by $k = 1$ bit at a time, the encoder state obtains its

value by observing the information sequence with a window size of $(K - 1)k = 2$ bits. A chain of state transitions of ..., 01, 10, 01, 10, 00, is obtained from an information sequence of ...010100. New encoded symbols are formed in response to these state transitions and the sequence of the encoded symbols become ..., 10, 00, 10, 11.

The encoder state is given by the last $(K - 1)k$ register stages, namely 2 in this example. The number of possible states is $2^{(K-1)k}$ ($= 4$ here). The state at each instant can be either 00, 10, 01 or 11, which we label as state a , b , c or d , respectively. Figure 2.3 tabulates the possible state transitions at any symbol instant. The branch emanating from the present state to the next state indicates the state transition. The broken line branch is the transition initiated by input bit of logical 1, whereas the solid branch is due to the input bit being a logical 0. The symbol attached to each branch is an $n = 2$ bit word and is the encoded symbol. The number of outgoing branches emanating from the present state is 2^k ($= 2$ here), which corresponds to the number of the possible input patterns. Figure 2.3 represents the fundamental structure that can be used to construct the state diagram and trellis diagram of the convolutional code.

The state diagram corresponding to the state transition diagram is shown in Figure 2.4. It consists of a total number of $2^{(K-1)k} = 4$ states connected by all the possible transitions shown in the state transition diagram. In the example being considered, the shift register is initially cleared setting the encoder to state a . The present input is a logical 1, the state changes from a to b , as illustrated by the dotted branch emanating from state a in Figure 2.4. The encoder output is the symbol 11 that is assigned to the transition branch. At the

FIGURE 2.2. The action of the encoding process for CC(2, 1, 3)

FIGURE 2.3. State transition diagram

next instant, the present state becomes state *b* and the encoder receives a new input of logical 0. This causes the state transition from *b* to *c*, and the encoded symbol becomes 10. The encoding cycle is repeated for subsequent inputs, which change the states and then generate the encoded symbols. By following the change of states throughout the encoding process, a particular path associated with states $a \rightarrow b \rightarrow c \rightarrow b \rightarrow c \rightarrow \dots$, can be observed. This path is unique to the input sequence.

Another representation of the encoding process is the trellis diagram shown in Figure 2.5. This is formed by concatenating the consecutive instants of the state transition diagram of Figure 2.3 starting from the all-zero state. The diagram illustrates all the possible paths, namely 16, that can occur for 4-bit information sequences. We observe in Figure 2.5 that after the initial transient i.e., after instant $J = 1$, the trellis contains four nodes at every symbol instant. The trellis is extended, for example to the $J = 5$ symbol instant, by appending the state transition diagram of Figure 2.3 every node, i.e., state. In our example, the path corresponding to the input sequence is drawn by the thick line in Figure 2.5 corresponding to the encoded sequence of 11, 10, 00, 10, ...

FIGURE 2.4. The state diagram

FIGURE 2.5. The trellis diagram

2.4 Hard-decision Viterbi Decoding^[27]

The Viterbi algorithm is best explained using the trellis diagram of the simple binary CC(2, 1, 3) code. This algorithm is actually a simple version of forward dynamic programming^{[40], [41]}, in which the shortest route is determined. We will assume that the demodulator provides only hard-decisions when regenerating the information sequence. In this case the Hamming distances between the received symbols and the estimated transmitted symbols in the trellis are used as a metric, i.e. a confidence measure. The Hamming distance $d(c_m, c_n)$ between two sequences c_m and c_n of length n is the number of places in which they differ. Figure 2.6 records the history of the paths selected by the Viterbi decoder. Suppose there are no channel errors, the input sequence to the decoder is the same as the encoded sequence 11, 10, 00, 10, ... as illustrated in Figure 2.2. At the first instant $J = 1$, the received symbol is 11 which is compared with the possible transmitted symbols 00 and 11 of the branches from node a to a and from node a to b , respectively. The metrics of these two branches are their Hamming distances, namely the differences between the possible transmitted symbols (00) or 11 and between the received symbol 11. Their distances are 2 and 0, respectively.

Now we define that the branch metric is the Hamming distance of an individual branch, and the path metric at the J -th instant is the sum of the branch metrics at all its branches from $J = 0$ to the J -th instant. Hence the path metrics, printed on top of each node in Figure 2.6, at instant $J = 1$ are 2 and 0 for the paths $a \rightarrow a$ and $a \rightarrow b$, respectively. At the second instant $J = 2$, the received symbol is 10 and the branch metrics are 1, 1, 0 and 2 for the branches $a \rightarrow a$, $a \rightarrow b$, $b \rightarrow c$, and $b \rightarrow d$, respectively. The path metrics, or accumulated metrics, are 3, 3, 0 and 2 for the corresponding paths $a \rightarrow a \rightarrow a$, $a \rightarrow a \rightarrow b$, $a \rightarrow b \rightarrow c$ and $a \rightarrow b \rightarrow d$. At the third instant, the received symbol is 00. There are eight possible branches as seen also in Figure 2.5 and their branch metrics are 0, 2, 2, 0, 1, 1, 1 and 1 for the branches $a \rightarrow a$, $c \rightarrow a$, $a \rightarrow b$, $c \rightarrow b$, $b \rightarrow c$, $d \rightarrow c$, $b \rightarrow d$ and $d \rightarrow d$, respectively. Let α_1 and α_2 denote the corresponding paths $a \rightarrow a \rightarrow a \rightarrow a$ and $a \rightarrow b \rightarrow c \rightarrow a$ that begin at the initial node a and remerge at node a at $J = 3$. Their respective path metrics are 3 and 2. Any further branches with $J > 3$ stemming from the node a at $J = 3$ will add identical branch metrics to the path metric of both paths α_1 and α_2 , and this means that the path metric of α_1 is larger at $J = 3$ and will continue to be larger at $J > 3$. The Viterbi decoder selects the path with the smallest metric and therefore decides to discard the α_1 path and retain the α_2 path. The α_2 path is called the *survivor*. This procedure is also applied at the other nodes b , c and d at $J = 3$. Notice that paths $a \rightarrow a \rightarrow a$ and $a \rightarrow a \rightarrow b$ cannot survive as their path metrics are larger than that of their counterparts of the merging pairs and they are therefore eliminated from the decoder memory. Thus there are four paths that survive at each instant and there are four path metrics, one for each of the surviving paths. Similarly, at the instant $J = 4$, there are again four survivors and the survivor with the smallest metric is the path $a \rightarrow b \rightarrow c \rightarrow b \rightarrow c$. This path corresponds to the transmitted sequence and hence the correct information sequence 1010... is delivered to the decoder output. This decoding is optimum in the sense that it minimizes the probability that the entire sequence is in error.

FIGURE 2.6. Example of Viterbi Decoding

The formal algorithm follows:

Viterbi Algorithm ^[21]

Storage:

k (time index)

$\hat{x}(x_k)$, $1 \leq x_k \leq M$ (survivor terminating in x_k);

$\Gamma(x_k)$, $1 \leq x_k \leq M$ (survivor length).

Initialization:

$k = 0$;

$\hat{x}(x_0) = x_0$; $\hat{x}(m)$ arbitrary, $m \neq x_0$;

$\Gamma(x_0) = 0$; $\Gamma(m) = \infty$, $m \neq x_0$

Recursion: Compute

$$\Gamma(x_{k+1}, x_k) = \Gamma(x_{k+1}, x_k) + \lambda |\xi_k = (x_{k+1}, x_k)|$$

for all $\xi_k = (x_{k+1}, x_k)$.

Find

$$\Gamma(x_{k+1}) = \min \Gamma(x_{k+1}, x_k)$$

for each x_{k+1} : store $\Gamma(x_{k+1})$ and the corresponding survivor $\hat{x}(x_{k+1})$.

Set k to $k+1$ and repeat until $k=K$.

With finite state sequences x the algorithm terminates at time K with the shortest complete path stored as the survivor $\hat{x}(x_K)$.

Certain trivial modifications are necessary in practice. When state sequences are very long or infinite, it is necessary to truncate the survivors to some manageable length δ . In other words, the algorithm must come to a definite decision on nodes up to time $k-\delta$ at time k . In general, if the truncation depth δ is chosen large enough, there is a high probability that all time- k survivors will go through the same nodes up to time $k-\delta$, so that the initial segment of the maximum-likelihood path is known up to time $k-\delta$ and can be put out as the algorithm's firm decision. It has been found that a delay $\delta > 5(v-1)$, where v is the length of the shift register, results in a negligible degradation in the performance relative to the optimum Viterbi algorithm.^[3] A simple method of truncating the sequence is to append a tail of logical zeros ($= (v-1)$) for the purpose of clearing the shift-register of all information bits. By this means, the encoder returns to the all-zero state. As the stuffing of logical zeros carries no information, the actual coding rate is now below R . In order to maintain the coding rate close to R , the encoded sequence (period of truncation) needs to be very long.

The complexity of the algorithm can be easily approximated. In terms of memory, the algorithm requires M storage locations, one for each state, where each location must be capable of storing a "length" $\Gamma(m)$ and a truncated survivor listing $\hat{x}(m)$ of δ symbols. In terms of computation, the algorithm must make $|\Xi|$ additions in each unit of time, one for each transition, and M comparisons among the $|\Xi|$ results. Thus, the amount of storage is proportional to the number of states, and the amount of computation to the number of transitions. With a shift-register process, $M = m^v$ and $|\Xi| = m^{v+1}$, so that the complexity increases exponentially with the length v of the shift register.

2.5 Punctured Convolutional Codes

For a convolutional code of rate k/n there are 2^k merging paths at each node in the trellis. The decoding of this code by the Viterbi algorithm selects the path with the highest metric out of the 2^k possibilities at each node. The number of calculations per selection at each node grows exponentially with k , rendering the implementation of the codec for operation at high speed a difficult task, particularly in the case of high-rate codes.

The Viterbi decoding of high rate codes, where $k > 1$ can be significantly simplified by employing punctured convolutional codes^[51]. Puncturing allows us to obtain a high-rate code by periodically deleting some of the coded parity bits from a low-rate encoder output. In addition, puncturing of the low-rate $1/n$ code results in the decoding trellis operating with k equal to unity.^[27] In this fashion we replace 2^k -ary comparisons at each state by binary comparisons. In a high-speed decoder this simplification has a major impact on decoder complexity.

An example is provided in Figure 2.7. A coding rate of $R = 2/3$ can be achieved by periodically deleting bits from the CC(2, 1, 3) half rate code. A puncturing matrix is assigned at the output of a half rate encoder, where an element of 1 in the matrix allows its input bit to appear at the output, whereas an element of 0 in the matrix deletes the incoming bit at its input. The top row of the matrix consists of 11 which does not delete any bits from the output of adder 1. The second row of the matrix is 10 which deletes every alternate bit from the output of adder 2. The encoded sequence is formulated by sampling the adder's output alternately. In this case, every fourth encoded bit is deleted, the encoder will produce three output bits for every two input bits resulting in a $R = 2/3$ code.

An example of punctured coding is shown in Figure 2.8 by using the encoder described in Figure 2.7. Suppose the input sequence to the encoder is ..., 0, 1, 0, 1. The output sequence is punctured every fourth bit and becomes ..., 0, 00, 0, 11. We then apply hard-decision Viterbi decoding as shown in Figure 2.6. At the first instant, $J = 1$, the received symbol is 11 which is compared with the possible transmitted symbols 00 and 11 of the branches $a \rightarrow a$ and $a \rightarrow b$. Their Hamming distances are 2 and 0, respectively. At the second instant, $J = 2$, the received symbol consists of only a single bit 0, as the other bit of the symbol was punctured at the encoder. The corresponding bit being punctured in the trellis is marked by X in Figure 2.8. The received bit is then compared with the possible transmitted bits 0, 1, 0 or 1 for the corresponding branches $a \rightarrow a$, $a \rightarrow b$, $b \rightarrow c$, and $b \rightarrow d$. The path metrics are 2, 3, 0 and 1 for the branches $a \rightarrow a \rightarrow a$, $a \rightarrow a \rightarrow b$, $a \rightarrow b \rightarrow c$ and $a \rightarrow b \rightarrow d$, respectively. Similarly, Viterbi decoding is applied to the third and fourth instants. At the instant $J = 4$, there are four survivors and the survivor with the smallest metric is the path $a \rightarrow b \rightarrow c \rightarrow b \rightarrow c$. This path corresponds to the transmitted sequence and hence the correct information sequence 1010 ... is delivered to the decoder output.

Rate compatible punctured codes (RCPC) are obtained by ensuring that all code bits of the higher rate punctured codes are used by the lower rate codes. For example (see Reference ^[26]), assume we have a rate-1/2 convolutional code with memory $M=2$ that is punctured periodically with period $P=4$. The puncturing table can be described by an $N \times P$ matrix (where $N=2$ for the rate-1/2 code):

$$a(1) = \begin{bmatrix} 1 & 1 & 1 & 0 \\ 1 & 0 & 0 & 1 \end{bmatrix}$$

A zero in the puncturing table means that the code symbol is not to be transmitted. In the table above the fourth bit from the first polynomial and the second and third bit from the second polynomial are not to be transmitted. Instead of transmitting $2 \cdot P = 2 \cdot 4$ only $P + 1 = P + 1 = 5$ bits are transmitted per $P = 4$ information bits. Therefore, we have generated a code with rate $R = 4/5$ by using only $P + 1$ "1"s instead of $2P$ "1"s in the puncturing table. If the rate-4/5 code is not powerful enough a more redundant code with lower rates

4/6, 4/7, 4/8 could be used. Instead of transmitting all the code bits of a completely different low rate code, the lower rate code should utilize the bits already transmitted. Then only additional incremental redundancy bits have to be transmitted. Additional "1" 's in the puncturing tables of the lower rate codes can therefore be placed only where zeros appeared in the puncturing matrix of the previous higher rate code, for example,

$$a(2) = \begin{bmatrix} 1 & 1 & 1 & 0 \\ 1 & 1 & 0 & 1 \end{bmatrix}, a(3) = \begin{bmatrix} 1 & 1 & 1 & 1 \\ 1 & 1 & 0 & 1 \end{bmatrix}, a(4) = \begin{bmatrix} 1 & 1 & 1 & 1 \\ 1 & 1 & 1 & 1 \end{bmatrix}$$

Thus for $l = 1$ to 4, we get a compatible family of codes derived from the mother code 1/2 with rates 4/5, 4/6, 4/7, 4/8.

References ^{[11],[15]} examined a family of maximal-free distance rate-compatible punctured codes (RCPC) for the (2, 1, 7) code used in later chapters.

2.6 Soft Output Viterbi Algorithm ("SOVA")

Forney^[21] considered "augmented outputs" from the Viterbi algorithm itself, such as the depth at which all paths are merged, the difference in length between the best two paths, and a list of the p best paths. References ^[19] and ^[20] describe implementations of the SOVA scheme in much the same vein. This is a Viterbi algorithm which uses hard (or soft) decisions to calculate its metrics, but also decides in a soft way by providing reliability information together with the output bits.

The fundamental information for allocating a reliability value to the choice of one path (the so-called survivor), of the two paths which merge at a state s_k , is the difference between the two accumulated metrics associated with this state. Let us call $\Gamma_s(k, s_k)$, the accumulated metric of the survivor at time k and state s_k , and $\Gamma_c(k, s_k)$, the accumulated metric of the other path coming at the same time to state s_k , which is called the concurrent path. The greater the difference $|\Gamma_s(k, s_k) - \Gamma_c(k, s_k)| \geq 0$, the more reliable the selection of the survivor at time k and state s_k . Because the two paths give two opposite binary decisions about the encoder input, $u(k)$, $\Gamma_s(k, s_k) - \Gamma_c(k, s_k)$ can be directly assigned as a first estimation of the weight of the decision given by the survivor about $u(k)$. To clarify this, see the example in Figure 2.9, of an S-state trellis. k is the current time at the input of the trellis. In

this example, the survivor and concurrent paths considered at time k , have diverged at time $k - \delta_m$. The two paths give opposite decisions at time $k - 1$ and $k - v$.

Assume that the number of states is S , where $S = 2^v$ and v is the code memory. Also assume that the classical VA makes a final decision with delay δ , δ being large enough so that all 2^v survivor paths have been merged with sufficiently high probability. The SOVA can be formulated as below. The steps marked by (*) augment the classical Viterbi algorithm.

Storage:

k (time index)

$$\hat{u}(s_k) = \{\hat{u}_{k-\delta}(s_k), \dots, \hat{u}_{k-\delta}(s_k)\}, \quad 0 \leq s_k \leq S-1 \text{ (hard decision values, } \hat{u} \in \{\pm 1\} \text{)}$$

$$\hat{L}(s_k) = \{\hat{L}_{k-\delta}(s_k), \dots, \hat{L}_{k-\delta}(s_k)\}, \quad 0 \leq s_k \leq S-1 \text{ (soft reliability values, } 0 \leq \hat{L} \leq \infty \text{)} (*)$$

$$\Gamma(s_k), \quad 0 \leq s_k \leq S-1, \text{ (accumulated metric value)}$$

Recursion:

a) Classical Viterbi step:

For each state s_k

Compute

$$\Gamma(s_{k-1}, s_k) = \Gamma(s_{k-1}, s_k) + \lambda |\xi_k(s_{k-1}, s_k)|$$

for all $\xi_k(s_{k-1}, s_k)$.

$$\text{Find } \Gamma(s_k) = \min \Gamma(s_{k-1}, s_k).$$

Store $\Gamma(s_k)$ and the corresponding survivor $\hat{u}_k(s_k)$.

b) Soft-deciding update: (*)

For each state s_k

$$\text{Store } \Delta = \max \Gamma(s_{k-1}, s_k) - \min \Gamma(s_{k-1}, s_k).$$

$$\text{Initialize } \hat{L}_k(s_k) = \infty.$$

$$\text{For } j = k - v \text{ to } j = k - \delta_m.$$

Compare the two paths merging in s_k

if $\hat{u}_j^{(1)}(s_j) \neq \hat{u}_j^{(2)}(s_j)$ then update

$$\hat{L}_j \leftarrow \min(\hat{L}_j, \Delta)$$

FIGURE 2.9. Example of the SOVA^[19,20]

2.7 Diversity Schemes

In mobile radio systems, the effects of signal fading can be combatted by the use of diversity techniques either at the base station or the mobile. Diversity reception works on the principle that if two or more independent samples of a random process are taken, these

samples will fade in an uncorrelated manner. The probability, therefore, of all the samples being below a given level simultaneously is very much less than the probability of a single sample being below that level.

There are two kinds of fading: long-term fading caused by shadowing which is characterized by a log-normal distribution and short-term fading caused by multipath which is characterized by a Rayleigh distribution.

To combat long-term fading, one uses “macroscopic diversity” which applies to separated antenna sites. To combat short-term fading, one uses microscopic diversity which applies to co-located antenna sites.

2.7.1 Microscopic Diversity

Microscopic diversity techniques can be adopted to mitigate the effect of Rayleigh fading; the term micro is used because decorrelation occurs within about $1/2$ wavelength (typically less than one foot for frequencies above 500 MHz). Microscopic diversity schemes require two or more co-located antennas. There are various ways in which to combine the output of these diversity branches explained below.

2.7.2 Macroscopic Diversity

Macroscopic diversity techniques are used to combat shadow fading; the term macro is used because much larger spatial separation is required to achieve decorrelation. The basic concept of macroscopic diversity is the use of a group of geographically distributed ports (base stations) to serve the user by selection of the best signal among those ports according to some selection algorithm.

2.7.3 Diversity Combining Techniques

Some of the most commonly applied combining techniques are illustrated in the following figure:

FIGURE 2.10. Diversity Combining Techniques

(A) Selective combining

(B) Switched combining

(C) Maximal-ratio combining

(D) Equal-gain combining

- selective combining: select strongest signal among M diversity branches

- **switched combining:** two diversity signals are selected, based on a given threshold level in one receiver. If signal A above a threshold L is selected to receive, it receives until it falls below the level L . Then the receiver switches to signal B regardless of whether signal B is above L or below. Signal B should be above L , but if it is below, then depending on the switching algorithm of the receiver, it can switch back to signal A or stay in signal B until signal B rises above level L . Switched combining performance is not always as good as that of selective combining. Since switched combining needs only one receiver, it is less costly and may be used on mobile units. However, the performance is greatly affected by threshold level and switching noise. To improve this scheme, the threshold level L needs to be changed dynamically in real time based on the present signal level received in the environment. [13]
- **maximal-ratio combining:** demodulated signals from each diversity branch are weighted for maximal ratio combining in proportion to their signal to noise power ratio, and then summed.
- **equal-gain combining:** the combined signal is the sum of the instantaneous fading envelopes of the individual branches. Equal-gain combining has only one dB degradation as compared to maximal-ratio combining.

2.7.4 Code Combining

Code combining is a technique for combining any number of repeated packets encoded with a rate- R code to obtain a lower rate, and thus more powerful error-correcting code, capable of allowing communications when channel error rates are less than 50 percent^[23]. Alternately if diversity antennas are used, the code combining would be done on packets received in parallel. Two critical features of code combining are as follows:

1) maximum-likelihood decoding, rather than bounded minimum distance (algebraic) decoding, of L noisy packets as a single low-rate code of rate R/L ; with bounded distance decoding, if the received word lies within distance t of a valid codeword, where t is the error correction capability of the code, the decoder chooses this unique codeword as the one which was transmitted. With maximum likelihood decoding (MLD), the decoder chooses the codeword closest to the received word. This distance can be

greater than t , the error correcting capability of the code. As shown in [23], a bounded distance decoder is not effective if the channel error probability is $p > 0.25$. However, with MLD and code combining, effective communications is possible as long as $p < 0.5$.

2) weighting of each packet by an estimate of its reliability (soft decision on packets).

Figure 2.11 illustrates the sequence of mappings used in code combining. We start with an information packet I composed of K bits. This packet may contain an error detection code so that the decoder can determine when to stop combining packets. The information packet is mapped into an encoded packet x , of N bits, by an error-correction code. Next, the encoded packet is transmitted over the noisy channel L times, producing the received sequences $\gamma^1, \gamma^2, \dots, \gamma^L$. Along with each encoded packet a number of extra bits may be transmitted to provide an estimate of the channel reliability. The decoder is referred to as a code combiner since it combines successive receptions until the code rate is low enough to provide complete error correction.

A maximum-likelihood code combiner treats the received packets $Y^1, Y^2, \dots, Y^L$ as a code of rate R/L and attempts to pick the most probable information packet. Each packet is assumed to have a reliability weight of w_i . The decoder may be a rate R/L soft or hard decoder, but will also have the capability of weighting the reliability of each received packet.

As an example, Figure 2.12 shows the trellis diagram for the rate $1/2$ memory 2, $(2, 1, 2)$ convolutional code. Figure 2.13 shows the trellis diagram for the same code repeated twice. As can be seen, the only difference between these two trellis diagrams is that the symbols on each branch are repeated twice in Figure 2.13. Therefore, the distance of each path in the trellis of Figure 2.13 (and also the free distance) is twice that of Figure 2.12. This can be extended to the case where each bit is repeated L times so that, if a rate R con-

volitional code with a minimum free distance d_{free} is repeated L times, the result is a code with minimum free distance Ld_{free} [52].

Figure 2.12: The Trellis for the (2, 1, 2) Convolutional Code

Figure 2.13: The Trellis for the Repeated CC(2, 1, 2)

An additional combining technique of interest was examined in [17,32].

In the Viterbi algorithm, the branch metrics associated with the state transitions are computed and then added to the previous path metrics. The contending path metrics are compared and the path metric with the largest metric is selected as the survivor. In this diversity method, the path metrics are computed for each diversity branch, and then simultaneously compared to select the survivor as shown. Therefore, this method has only one set of path-history storage. (ACS = Add-Compare-Select)

2.8 Interleaving and Coding applied in GSM (Global System for Mobile Communications)

It is illustrative to examine the coding and interleaving used in GSM; this section will give a brief example for full rate speech. Although this is only a low-rate real time application, it can provide an indication of the acceptable delay range for delay-sensitive services.

The terminals and base stations insert information into the channel in time *slots* (also referred to as a "burst") of duration 577 μ s. Eight consecutive time slots comprise a TDMA frame of duration 4.62 ms, while 26 frames comprise a *multiframe* with duration 120 ms.

Every 20 ms, the speech coder and its associated channel coder release 456 bits. An interleaver collects the speech information generated in a 40 ms interval (912 bits) and permutes the order of the serial bit stream as indicated in Figure 2.15. The purpose of the interleaving is to reduce the burstiness of transmission errors in the decoded bit stream. When the bit stream is interleaved prior to transmission and afterward returned to its original order, errors that are clustered in the sequence of 912 channel bits tend to be randomly dispersed in the bit stream presented to the speech coder.

The input rate for speech transmission mode on a full rate traffic channel (TCH/FS in Table 2.2) is equal 13 kbit/s, by blocks of 260 bits (i.e. one block every 20 ms). When errors occur, the speech is disturbed in different ways depending on the role of the bits in error. The bits receive unequal protection across this block as seen in the table. Blocks of full rate speech are interleaved on 8 bursts: the 456 bits of one block are split into 8 groups of 57 bits (a half-burst), each one carried by a different burst. A burst therefore contains the contribution of two successive speech blocks A and B. In order to destroy the proximity relations between successive bits, bits of block A use the even positions inside the burst and bits of block B the odd positions^[42]. This is illustrated in Table 2.1. The corresponding delay taking into account one additional burst for the SACCH is equal to $(8 \times 8) + 1 = 65$ burst periods = 37.5 ms. Similarly, in the digital cellular standard for North America, IS-54, the interleaving delay for speech transmission is restricted to 40 ms.

Note that the above example was for full rate speech and the interleaving employed for other packets varies as seen in Table 2.2.

TABLE 2. 1: Interleaving Algorithm for Full Rate Speech^[41]

0	8	.	.	.	.	.	.	.	.	448	even bits of burst N
1	9	.	.	.	.	.	.	.	.	449	even bits of burst N + 1
2	10	.	.	.	.	.	.	.	.	450	even bits of burst N + 2
3	11	.	.	.	.	.	.	.	.	451	even bits of burst N + 3
4	12	.	.	.	.	.	.	.	.	452	odd bits of burst N + 4
5	13	.	.	.	.	.	.	.	.	453	odd bits of burst N + 5
6	14	.	.	.	.	.	.	.	.	454	odd bits of burst N + 6
7	15	.	.	.	.	.	.	.	.	455	odd bits of burst N + 7
<----- 57 columns ----->											

TABLE 2.2 Interleaving and coding for the different transmission modes^[41]

Channel and transmission mode		Input rate (kbit/s)	Input block	Coding	Output Block	Interleaving
	Ia		50	Parity (3 bits) Convolutional 1/2		
TCH/FS	Ib	13	132	Convolutional 1/2	456	On 8 half-bursts
	II		78	none		
TCH/F9.6 TCH/H4.8		12 6	240	Convolutional 1/2 punctured 1 bit out of 15	456	Complex, on 22 unequal burst portions

TABLE 2.2 Interleaving and coding for the different transmission modes ^[41]

Channel and transmission mode	Input rate (kbit/s)	Input block	Coding	Output Block	Interleaving
TCH/F4.8	6	120	Addition of 32 null bits Convolutional 1/3	456	Complex, on 22 unequal burst portions
TCH/F2.4	3.6	72	Convolutional 1/6	456	On 8 half bursts
TCH/H2.4	3.6	144	Convolutional 1/3	456	Complex, on 22 unequal burst portions
SCH		25	Parity (10 bits) Convolutional 1/2	78	On 1 S burst
RACH (+ handover access)		8	Parity (6 bits) Convolutional 1/2	36	On 1 access burst
fast associated signalling TCH/F and /H		184	Fire code 224/184	456	On 8 half bursts
TCH/8, SACCH: BCCH, PAGCH					On 4 full bursts

GSM characterizes the transmission modes (TCH = Traffic Channel) used in Table 2.1 as follows:

- the TCH/F (where F stands for “full rate”) allows the transmission of 13 kbit/s speech or of data at 12, 6 or 3.6 kbit/s;
- the TCH/H (where H stands for “half rate”) allows speech coded at a rate around 7 kbit/s to be transported, or data at a rate of 6 or 3.6 kbit/s;
- SCH - synchronization channel; channel broadcast by the base station to enable the mobile station to stay synchronized with it.
- RACH - random access channel; allows the mobile stations to transmit their requests to the network in a random manner.
- BCCH - broadcast control channel; information is broadcast in each cell on this channel providing information required by the mobile.
- TCH/8 - separate signalling channel between the mobile and network.

- **fast associated signalling** - the TCH is used by messages to indicate the call establishment progress, or to authenticate the subscriber, or to command a handover.
- **SACCH** - slow associated control channel - used for non-urgent procedures, mainly the transmission of the radio measurement data needed for the decisions concerning handover.
- **PAGCH** - paging and access grant channel - paging messages are transmitted on this channel.

Chapter 3

Comparison of Diversity Schemes in Additive White Gaussian Noise

3.1 Introduction

This purpose of this chapter is to compare the performance of the packet diversity techniques described below in an additive white Gaussian noise (AWGN) channel.

- code combining

In this method, the noise corrupted encoded packet from each of L diversity branches is combined (if none of the L packets' CRC check are correct) and decoded by means of a rate r/L (where L is the number of diversity branches) convolutional code that has the same number of states as the original convolutional code. If at least one of the packets'

CRC checks, the corresponding packet is assumed correct, and combining is not necessary. It is assumed that there is a CRC protecting the coded packet so that the code-combining is "non-blind". Notice that in the case of single branch communication, one could buffer the noise-corrupted packets and combine them with subsequent retransmissions (hybrid ARQ-type-II).

- ideal post Viterbi decoding selection

In this method, each diversity branch signal received is individually demodulated and Viterbi-decoded. Decoded results are compared and a selection is made based on the

branch with the fewest decoded errors. This is an ideal selection scheme representing the best possible performance of a selection based on CRC-detection of decoder errors.

Figure 3.2: Post-Viterbi-decoding selection

The communication system used in this section is described in Figure 3.4. A variable-rate hard-decision Viterbi decoder is employed by using the concept of punctured convolutional coding (PCC). The mother code used was the rate 1/2 constraint-length 7 convolutional code, (2,1,7) with generator polynomials (133,171) (see Figure 3.3 below). Reference ^[11] examined a family of rate compatible punctured convolutional (RCPC) codes for the (2, 1, 7) code used here. Table 3.1 lists the optimum codes (d = Hamming free distance) found. The rate-3/4 and rate-7/8 code were used here in addition to the mother rate-1/2 code.

Convolutional codes are used in this work based on the observation that lower rate

Figure 3.3: Convolutional Encoder

codes resulting from the packet repetition are very close to the maximum possible free distance that could be attained by means of an equal rate convolutional code. ^[1,24]

This observation was also made by Chase ^[22]. The convolutional code used is moderate in complexity and retains its effectiveness as it is repeated.

The minimum free distance of a convolutional code is upper bounded ^[28] by

$$d' \leq \min_h \frac{2^h}{2^h + 1} \cdot \frac{K + h - 1}{2R} \quad (\text{EQ 3.1})$$

where R is the code rate, K is the constraint length, and h is a parameter chosen to optimize the bound. Table 3.2 compares the minimum distance of the repeated $K = 7$ convolutional code and this upper bound on the best known $K = 7$ convolutional codes. For $R = 1/2$ and $1/4$ the selected code is optimum in the sense that no convolutional code can be found with a larger minimum distance. For rates below $1/4$ the difference between d and d' is small enough that even if codes with distance d' can be found (note d' is only an upper bound on minimum distance) the actual performance difference would be negligible ^[22].

TABLE 3.1. Parameters of examined codes^[11]

Code	Coding Rate	Puncturing Table (0: deleting bit position)	d _{free}
1/2	0.5	1 (133) --- Code generator 1 (171) --- Code generator	10
2/3	0.667	11 19	6
3/4	0.75	110 101	5
4/5	0.8	1111 1000	4
5/6	0.833	11010 10101	4
6/7	0.857	111010 100101	3
7/8	0.875	1111010 1000101	3
8/9	0.889	11110100 10001011	3
9/10	0.9	111101110 100010001	3
10/11	0.909	1110110111 1001001000	3
11/12	0.917	11110111110 10001000001	3
12/13	0.923	111111110101 100000001010	3
13/14	0.929	1101000001111 1010111110000	3
14/15	0.933	11010110110110 10101001001001	3
15/16	0.938	111001011010010 100110100101101	3
16/17	0.941	1101010010000101 1010101101111010	3

TABLE 3.2. Minimum Free Distance for a Repeated Constraint Length 7 Convolutional Code Compared to Upper Bounds on Optimum Codes [22]

Diversity	R = Code Rate	d = Minimum Free Distance for the Selected Code	d' = Upper Bound on the Minimum Free Distance
1	1/2	10	10
2	1/4	20	20
4	1/8	40	41
8	1/16	80	82
16	1/32	160	164
32	1/64	320	329
64	1/128	640	658

The modulation used in the system is assumed to be BPSK (binary phase shift keying) which has a corresponding probability of bit error [3]:

$$P_b = Q\left(\sqrt{2\gamma_b}\right) \quad (\text{EQ 3.2})$$

where γ_b is the signal-to-noise ratio per information bit. Also, the probability of symbol error is:

$$P_{cs} = Q\left(\sqrt{2\gamma_{cs}}\right), \quad \gamma_{cs} = R\gamma_b \quad (\text{EQ 3.3})$$

where γ_{cs} is the signal-to-noise ratio per coded symbol and R is the code rate.

The function $Q()$ is the so-called error function [3]

$$Q(x) = \frac{1}{\sqrt{2\pi}} \int_x^{\infty} e^{-t^2/2} dt$$

3.2 Simulation Description

The diversity techniques were compared based on the cell loss probability (at least one uncorrectable error in an ATM cell) and the probability of bit error. These probabilities were obtained by a simulation program written in C. The packet format used in the simulation is that of an ATM cell (53 bytes) augmented by a Cyclic Redundancy Check (CRC)

that protects the ATM cell payload. Subsequent application of a rate-1/2 convolutional code (CC) doubles the packet size and the required transmission rate. Figure 3.5 below describes the packet format before it is augmented by a CRC, after it is augmented by a CRC, and finally after it is convolutionally encoded. It is assumed that there is an additional CRC protecting the coded packet. This will determine whether code-combining should be done using the multiple copies of the packet received in parallel from the n diversity branches. In this manner the code-combining will be non-blind and the packet

will not have to be decoded twice. When hybrid-ARQ is examined in Chapter 5 it is assumed that there is a CRC protecting the information bits only.

The simulation code was split into the following main modules:

- codegen.c: generates random ATM packets and convolutionally encodes the packets
- noisegen.c: generates an error sequence for the ATM packet based on the channel assumption (AWGN in this chapter) and the modulation (BPSK).

- decode.c:** hard-decision Viterbi decoding and code-combining algorithms. A static array is also defined here for the first $v-1$ stages (where v is the constraint-length of the code) of the trellis after which it repeats.

The input parameters for the simulation are N_{div} , the number of diversity branches, the punctured matrix in file format (from Table 3.2), the SNR, and the required number of packets to generate in the simulation..

The following steps are used in the simulation. In this simulation we consider two and three independent diversity branches.

1. Generate multiple copies (for two or three diversity branches) of an ATM packet consisting of random bits.
2. Encode the copies of the packet using the (2, 1, 7) convolutional code and the appropriate puncturing matrix from Table 3.1.
3. Generate random errors amongst the independent copies using equation (3.3). Add M "0" bits to the packets to properly terminate the encoder memory and the decoder trellis. M is the memory of the convolutional encoder.
4. Decode the copies of the packet independently using hard-decision Viterbi decoding; count the decoded bit errors and lost cells and store these statistics for the no diversity case. (Recall that hard decision decoding is used since the transmission rates considered here are OC-3 (= 155 Mb/s) and soft-decision decoding would not be feasible.)
5. Select the decoded packet with the fewest errors and store the number of decoded bit errors and lost cells for the post-Viterbi decoding selection case.
6. The CRC of the coded packets is checked (Note this CRC is on the coded packet). If all copies are in error then the coded packets are combined and decoded using the Viterbi algorithm and the code combining approach described earlier. The statistics are stored for the equal weights code combining case.
7. Repeat steps 2. through 6. until the required number of packets has been generated.

The number of packets generated at the input was of the order 10^5 for most cases. The number of packets generated at the lower error rates was two orders of magnitude more

than the loss probability. For example at a cell loss probability of 10^{-5} , 10^7 packets would be generated to obtain a reasonable confidence level.

3.3 Results

The results of the simulation, namely, cell loss probability (at least one uncorrectable error in an ATM cell) and decoded bit error probability are shown in Figures 3.6 to 3.11 as a function of the energy per information bit to noise power density (E_b/N_o).

We can compare the results for no diversity with the Chernoff upper bound [3, p.465],

$$P_b \leq \frac{d}{dN} T(D, N) \Big|_{N=1, D=\sqrt{4p(1-p)}} \quad (\text{EQ 3.4})$$

where p is the probability of symbol error given by equation (3.3) and $T(D, N)$ is the transfer function of the rate-1/2 constraint-length 7 convolutional code.

The convolutional code used here has a bit error bound of [3, 12],

$$P_b \leq \frac{1}{2} [36D^{10} + 211D^{12} + 1404D^{14} + 11633D^{16} + \dots] \quad (\text{EQ 3.5})$$

This curve is displayed in Figure 3.6 for the mother rate-1/2 code.

The results shown in Figure 3.6 apply to an AWGN channel with two independent diversity branches, and a (2, 1, 7) convolutional code. Similarly, the results shown in Figure 3.7 and Figure 3.8 apply to an AWGN channel with two-branch diversity, and a punctured convolutional code (PCC) of rate 3/4 and 7/8 which is generated by periodically deleting code symbols generated by the mother code (2, 1, 7). Figures 3.9 to 3.11 apply to three independent diversity branches for rate-1/2, rate-3/4, and rate-7/8 convolutional codes as well.

3.4 Discussion

The results clearly show the superiority of the code-combining approach over the post-Viterbi decoding selection scheme. The code-combining approach is about 2 dB superior to the post-Viterbi decoding selection scheme at a decoded bit error probability of 10^{-4} for

two diversity branches (Figure 3.6) and it is about 3 dB superior for three diversity branches (Figure 3.9) at the same error probability. It is interesting to note that the code-combining curve in Figure 3.8 (two-branch diversity, PCC-7/8) is comparable to the single coded branch in Figure 3.6 (CC(2, 1, 7)); the code-combining curve represents a repeated code of rate $R/2=7/16$. In this manner the coding overhead is distributed amongst the diversity branches.

In the next chapter we will examine the same packet diversity techniques in a variety of fading conditions.

Figure 3.6: Two-Branch Diversity of BPSK in AWGN using CC(2, 1, 7)

Figure 3.7: Two-Branch Diversity of BPSK in AWGN using PCC-3/4

Figure 3.8: Two-Branch Diversity of BPSK in AWGN using PC(7/8)

Figure 3.9: Three-Branch Diversity of BPSK in AWGN using CC(2, 1, 7)

Figure 3.10: Three-Branch Diversity of BPSK in AWGN using PC(1/4)

Figure 3.11: Three-Branch Diversity of BPSK in AWGN using PC(1/8)

Chapter 4

Comparison of Diversity Schemes in Rayleigh Fading

4.1 Performance in a Slow Rayleigh Fading Environment

4.1.1 Introduction

One of the impairments that a mobile communication channel encounters is amplitude fading which is characterized by a Rayleigh distribution. For digital transmission over a fading channel, the time variation causes a changing error probability with the effect of causing bursts of errors at the receiver output. Microscopic diversity techniques can be adopted to mitigate the effect of Rayleigh fading. This chapter analyzes the performance of various diversity schemes in a slow nonselective Rayleigh fading channel. The term slow is applied since we assume that the “instantaneous” signal-to-noise ratio γ *remains the same over a coded block of N bits* (where N = a coded ATM cell in our case; see Figure 3.2). This should be a reasonable assumption for indoor wireless applications where the speeds associated with the portable units are slow and fades would typically endure over several packets. Also, if a system employing sectorized antennas where the sector may be switched between packets such as ALTAIR¹ (described in [25]) is used it would be reasonable to assume that the packets are independent.

Such a system is analyzed in [30]. There, the traffic in each microcell is organized into frames of fixed duration, each consisting of both uplink and downlink traffic. The first portion of every frame is reserved for beacon envelopes and registration slots from all the base station's antenna sectors. The remainder of the frame is partitioned into sector segments, one for each of the base's antenna sectors. During each sector segment, the base transmits and receives only on that antenna sector. Each sector segment allows communication with a different set of portables. Each portable also has a sectorized antenna, with only one of its sectors used at a time. By monitoring the beacon envelopes, each portable determines the best combination of its own and base sector to use with its base station. [30]

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1. ALTAIR is a trademark of Motorola

Scheduling of each frame is done by the base station. Random scheduling ensures that any cochannel interference is essentially independent from frame to frame.

If this assumption were not valid the Rayleigh samples for each packet would need to be correlated. One method discussed to achieve this would be to essentially pass the samples through a low-pass filter as indicated below. The frequency of variance between the samples could then be altered by changing the filter coefficients. The autocorrelation

function is described in [54, p. 195]

For very slow nonselective Rayleigh fading with no diversity, the instantaneous signal-to-noise ratio γ has the probability density function $f(\gamma)$.

$$f(\gamma) = \frac{1}{\Gamma} e^{-\gamma/\Gamma} \quad (\text{EQ 4.1})$$

where Γ is the average signal-to-noise ratio per information bit. If we average the bit error probability, $P_e(\gamma)$ (given by (3.1) for BPSK), over the probability distribution of the instantaneous SNR, γ , we obtain^[3],

$$\begin{aligned} P_b &= \int_0^{\infty} P_e(\gamma) f(\gamma) d\gamma \\ &= \frac{1}{2} \left[1 - \sqrt{\frac{\Gamma}{1+\Gamma}} \right] \end{aligned} \quad (\text{EQ 4.2})$$

The probability of bit error under fading conditions, P_b , and the probability of more than M errors in a block of N bits under fading conditions, are not related in a simple manner. If signal strength is constant during the transmission of any one block of N bits (as is assumed in this section), but can fade from block to block, the probability of more than M bit errors in a block under conditions of Rayleigh fading is given by the average of $P(M, N)$ over the Rayleigh distribution [7.8]:

$$P(M, N) = \sum_{m=M+1}^N \binom{N}{m} \cdot \int_0^{\infty} P_e^m(\gamma) (1 - P_e(\gamma))^{N-m} f(\gamma) d\gamma \quad (\text{EQ 4.3})$$

The cell loss probability is then $P(M, N) |_{M=0, N=424}$

The same slow Rayleigh fading channel is assumed for each diversity branch with independent fading among branches.

The communication system model used in this section is described in the following figure.

4.1.2 Simulation Description

The simulation follows the following steps for two and three independent diversity branches:

1. Read the puncturing matrix file. Read the number of diversity branches = N_{div} . Read the average SNR per information bit, Γ , per branch.
2. Generate N_{div} independent sequences of Rayleigh distributed variables to represent the instantaneous SNR. The instantaneous SNR is obtained from the PDF given in equation (4.1) for the given average SNR, Γ ; where Γ has been scaled by the code rate to keep the energy per information bit constant. As stated earlier this instantaneous SNR is constant over each packet. The corresponding probability of bit error is then given by equation (3.2) for BPSK.
3. Generate multiple copies (= N_{div}) of an ATM packet consisting of random bits.
4. Encode the copies of the packet using the (2, 1, 7) convolutional code and the appropriate puncturing matrix from Table 3.2.
5. Generate random errors amongst the independent copies using the error sequences generated in step (2). Add M "0" bits to the packets to properly terminate the encoder memory and the decoder trellis. M is the memory of the convolutional encoder.
6. Decode the copies of the packet independently using the Viterbi algorithm; count the decoded bit errors and lost cells and store these statistics for the no diversity case.
7. Select the decoded packet with the fewest errors and store the statistics for the post-Viterbi decoding selection case.
8. The CRC of the coded packets is checked. (Note this CRC is on the coded packet) If all copies are in error then they are combined and decoded using the Viterbi algorithm and the code combining approach described earlier. The statistics are stored for the equal weights code combining case
9. Repeat steps 2. through 8.

The number of packets generated at the input was of the order 10^5 for most cases. The number of packets generated at the lower error rates was two orders of magnitude more than the loss probability. For example at a cell loss probability of 10^{-5} , 10^7 packets would be generated to obtain a reasonable confidence level.

4.1.3 Results

The results shown in Figures 4.3 to 4.5 apply to two independent diversity branches and Figures 4.6 to 4.8 apply to three independent diversity branches in slow Rayleigh fading. These results were obtained by the simulation described above, while the uncoded curve in the figures were given by (4.2) and (4.3) for the probability of bit error and cell loss probability, respectively.

Figure 4.3: Two-Branch Diversity of BPSK in Slow Rayleigh Fading using CC(2, 1, 7)

Figure 4.4: Two-Branch Diversity of BPSK in Slow Rayleigh Fading using PCC-3/4

Figure 4.5: Two-Branch Diversity of BPSK in Slow Rayleigh Fading using PCC-7/8

Figure 4.6: Three-Branch Diversity of BPSK in Slow Rayleigh Fading using CC(2, 1, 7)

Figure 4.7: Three-Branch Diversity of BPSK in Slow Rayleigh Fading using PCC=3/4

Figure 4.8: Three-Branch Diversity of BPSK in Slow Rayleigh Fading using PCC=7/8

4.1.4 Discussion

Coding without diversity offers no improvement as demonstrated in the curves. Convolutional codes rely on adjacent bits to correct an error. In the slow Rayleigh fading channel, burst errors occur due to deep fades. The result is that the convolutional code becomes occasionally overloaded and the BER performance deteriorates. More bit errors are introduced at the decoder output because substantial burst errors could not be corrected. The Viterbi decoder selected the incorrect path and that precipitated more errors than if coding had not been used. This leads to the subsequent sections in which block interleaving will be introduced into the communication system model. Steele [27, Ch. 4, pp. 390-400] examined CC(2, 1, 5) in slow Rayleigh fading and emphasized the coding gain achieved by using soft-decision decoding (the soft decision decoding was assumed to utilize infinite quantization levels from the demodulator) as opposed to hard-decision decoding. In Chapter 5, reliability weights will be assigned to the code symbols from each branch for the code-combining approach; a soft decision on each individual packet will be used, while a hard decision on the individual bits will be made.

The code-combining approach has the same performance as the post-Viterbi decoding selection scheme in Figure 4.3 - two-branch diversity with CC(2, 1, 7) employed. The cell-loss probability is greater than 10^{-3} for an average E_b/N_0 of 15 dB and would not be acceptable for most services [14]. The results shown in Figure 4.4 and 4.5 employ a punctured convolutional code of rate 3/4 and 7/8, respectively. The code-combining approach is approximately 1 dB inferior to the post-Viterbi decoding selection scheme at a probability of bit error of 10^{-3} in both these figures. For two diversity branches the performance of the code-combining approach was no better than, or in fact worse for rates 3/4 and 7/8, than post-Viterbi decoding selection diversity. This would be due to the fact that the errors occur in bursts and since the codewords from each diversity branch are assigned an equal weight, an overloaded or error-prone one will spread errors into one which may not be.

The results for three-branch diversity with CC(2, 1, 7) employed in Figure 4.6 show a cell-loss probability of 10^{-4} at an average E_b/N_0 of 15 dB when the code-combining approach is used. This is about 2 dB superior to the post-Viterbi decoding selection scheme. The average E_b/N_0 for the same cell-loss probability (for the code-combining

approach) is about 16 dB and 18 dB when a punctured convolutional code of rate 3/4 and 7/8 is used in Figures 4.7 and 4.8, respectively.

4.2 Performance in a Slow Rayleigh Fading Environment with Block Interleaving

Convolutional codes are designed to combat random independent errors which usually occur in a memoryless channel. However, fading channels cause errors to occur in bursts during the signal fades. Interleaving is deployed to disperse the burst errors when the received signal level fades, and to reduce the concentration of the errors that must be corrected by the channel code. If the errors are scrambled enough, then adjacent symbols appear independent to the decoder. Before a sequence of symbols is transmitted the symbols from several codewords are interleaved. Then when an error burst occurs the errors will be shared among the interleaved codewords and a less powerful code is required to correct them. As the interleaving period increases, the error performance can be expected to improve in the sense that noise bursts are more dispersed. On the other hand, the delays due to interleaving and deinterleaving increase.

The communication system used in this section is described in the figure below.

Here, we assume that block interleaving is used. The encoded symbols are written row by row into a matrix consisting of N columns where $N = 1/R_c * (\text{ATM cell length})$ and D rows. The interleaver used in the simulation is shown in Figure 4.10. The fades are assumed to be one coded ATM packet in duration as in the previous sections. The results obtained were for the rate-1/2 code employed. The results for rates 3/4 and 7/8 are not included since it was found that close to perfect interleaving would be required to achieve acceptable cell loss probabilities. The fade duration in this case is assumed to be $2 * 53$ bytes (ATM cell length) - the same fade duration employed without interleaving. The coded symbols are transmitted column by column. The symbol errors are generated in the matrix as shown by the dotted line.

Figures 4.11 to 4.13 show the results for two diversity branches when interleaving of depths 10, 20, and 100 coded ATM cells is used and the CC(2, 1, 7) is employed. The corresponding figure for two diversity branches without interleaving is Figure 4.3.

The probability of bit error for the code-combining approach provides a coding gain of about one dB at 10^{-3} for an interleaving depth of 10 ATM cells when compared with Figure 4.3 (code-combining with no interleaving). However, the cell-loss probability for the

code-combining approach is actually worse; this small amount of interleaving has the effect of spreading more errors amongst the cells that the convolutional code is unable to correct. Likewise, the error rates for the post-Viterbi-decoding selection scheme in Figure 4.11 worsens due to this distribution of bursts of errors amongst the cells. The code-combining approach in Figure 4.12 (interleaving depth of 20 cells) gives a coding gain of about 3 dB over the no interleaving case at the same probability of bit error, although the interleaving is still not sufficient and results in a higher cell loss probability as well. Interleaving of 100 cells (Figure 4.13) provides a cell-loss probability in the neighbourhood of less than 10^{-4} and a probability of bit error less than 10^{-6} for an average E_b/N_0 of 15 dB which is acceptable for a limited set of applications^[14]; yet this results in further delay.

Let us examine the delay for speech transmission in existing GSM systems for the purposes of comparison. The input rate for speech transmission on a full rate traffic channel is equal to 13 kbit/s, by blocks of 260 bits (i.e. one block every 20 ms)^[1, Ch. 4]. The coded block length is 456 bits. Blocks of full rate speech are interleaved on 8 bursts: the 456 bits of one block are split into 8 groups of 57 bits, each one carried by a different burst. A burst contains the contribution of two successive speech blocks. The corresponding delay is about 37.5 ms (see section 2.10 for more details of the interleaving and coding used in GSM). If this 13 kbit/s source is used in our case and interleaving of 10 ATM cells is used the incurred delay would be ~ 300 ms which would not be acceptable and the cell-loss probability displayed in Figure 4.11 would not be low enough. If a data source of 1 Mb/s is used the delay incurred with interleaving of 100 ATM cells would be ~ 40 ms which could be acceptable for some broadband applications, however, the cell-loss probability displayed in Figure 4.13 would not be good enough. The following section will examine these diversity schemes assuming perfect interleaving and will be compared as an upper bound against the results from this section.

Figure 4.11: Two-Branch Diversity of BPSK in Slow Rayleigh Fading using CC(2, 1, 7) and Interleaving of 10 Cells

Figure 4.12: Two-Branch Diversity of BPSK in Slow Rayleigh Fading using CC(2, 1, 7) and Interleaving of 20 Cells

Figure 4.13: Two-Branch Diversity of BPSK in Slow Rayleigh Fading using CC(2, 1, 7) and interleaving of 100 Cells

4.3 Performance in a Fast Rayleigh Fading Environment (i.e. perfect interleaving)

In this section the assumption is made that there is perfect interleaving and that all code symbol errors are independent. This is an idealization and is not realistic if we consider real-time applications like video traffic, voice and multimedia sessions which pose the most stringent performance requirements in terms of delay^[14]. However, it does give an upper bound in terms of cell-loss probability and it is useful for comparison with the limited interleaving results presented in the previous section.

The modifications made in the simulation would be in Step 2. Previously the SNR was assumed to be independent from packet to packet. Since perfect interleaving is considered here, the SNR will be assumed to be independent from symbol to symbol. The Rayleigh distributed values of instantaneous SNR are thus generated per code symbol as opposed to coded packet.

Figures 4.14 to 4.16 show the results for two-branch diversity in which the convolutional coding employed is rate $1/2$, $3/4$, and $7/8$, respectively. The corresponding figures for the slow fading case are figures 4.3 through 4.5. The post-Viterbi-decoding selection scheme as well as the other displayed curves show considerable improvement over the slow fading results. This would follow from the fact that the symbol errors are completely random and non-bursty and the convolutional code employed is powerful enough to correct most of them.

The code-combining results for two-branch diversity and CC(2, 1, 7) are compared in Figure 4.20. There is an increasing coding gain as expected in terms of the probability of bit error. However, we see that for interleaving depths of 10 and 20 coded cells the cell loss probability is actually worse than the no-interleaving case. As stated in the previous section this small amount of interleaving has the effect of spreading more errors amongst the cells that the convolutional code is unable to correct, thus corrupting a higher percentage of otherwise correctly received cells.

Figure 4.14: Two-Branch Diversity of BPSK in Fast Rayleigh Fading using CC(2, 1, 7)

Figure 4.15: Two-Branch Diversity of BPSK in Fast Rayleigh Fading using PC(1, 3/4)

Figure 4.16: Two-Branch Diversity of BPSK in Fast Rayleigh Fading using PCV-7/8

Figure 4.17: Three-Branch Diversity of BPSK in Fast Rayleigh Fading using CC(2, 1, 7)

Figure 4.18: Three-Branch Diversity of BPSK in Fast Rayleigh Fading using PC(1-3/4)

Figure 4.19: Three-Branch Diversity of BPSK in Fast Rayleigh Fading using PC(1-7/8)

FIGURE 4.20. Comparison of Code Combining in Slow Rayleigh Fading Using Different Depths of Interleaving for Two-Branch Diversity using CC(2, 1, 7)

a = uncoded branch
b = interleaving depth 10
c = no interleaving
d = interleaving depth 20
e = interleaving depth 100
f = full interleaving

4.4 Performance in a Slow Rayleigh Fading Environment with Correlated Diversity Branches

All the results up to this point have been premised on the assumption of independence between the fading on the diversity branches. It is clear that there will be many cases in which this cannot be achieved because of insufficient antenna spacing (due to siting limitation in spaced antenna diversity), etc. During our discussions Professor Falconer showed measurements^[53] of a system employing sectorized antennas such as that in ^[30]. There was a high degree of correlation between the signal strength of the line-of-sight antenna and adjacent antennas after normalizing the signal strengths. The purpose of this section is to analyze how correlation affects the examined diversity techniques. This section examines the packet diversity techniques in a slow Rayleigh fading environment as in Section 4.1, however, here the Rayleigh fading experienced on the diversity branches will be correlated.

Several authors have investigated diversity combining over correlated Rayleigh fading channels^[33, 35, 36, 37, 38]. The correlation in ^[35] was assumed to be equal as well as linear between all channels. For low diversity order, this is a valid assumption. For high order diversity, the assumption provides an upper bound for the channel correlation. The approach used by Packard^[33] to generate correlated Rayleigh variables is used in the simulation for this section. This is briefly described below:

- First, generate a Rayleigh variable Z_1 using the distribution in (4.1). This variable can then be resolved into two independent Gaussian variables X_1 and Y_1 ^[31]. We perform the following steps to accomplish this end.

1. Generate U_1 and U_2 , two independent random variables uniformly distributed in the unit interval.

2. Generate a Rayleigh variable:

$$Z_1 = -\Gamma \log U_1 \quad (\text{from equation 4.1}) \quad \text{and} \quad \theta = 2\pi U_2$$

3. Let $X_1 = \sqrt{Z_1} \cos \theta$ and $Y_1 = \sqrt{Z_1} \sin \theta$, since $Z_1 = \sqrt{X_1^2 + Y_1^2}$

Then X_1 and Y_1 are independent, zero-mean, variance = $\Gamma/2$ Gaussian random variables.

- Secondly, we consider two correlated Rayleigh variables Z_1 and Z_2 having the components (X_1, Y_1) and (X_2, Y_2) , respectively (with variances = σ^2). We assume a simple linear correlation

$$X_2 = cX_1 + u, \quad Y_2 = cY_1 + v \quad (\text{EQ 4.4})$$

where u and v are the uncorrelated portions of X_2 and Y_2 . It has been shown^[33] that the correlation coefficient ρ for Z_1 and Z_2 is given by

$$\rho = \frac{\pi}{4(4-\pi)} \left(c^2 + \frac{c^4}{16} + \frac{c^6}{64} + \dots \right) \approx c^2 \quad (\text{EQ 4.5})$$

We now consider the variables defined by the linear transformations:

$$\sqrt{2}\xi_1 = X_1 + X_2 = X_1(1+c) + u, \quad (\text{EQ 4.6})$$

$$\sqrt{2}\xi_2 = X_1 - X_2 = X_1(1-c) - u,$$

$$\sqrt{2}\eta_1 = Y_1 + Y_2 = Y_1(1+c) + v,$$

$$\sqrt{2}\eta_2 = Y_1 - Y_2 = Y_1(1-c) - v.$$

These variables are normally distributed since they are linear combinations of normally distributed variables^[35]. They are also uncorrelated as may be seen from the following:

$$2\overline{\xi_1 \xi_2} = \overline{X_1^2} - \overline{X_2^2} = \sigma^2 - \sigma^2 = 0,$$

$$2\overline{\eta_1\eta_2} = \overline{Y_1^2} - \overline{Y_2^2} = \sigma^2 - \sigma^2 = 0.$$

Since

$$2\overline{\xi_1\xi_2} = \overline{X_1^2}(1 - c^2) - \overline{u^2} = 0$$

then

$$\overline{u^2} = (1 - c^2)\sigma^2 \quad (\text{EQ 4.6})$$

and similarly

$$\overline{v^2} = (1 - c^2)\sigma^2 \quad (\text{EQ 4.7})$$

The uncorrelated portions of X_2 and Y_2 can then be described by the normally distributed variables u and v possessing the above variances. The simulation described in section 4.1 is then modified as follows:

1. Read the puncturing matrix file. Read the number of diversity branches $= N_{div}$ ($= 2$ for the cases examined). Read the average SNR per information bit, Γ , per branch.
2. Generate a sequence of Rayleigh distributed variables (variable Z_1 above) for one diversity branch. The instantaneous SNR is obtained from the PDF given in equation (4.1) for the given average SNR, Γ . Resolve this sequence of Rayleigh variables into its Gaussian components (X_1 and Y_1). The uncorrelated portions of X_2 and Y_2 , namely u and v are then generated to obtain Z_2 from the linear relation in equation (4.3).
3. Using the correlated sequences of instantaneous SNRs (which remain constant over each packet), we generate the corresponding probability of bit error given by equation (3.2) for BPSK.
4. Generate multiple copies ($= N_{div}$) of an ATM packet consisting of random bits.

5. Encode the copies of the packet using the (2, 1, 7) convolutional code and the appropriate puncturing matrix from Table 3.2.

6. Generate random errors amongst the independent copies using the error sequences generated in step (4). Add M "0" bits to the packets to properly terminate the encoder memory and the decoder trellis. M is the memory of the convolutional encoder.

7. Decode the copies of the packet independently using the Viterbi algorithm; count the decoded bit errors and lost cells and store these statistics for the no diversity case.

8. Select the decoded packet with the fewest errors and store the statistics for the post-Viterbi decoding selection case.

9. The CRC of the coded packets is checked. If all copies are in error then they are combined and decoded using the Viterbi algorithm and the code combining approach described earlier. The statistics are stored for the equal weights code combining case

10. Repeat steps 2. through 9.

The number of packets generated at the input was of the order 10^5 for most cases. The number of packets generated at the lower error rates was two orders of magnitude more than the loss probability. For example at a cell loss probability of 10^{-5} , 10^7 packets would be generated to obtain a reasonable confidence level.

The results shown in Figures 4.21 and 4.22 apply to two correlated diversity branches in slow Rayleigh fading with CC(2, 1, 7) employed. Figure 4.3 is the corresponding figure for the independent diversity branch case. We see from these figures that the diversity gain decreases as the correlation coefficient is increased. For example, in Figure 4.3 to achieve a bit error probability of 10^{-2} the required average E_b/N_0 is approximately 9 dB for either equal weights code combining or post-detection selection diversity; whereas with correlated diversity branches the required average E_b/N_0 to achieve the same bit error probability for either diversity scheme is approximately 12 dB and 13 dB for a correlation coefficient of 0.7 and 0.9, respectively. It has been shown for various diversity combining techniques that for a correlation coefficient less than 0.5 the performance loss is small

(dual diversity in [33], maximal-ratio combining in [35], and switch-diversity in [39]).

Another observation is that equal weights code combining becomes increasingly more superior to post-detection selection diversity at low average E_b/N_0 values as the correlation coefficient is increased. Although the second diversity branch provides limited information (because of the high correlation with the first diversity branch), the errors would be more evenly distributed between the branches, thereby preventing an error-prone branch to corrupt a somewhat less error-prone branch.

The results shown in Figures 4.23 and 4.24 apply to two correlated diversity branches in slow Rayleigh fading with a rate 3/4 punctured convolutional code. Figure 4.4 is the corresponding figure for the independent diversity branch case. As well, the results shown in Figures 4.25 and 4.26 apply to two correlated diversity branches in slow Rayleigh fading with a rate 7/8 punctured convolutional code and Figure 4.5 is the corresponding figure for the independent diversity branch case. We see in these cases as well that the diversity gain decreases as expected with increasing correlation between the diversity branches. Looking at the cell-loss probability for these higher rate cases, we see again that the equal weights code combining becomes increasingly more superior to post-detection selection diversity at low average E_b/N_0 values as the correlation coefficient is increased. Alternatively, at high average values of E_b/N_0 the equal weights code-combining approach is no worse than the post detection selection diversity case in terms of cell-loss probability.

Figure 4.21: Probability of Error for BPSK in Slow Rayleigh Fading using CC(2, 1, 7) and Two Correlated (correlation coefficient = 0.7) Diversity Branches

Figure 4.22: Probability of Error for BPSK in Slow Rayleigh Fading using CC(2, 1, 7) and Two Correlated (correlation coefficient = 0.9) Diversity Branches

Figure 4.23: Probability of Error for BPSK in Slow Rayleigh Fading using PCC(1/3/4) and Two Correlated (correlation coefficient = 0.7) Diversity Branches

Figure 4.24: Probability of Error for BPSK in Slow Rayleigh Fading using PCC(1/3/4) and Two Correlated (correlation coefficient = 0.9) Diversity Branches

Figure 4.25: Probability of Error for BPSK in Slow Rayleigh Fading using PC(17/8) and Two Correlated (correlation coefficient = 0.7) Diversity Branches

Figure 4.26: Probability of Error for BPSK in Slow Rayleigh Fading using PC(17/8) and Two Correlated (correlation coefficient = 0.9) Diversity Branches

4.5 Performance in a Log-normal Shadowing Channel with Slow Rayleigh Fading Superimposed

In this section we examine the effects of shadow fading on the channel. The code combining results are not expected to be significantly better in terms of cell loss probability based on the results obtained in section 4.1 for a slow Rayleigh fading channel.

The shadow fading is modeled as a log-normal distributed random variable with 0 dB mean and a standard deviation $\sigma=10$ dB.

The simulation is performed by the following steps for a given Ω (area mean signal-to-noise ratio) and a number of diversity branches: (similar steps as in ^[9] are taken)

- (1) A sample of the shadow fading envelope is generated, and the local mean signal-to-noise ratio Γ is computed (i.e., the signal-to-noise ratio averaged over Rayleigh fading)

- (2) Given the local mean signal-to-noise ratio Γ in dB, for each microdiversity branch, 1000 packets are considered. The instantaneous signal-to-noise ratio, γ , is obtained for each given packet using the Rayleigh probability density function given in (4.1). The average bit error probability encountered by each packet is then computed using (3.1).

- (3) The random errors for each coded packet are generated using the probability computed in step (2). Add M "0" bits to the packets to properly terminate the encoder memory and the decoder trellis. M is the memory of the convolutional encoder. Hard decision Viterbi decoding is performed on the independent packets; the decoded bit errors and lost cells are counted and stored for the no diversity case.

- (4) The decoded packet with the fewest errors is selected and the statistics for the post-Viterbi decoding selection case are stored. If the CRC of the coded packets for each diversity branch indicates an error, equal-weights code combining is done. The statistics for the code-combining case are stored.

- (5) Steps 2) through 4) are repeated 1000 times to find the corresponding instantaneous error probabilities.

(6) Steps 1) to 5) are repeated 1000 times to obtain the error probabilities averaged over shadow fading samples.

Figure 4.27 shows the results for two diversity branches and CC(2, 1, 7) employed. The results are comparable to those in [9] obtained for a BCH(31, 21, 5) code. There is no difference in terms of error probability between the diversity techniques here.

Figure 4.27: Two Branch Diversity in Log-normal Shadowing with Slow Rayleigh Fading Superimposed using CC(2, 1, 7)

4.6 Conclusions

The comparative performance of post-Viterbi decoding selection diversity and equal-weights code combining was evaluated over a slow Rayleigh fading channel for two and three independent diversity branches and punctured convolutional codes of rate 1/2, 3/4, and 7/8. It was seen that for two diversity branches the performance of the code-combining approach was no better than, or in fact worse for rates 3/4 and 7/8, than post-Viterbi decoding selection diversity. The errors occur in bursts and would be spread from error-prone packets into the packets from the other diversity branches when performing the

code combining. Interleaving of depths 10 to 100 coded ATM cells offered moderate improvements for the code-combining approach in slow fading, however, at the cost of considerable delay. Table 4.1 summarizes the comparison between the coding gains for the various coding rates under slow fading conditions for a bit error probability of 10^{-3} and a cell loss probability of 10^{-3} . (The entries "--" were off the simulation plots.)

TABLE 4.1. Coding Gains over a Rayleigh Fading Channel with Independent Diversity Branches

	E_b/N_0 Value		Coding Gain	
	BER = 10^{-3}	Cell Loss Prob. = 10^{-3}	BER = 10^{-3}	Cell Loss Prob. = 10^{-3}
No Coding	23.97 dB	29.5 dB	0 dB	0 dB
CC(2,1,7) R=1/2				
No Diversity	> 23.97 dB	--	NA	NA
2-branch PDSB	14.5 dB	17.9 dB	9.47 dB	11.6 dB
2-branch Code Comb.	14.5 dB	17.9 dB	9.47 dB	11.6 dB
3-branch PDSB	10 dB	13.3 dB	13.97 dB	16.2 dB
3-branch Code Comb.	9 dB	12.4 dB	14.97 dB	17.1 dB
PCC-3/4				
No Diversity	> 23.97 dB	--	NA	NA
2-branch PDSB	16.0 dB	18.57 dB	7.97 dB	10.93 dB
2-branch Code Comb.	16.9 dB	19.0 dB	7.07 dB	10.5 dB
3-branch PDSB	10.7 dB	14.0 dB	13.27 dB	15.5 dB
3-branch Code Comb.	10.7 dB	13.3 dB	13.27 dB	16.2 dB
PCC-7/8				
No Diversity	> 23.97 dB	--	NA	NA
2-branch PDSB	16.4 dB	19.3 dB	7.57 dB	10.2 dB
2-branch Code Comb.	17.6 dB	19.3 dB	6.37 dB	10.2 dB
3-branch PDSB	11.7 dB	15.0 dB	12.27 dB	14.5 dB
3-branch Code Comb.	11.7 dB	14.3 dB	12.27 dB	15.2 dB
Interleaving 10 Cells				
CC(2, 1, 7) R=1/2				
No Diversity	> 23.97 dB	--	NA	NA
2-branch PDSB	17.6 dB	--	6.37 dB	NA
2-branch Code Comb.	13.3 dB	--	10.67 dB	NA

	E_b/N_0 Value		Coding Gain	
	BER = 10^{-3}	Cell Loss Prob. = 10^{-3}	BER = 10^{-3}	Cell Loss Prob. = 10^{-3}
Interleaving 20 Cells CC(2, 1, 7) R=1/2				
No Diversity	--	--	NA	NA
2-branch PDSB	17.9 dB	--	6.07 dB	NA
2-branch Code Comb.	11.7 dB	--	12.27 dB	NA
Interleaving 100 Cells CC(2, 1, 7) R=1/2				
No Diversity	15.5 dB	--	8.47 dB	NA
2-branch PDSB	12.1 dB	19.5 dB	11.87 dB	10.0 dB
2-branch Code Comb.	6.9 dB	12.6 dB	17.07 dB	16.9 dB

We also examined the diversity techniques in a slow Rayleigh fading environment in which the diversity branches were correlated. It was interesting to note that as the correlation coefficient is increased the equal-weights code-combining outperformed the post-Viterbi decoding selection scheme. This makes sense since the weights used in the equal weights code-combining approach are most accurate or correct when the correlation coefficient is higher.

The performance of the code-combining approach can be improved with the use of accurate channel state information (similar to that employed in ^[43]) in determining reliability information for the combined packets. A hybrid-ARQ scheme will be examined in the next section in which reliability information is assigned to buffered packets which are then combined with retransmitted packets.

Chapter 5

Hybrid-ARQ Schemes in Rayleigh Fading

5.1 Introduction

It was demonstrated in the previous chapter that retransmissions may be preferable to the required amount of interleaving in slow Rayleigh fading channels.

This chapter will examine a type-II hybrid-ARQ scheme used in combination with rate-compatible punctured codes.

5.2 Performance in a Slow Rayleigh Fading Environment

Various ARQ schemes have been devised that maintain throughput at high channel bit error rates. Hybrid versions which employ error-correction coding have proven most effective. However, even these schemes fail to provide useful throughput at very high bit error rates, despite a non-zero channel capacity. Chase^[22] demonstrated that a selective-repeat ARQ link employing code combining can maintain useful throughput at high error rates (error rates approaching 50 %).

Automatic-repeat-request (ARQ) and forward-error-correction (FEC) are two techniques commonly used to maintain reliable communication over noisy channels. ARQ schemes offer very high reliability but their throughput efficiency falls off rapidly as the channel error rate increases. Several authors^[22] have suggested that a "hybrid" ARQ system which judiciously combines ARQ and FEC can offer performance superior to either technique alone.

In this section we consider the incorporation of code-combining into selective-repeat ARQ systems. The code-combining algorithm employed in the simulations is illustrated in the flowchart of Figure 5.1. ATM cells of the format in Figure 3.5 are decoded and their CRC is checked. We assume that there is an error-free feedback link to transmit ACK/NAK information.¹ Packets requiring retransmission are buffered and code-combined

---

1. In practice ACK/NAK information will be protected by low rate codes and the ARQ protocol will be capable of dealing with lost ACK/NAK information.

with subsequent transmissions, resulting in a code of rate- R/L , where L is the number of retransmissions

Forney^[21] considered “augmented outputs” from the Viterbi algorithm itself, such as the depth at which all paths are merged, the difference in length between the best two paths, and a list of the p best paths. References ^{[19],[20]} describe implementations of the “Soft-Output Viterbi Algorithm” or SOVA scheme in much the same vein. This is a Viterbi algorithm which uses hard (or soft) decisions to calculate its metrics, but also decides in a soft way by providing reliability information together with the output bits (see Section 2.8 for more details). These reliability weights are determined for buffered packets requiring retransmission.

Figures 5.2 and 5.3 show the error rates using this scheme for up to two and three retransmissions respectively and a rate-7/8 convolutional code. The curves displayed are similar to Figures 4.5 and 4.8 with the exception that the code-combining is not done on parallel diversity branches but on buffered packets. The code-combining is done using the buffered packets in this instance, if the CRC protecting the information bits does not check. There isn't an additional CRC protecting the coded ATM packet in this case, since the buffered packets have been decoded and the code-combining will therefore be “non-blind”. We see from Figure 5.3 that for error-free transmission, one cell in 10^5 would need to be retransmitted three times at an E_b/N_o of 19 dB. The weighted code-combining approach does not offer any significant performance improvements over the equal-weights code-combining approach for up to two or three retransmissions. The code-combining approach is approximately one dB superior to the post-Viterbi-decoding selection approach at a cell loss probability of 10^{-3} for up to two transmissions. This difference is approximately 1.5 dB at a cell loss probability of 10^{-3} for up to three transmissions. Figures 5.4 and 5.5 show the error rates using up to two and three retransmissions respectively and a rate-1/2 convolutional code. The curves displayed are similar to Figures 4.3 and 4.6 with the exception that the code-combining is not done on parallel diversity branches but on buffered packets. We see from Figure 5.5 that for error-free transmission, one cell in 10^5 would need to be retransmitted three times at an E_b/N_o of 15 dB. The weighted code-combining approach does not offer any significant performance improve-

ments over the equal-weights code-combining approach for up to two and three retransmissions.

Another observation that was made elsewhere^[22] is fixed-rate ARQ (= post-Viterbi-decoding selection in Figure 5.2 -5.5) extends the region of useful throughput to high channel error rates when a low rate code is used, while imposing a penalty on the throughput of low error rates. The code combining approach is adaptive since each retransmission results in a lower rate code of R/L , where L is the number of retransmissions. In addition the code repeated L times has a minimum free distance L times that of the basic code which is nearly optimal^[23]. Since RCPC codes are being used as well, the code rate can be further decreased if the channel is too noisy by puncturing fewer bits.

Figure 5.1: Flowchart of Hybrid-ARQ Employing Code Combining

Figure 5.2: Error Rates for BPSK Using Equal Weights Code Combining with up to Two Retransmissions and PCC-7/8

Figure 5.3: Error Rates for BPSK Using Non-Equal Weights Code Combining with up to Three Retransmissions and PCC-7/8

Figure 5.4: Error Rates for BPSK Using Non-Equal Weights Code Combining with up to Two Retransmissions and CC(2, 1, 7)

Figure 5.5: Error Rates for BPSK Using Non-Equal Weights Code Combining with up to Three Retransmissions and CC(2, 1, 7)

5.3 Hybrid ARQ using RCPC Codes in an Ideally Interleaved Rayleigh Channel

In this section an ARQ/FEC scheme is analyzed which uses rate compatible punctured convolutional codes (RCPC) codes.

It has been shown that packetized data transmission for mobile satellite or radio services achieves reliable transmission and good throughput in a fading environment if a combination of automatic repeat request (ARQ) and forward error correction (FEC) is used^[16]. Rate compatible codes have been found which can be used for successive parity transmission. The same Viterbi decoder is used for repeated ML-decoding of RCPC codes with decreasing rate, thus minimizing the number of transmitted code bits.

RCPC codes are obtained by puncturing a $1/N$ convolutional code in a rate compatible manner to obtain a family of codes with rate

$$R = \frac{P}{P+1} \quad l = 1, \dots, P(N-1) \quad (\text{EQ 5.1})$$

P is the period of the puncturing. All code bits of a higher rate code are used as code bits of a lower rate code. This allows successive parity transmission. For instance, with $P = 7$ and $N = 2$ the code rate is varied between $7/8$ and $1/2$. RCPC codes are almost as good as the best known convolutional codes with the same complexity of the VA^[16]. However, the same encoder and the same Viterbi-decoder can be used, only the puncturing rule or the multiplexer rule is changed. The principle of the proposed ARQ/FEC scheme is not to repeat information or parity bits if the transmission is unsuccessful, but to transmit additional code bits of a lower rate RCPC-code until the code is powerful enough to enable decoding. This includes several decoding attempts on the receive side, which seems especially feasible for applications in mobile or low rate systems where the data rate is in the kbit/s range and the VA decoder would run on a chip at a higher rate.

The $(2, 1, 7)$ code is used as the mother code in this section. The weight spectra of the RCPC codes generated from this mother code are replicated below from^[15]. The upper

bound on the error event and bit error probabilities of any rate k/n codes with Viterbi decoding on discrete memoryless channels are given by [24]:

$$P \leq \frac{1}{k} \sum_{d=d_{free}}^{\infty} a_d P_d \quad (\text{EQ 5.2})$$

$$P_b \leq \frac{1}{k} \sum_{d=d_{free}}^{\infty} c_d P_d \quad (\text{EQ 5.3})$$

where d_{free} is the minimum free distance of the code, a_d is the number of incorrect paths of Hamming weight d for $d \geq d_{free}$ that diverge from the correct path and re-emerge with it at some later stage. c_d is the total number of error bits produced by the incorrect paths and P_d is the probability of picking an incorrect path in the Viterbi decoding process. For coherent BPSK in AWGN

$$P_d = Q\left(\sqrt{2d(k/n)(E_b/N_o)}\right) \quad (\text{EQ 5.4})$$

To choose good codes, one should maximize d_{free} and minimize the terms a_d and c_d .

TABLE 5. Weight Spectra of RCPC Codes Generated from (2, 1, 7) Mother Code and Puncturing Period $P = 7$; (M = Code memory, G_1 and G_2 are the generator polynomials)

Mother Code						
M	G_1	G_2	Puncturing Matrix $A(l)$	RCPC code rate- $P/(P+l)$	d_{free}	(a_d , for $d=d_f, d_f+1, d_f+2, \dots$) [c_d , for $d=d_f, d_f+1, d_f+2, \dots$]
7	133	171	1111111 1110111	7/13	8	(6,43,80,174,423,1102,3213) [16,152,382,1086,3063,9065,28767]
			1111111 1110110	7/12	7	(4,37,102,220,622,1967,6037) [10,141,520,1545,5109,17829,61636]
			1111111 1010110	7/11	6	(4,28,94,296,957,3411) [10,119,496,2170,8391,34229]
			1111111 1000110	7/10	5	(4,31,131,535,2282,9649) [14,174,889,4915,25538,126149]
			1111111 1000100	7/9	4	(2,38,202,1149,6761,39572) [5,279,1889,13785,100491,694005]
			1111111 1000000	7/8	3	(2,42,468,4939,52821) [14,389,6792,97243,1317944]

The ARQ/FEC Protocol with RCPC Codes follows [16]:

The ARQ/FEC protocol performs the following steps while transmitting a block I of information as shown in Figure 5.6

Encoding:

- (i) Add n_c parity bits to form an error detection code C_o .
- (ii) Add M "0" or known bits to properly terminate the encoder memory and the decoder trellis for the "information" block $(n + n_c)$. M is the memory of the convolutional encoder.
- (iii) Use the $1/N$ encoder to encode the convolutional code and store it at the transmitter, possibly in a matrix as in Figure 5.6.

Transmitting and decoding:

- (1) Transmit the first P columns of the matrix.
  - (1a) Invert the convolutional code using generator G_1^{-1} .
  - (1b) Check the syndrome of code C_o . If it is zero then output I and send ACK to transmitter.
- (2) Transmit additional columns of the matrix up to index l according to the puncturing table.
- (3) Decode RCPC-Code of rate $P/(P+1)$ with Viterbi algorithm.
- (4) Check the syndrome^[3, p. 410] of Code C_o . If the syndrome is zero, output I and send ACK to the transmitter. If the syndrome of C_o is not zero increase l and repeat steps (2) to (4)

FIGURE 5.6 ARQ/FEC Encoding Scheme with RCPC Codes^[16]

The throughput is

$$R_{av} = \frac{n}{n + N_c + M} \cdot \frac{P}{P + l_{av}} \quad (\text{EQ 5.5})$$

$$R_{av} \approx \frac{n}{n + M} \cdot \frac{P}{P + l_{av}} \quad (\text{EQ 5.6})$$

where,

P = Puncturing period = 7

n = ATM cell = 424 bits

2 of/de 2

PM-1 3½"x4" PHOTOGRAPHIC MICROCOPY TARGET
NBS 1010a ANSI/ISO #2 EQUIVALENT

PRECISION^{SM} RESOLUTION TARGETS

$$N_c (< n) = \text{CRC}$$

l_{av} = average number of retransmissions

M = Memory of code (= constraint length - 1)

The throughput can be interpreted as the average number of additionally transmitted columns. $P/(P + l_{av})$ is the effective code rate of the FEC code. In Figure 5.7 and 5.8 the performance of a fully interleaved Rayleigh channel with BPSK transmission is given. In this case decoding starts with the 7/8 code described in Table 5.1 and ends with the mother 1/2 code.

5.4 Discussion

In reference ^[16] relatively high throughputs could be attained for low frame error rates. Specifically, a frame error rate of less than 10^{-6} and a throughput of 0.7 could be obtained with $E_s/N_0 = 10$ dB. The frame used had $n = 8 \times 48$, $n_c = 34$. These were obtained with full channel state information, though.

In our case without any channel state information used, we see that even with 6 retransmissions the cell loss probability (Figure 5.8) is still high at < 5 dB E_s/N_0 . It has also been assumed that perfect interleaving is employed. We also see at these low E_s/N_0 values that the throughput levels off at 1/2 or the rate of the mother code. In other words we require $l = 61$ transmissions. It would then be preferable to either start with a lower rate code in this case or retransmit the entire packet instead of transmitting additional parity bits.

FIGURE 5.7. Hybrid ARQ Using RCPC Codes in an Interleaved Rayleigh Channel; Mother Code: (2, 1, 7)

FIGURE 5.8. Hybrid ARQ Using RCPC Codes in an Interleaved Rayleigh Channel; Mother Code: (2, 1, 7)

Chapter 6

Conclusions and Suggestions for Future Work

6.1 Conclusions

A simulation tool to analyze the performance of antenna diversity techniques operating on a packet-basis was developed and tested. The diversity techniques of interest were code diversity combining and ideal post-Viterbi decoding selection diversity which could be used as an approximation to ARQ (since the post detection diversity scheme is ideal and the packet with the fewest errors is selected). The codes used throughout the simulations were punctured convolutional codes.

It was demonstrated that the code-combining approach was superior to the post-detection scheme in a two- and three-branch diversity AWGN channel.

The packet diversity techniques were then compared over a slow Rayleigh fading channel with two and three diversity branches and punctured convolutional codes of rate $1/2$, $3/4$, and $7/8$. It was seen that for two diversity branches the performance of the code-combining approach was no better, or in fact worse for rates $3/4$ and $7/8$, than post-Viterbi decoding selection diversity for no interleaving. The diversity schemes were then examined using perfect interleaving for rates $1/2$, $3/4$, $7/8$. The coding gain attained for code-diversity combining over no-diversity and post-Viterbi decoding selection diversity was of the same order as that obtained in the AWGN case. Limited interleaving was also examined for a coding rate of $1/2$. The amount of interleaving required was at least 100 coded ATM cells. The simulation was then modified by correlating the diversity branches. The diversity gain decreased as expected with increasing correlation. It was also noted that as the correlation coefficient was increased the equal weights code-combining outperformed the post-Viterbi decoding selection scheme. This suggested the use of more accurate weights assigned to the combined packets in the code-combining approach.

The final chapter compared the diversity techniques using a "SOVA" approach in a hybrid-ARQ scheme. The weighted code-combining approach did not offer any signifi-

cant performance improvements over the equal-weights code-combining approach for up to two or three retransmissions when a rate-7/8 or a rate-1/2 convolutional code was used. An ARQ scheme using rate-compatible punctured convolutional codes was then analyzed in which additional parity bits are transmitted for each required retransmission. It was seen that without any channel state information high cell-loss probability was the outcome for $E_s/N_0 < 5$ dB. Complete retransmission of the packet would then be preferable or alternately a lower initial code rate should be used.

Improvements to the code-combining approach could be implemented by using channel-state information to erase unreliable symbols^[50]. This approach would require less interleaving in slow Rayleigh fading conditions for the code-combining approach to surpass the post-Viterbi decoding selection scheme.

Future work would be to modify the slow fading channel assumption in Chapter 4 to be frequency selective by making the channel bit error probability a function of delay spread as well ^[9]. The assumption that the instantaneous SNR values are independent from frame to frame could also be modified by correlating these random variables using the “low-pass filter” approach briefly described in Chapter 4.

It should be noted that the code-combining scheme can be easily implemented by digital devices and is nearly as simple as selection combining. The implementation of the decoder is fairly simple as well with this approach.

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