

Eldad Perahia and Robert Stacey

Next Generation Wireless LANs

Throughput, Robustness, and Reliability in 802.11n

CAMBRIDGE

Next Generation Wireless LANs

If you've been searching for a way to get up to speed quickly on IEEE 802.11n without having to wade through the entire standard, then look no further. This comprehensive overview describes the underlying principles, implementation details, and key enhancing features of 802.11n. For many of these features, the authors outline the motivation and history behind their adoption into the standard. A detailed discussion of the key throughput, robustness, and reliability enhancing features (such as MIMO, 40 MHz channels, and packet aggregation) is given, in addition to a clear summary of the issues surrounding legacy interoperability and coexistence. Advanced topics such as beamforming and fast link adaption are also covered. With numerous MAC and physical layer examples and simulation results included to highlight the benefits of the new features, this is an ideal reference for designers of WLAN equipment, and network managers whose systems adopt the new standard. It is also a useful distillation of 802.11n technology for graduate students and researchers in the field of wireless communication.

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**To my wife Sarah and our son Nathan
– Eldad Perahia**

**To my father, who nurtured and guided an inquiring mind
– Robert Stacey**

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Foreword

The first version of the 802.11 standard was ratified in 1997 after seven long years of development. However, initial adoption of this new technology was slow, partly because of the low penetration of devices that needed the “freedom of wireless.”

The real opportunity for 802.11 came with the increased popularity of laptop computers just a few years later. This popularity brought a rapidly growing user base wanting network connectivity not only while connected to an Ethernet cable at home or at work, but also in between: in hotels, airports, conference centers, restaurants, parks, etc. 802.11 provided a cheap and easy way to make laptop mobility a reality for anyone who wanted it.

However, technology by itself is rarely sufficient, particularly in the networking space, where interoperability of devices from multiple vendors is almost always the key to market success. Having been formed as WECA in 1999, the Wi-Fi Alliance was ready to provide certification of multi-vendor interoperability.

With the right technology from the IEEE 802.11 Working Group, certified interoperability from the Wi-Fi Alliance, and a real market need based on a growing installed base of laptops, the conditions were ripe for the Wi-Fi market to take off, and indeed it did. By 2007 virtually every new laptop contains Wi-Fi as standard equipment. More importantly, and unlike some other “successful” wireless technologies, many of these devices are used regularly. With this wide use came a growing understanding of the power of cheap, easy-to-deploy, and easy-to-manage interoperable Wi-Fi networks.

The natural next step was for people to ask, “What else can we use Wi-Fi for?” The answer is increasingly becoming “everything, everywhere!” Not just laptops, but now almost anything mobile and even many fixed devices contain Wi-Fi, and they are used in a phenomenal range of applications, including data, voice, games, music, video, location, public safety, vehicular, etc. In 2007, more than 300 million Wi-Fi devices were shipped. By 2012, some analysts are forecasting that more than one billion Wi-Fi devices will be shipped every year.

The 2.4 GHz 802.11b 11 Mb/s DSSS/CCK PHY and the basic 802.11 contention-based MAC provided the basis for a great industry. However, the rapid growth of the Wi-Fi market challenged the capabilities of the technology. It was not long before better security (802.11i certified by the Wi-Fi Alliance as WPA/WPA2™) and better Quality of Service (802.11e certified by the Wi-Fi Alliance as WMM™ and WMM Power Save) were defined, certified, and deployed.

It was also not long before higher data rates were demanded for greater data density and to support the many new and exciting devices and applications. 802.11a, providing 54 Mbps based on OFDM in the 5 GHz band, failed to garner significant support because two radios were required to maintain backward compatibility with 2.4 GHz 802.11b devices; the cost of two radios was often too high. The real success story was 802.11g, which provided 54 Mbps based on OFDM in the 2.4 GHz band in a way that was backward-compatible with 802.11b.

The success of 802.11g drove the use of Wi-Fi to new heights and expanded the demands on the technology yet again; everyone wanted more. Fortunately, the technology continued to develop and in 2002 the IEEE 802.11 Working Group started defining the next generation of PHY and MAC features as part of 802.11n. 802.11n will define mechanisms to provide users some combination of greater throughput, longer range and increased reliability, using mandatory and optional features in the PHY (including MIMO technology and 40 MHz channels) and the MAC (including more efficient data aggregation and acknowledgments).

Interestingly, 802.11n operates in both the 2.4 GHz and 5 GHz bands. It is expected that 5 GHz operation will be more popular than when 802.11a was introduced, because 2.4 GHz is now more congested, the number of available channels in the 5 GHz band has been expanded with the introduction of DFS and TPC technology, there is more need for high throughput 40 MHz channels, and the cost of dual-band radios has decreased.

The 802.11n standard is not yet complete, and is unlikely to be ratified by the IEEE until at least mid 2009. Until August 2006, the Wi-Fi Alliance had a policy to not certify 802.11n products until the standard was ratified. However, some vendors decided the market could not wait for ratification of the 802.11n standard and started releasing pre-standard products. These products were often not interoperable at the expected performance levels because they were not based on a common interpretation of the draft 802.11n specification. The problem for the Wi-Fi Alliance was that these products were adversely affecting the reputation of Wi-Fi. The Wi-Fi Alliance decided the only way forward was to certify the basic features of 802.11n from a pre-standard draft. Such a decision is not without precedent. In 2003, certification of WPA started before the 802.11i standard was ratified and in 2004 certification of WMM started before 802.11e was ratified. The Wi-Fi Alliance commenced certification of 802.11n draft 2.0 on 26 June 2007.

The decision has turned out to be the right one for the industry and for users. The Wi-Fi CERTIFIED 802.11n draft 2.0 programme has been remarkably successful, with more than 150 products certified in less than five months. This represents a significantly higher number of certified products than for the 802.11g programme during a similar period after launch. The Wi-Fi Alliance's certification program has helped ensure interoperability for the many products that will be released before the ratification of the 802.11n standard. This is particularly important given that the likely ratification date of the 802.11n standard has been extended by more than a year since the decision to start a certification program was announced by the Wi-Fi Alliance. The next challenge for the Wi-Fi Alliance is to ensure a backward-compatible transition path from the 802.11n draft 2.0 as certified by the Wi-Fi Alliance to the final ratified standard.

Standards are never the most accessible of documents. The 802.11 standard is particularly difficult to understand because it has been amended so many times by different groups and editors over a long period. A draft amendment to the standard, such as 802.11n D2.0, is even harder to interpret because many clauses are still being refined and the refinement process often has technical and political aspects that are only visible to those participating full time in the IEEE 802.11 Working Group.

Books like this one are invaluable because they provide the details and the background that allow readers to answer the questions, "What is likely to be in the final standard and how does it work?" Eldad and Robert should be congratulated on taking up the challenge.

Dr. Andrew Myles
Chairman of the BoD
Wi-Fi Alliance
6 December 2007

Preface

Having worked on the development of the 802.11n standard for some time, we presented a full day tutorial on the 802.11n physical layer (PHY) and medium access control (MAC) layer at the IEEE Globecom conference held in San Francisco in December 2006. Our objective was to provide a high level overview of the draft standard since, at the time, there was very little information on the details of the 802.11n standard available to those not intimately involved in its development. After the tutorial, we were approached by Phil Meyler of Cambridge University Press and asked to consider expanding the tutorial into a book.

Writing a book describing the standard was an intriguing prospect. We felt that a book would provide more opportunity to present the technical details in the standard than was possible with the tutorial. It would fill the gap we saw in the market for a detailed description of what is destined to be one of the most widely implemented wireless technologies. While the standard itself conveys details on what is needed for interoperability, it lacks the background on why particular options should be implemented, where particular aspects came from, the constraints under which they were designed, or the benefit they provide. All this we hoped to capture in the book. The benefits various features provide, particularly in the physical layer, are quantified by simulation results. We wanted to provide enough information to enable the reader to model the physical layer and benchmark their simulation against our results. Finally, with the amended standard now approaching 2500 pages, we hoped to provide an accessible window into the most important aspects, focusing on the throughput and robustness enhancements and the foundations on which these are built.

The book we came up with is divided into three parts. The first part covers the physical layer (PHY), and begins with background information on the 802.11a/g OFDM PHY on which the 802.11n PHY is based and interoperates, and proceeds with an overview of spatial multiplexing, the key throughput enhancing technology in 802.11n. This is followed by details on exactly how high throughput is achieved in 802.11n using spatial multiplexing and wider, 40 MHz channels. This in turn is followed by details on robustness enhancing features such as receive diversity, spatial expansion, space-time block codes, and low density parity check codes.

The second part covers the medium access control (MAC) layer. This part provides background on the original 802.11 MAC as well as the 802.11e quality of service (QoS) enhancements. It gives an overview of why changes were needed in the MAC to achieve higher throughput, followed by details on each of the new features introduced. Given the large installed base of 802.11 devices, coexistence and interoperability are considered

crucial to the smooth adoption of the standard. To this end, the book provides a detailed discussion on features supporting coexistence and interoperability.

In the third part we provide details on two of the more advanced aspects of the standard, transmit beamforming and link adaptation. These topics are covered in a section of their own, covering both the PHY and the MAC.

Writing this book would not have been possible without help from our friends and colleagues. We would like to thank Thomas (Tom) Kenney and Brian Hart for reviewing the PHY portion of the book and Solomon Trainin, Tom Kenney, and Michelle Gong for reviewing the MAC portion of the book. They provided insightful comments, suggestions, and corrections that significantly improved the quality of the book.

One of the goals of this book is to provide the reader with a quantitative feel of the benefit of the PHY features in the 802.11n standard. This would have been impossible without the extensive simulation support provided to us by Tom Kenney. He developed an 802.11n PHY simulation platform that includes most of the 802.11n PHY features and is also capable of modeling legacy 802.11a/g. The simulation includes all the 802.11n channel models. Furthermore, Tom modeled receiver functionality such as synchronization, channel estimation, and phase tracking. The simulation also included impairments such as power amplifier non-linearity and phase noise to provide a more realistic measure of performance.

The simulation supports both 20 MHz and 40 MHz channel widths. With the 40 MHz simulation capability, Tom generated the results given in Figure 5.8 in Section 5.1.5 modeling MCS 32 and Figure 5.9 in Section 5.1.7 which illustrates the range and throughput improvement of 40 MHz modes. With the MIMO/SDM capability of the simulation in both AWGN channel and 802.11n channel models, Tom produced the results for Figures 5.12–5.15 in Section 5.3. By designing the simulation with the flexibility to set the transmitter and receiver to different modes, he also produced the results given in Figure 5.18 in Section 5.4 modeling the behavior of a legacy 802.11a/g device receiving a GF transmission. Tom also incorporated short guard interval into the simulation with which the results for sensitivity to time synchronization error in Figures 5.20–5.22 in Section 5.5 were generated.

Tom designed the simulation with the ability to select an arbitrary number of transmitter and receiver antennas independent from the number of spatial streams. Using this capability he produced the results for receive diversity gain in Figures 6.2–6.4 in Section 6.1 and spatial expansion performance in Figures 6.5 and 6.6 in Section 6.2. Tom also incorporated space-time block coding and low density parity check coding into the simulation and generated the results given in Figures 6.8, 6.9, 6.14, 6.15, and 6.16 in Section 6.3 and Figure 6.24 in Section 6.4.

To accurately model the performance of a transmit beamforming system, it is important to include aspects like measurement of the channel state information, beamforming weight computation, and link adaptation. Tom incorporated all of these functions into the simulation to generate the waterfall curves in Figures 12.11–12.16 and the throughput curves in Figures 12.17 and 12.18 in Section 12.18.

We sincerely hope our book provides you with insight and a deeper understanding of the 802.11n standard and the technology upon which it is built.

Abbreviations

μ s	microseconds
2G	second generation (cellular)
3G	third generation (cellular)
AC	access category
ACK	acknowledgement
ADC	analog-to-digital converter
ADDBA	add block acknowledgement
ADDTS	add traffic stream
AGC	automatic gain control
AID	association identifier
AIFS	arbitration inter-frame space
A-MPDU	aggregate MAC protocol data unit
A-MSDU	aggregate MAC service data unit
AoA	angle of arrival
AoD	angle of departure
AP	access point
APSD	automatic power save delivery
A-PSDU	aggregate PHY service data unit
AS	angular spectrum
ASEL	antenna selection
AWGN	additive white Gaussian noise
BA	block acknowledgement
BAR	block acknowledgement request
BCC	binary convolution code
BF	beamforming
BICM	bit interleaved coded modulation
bps	bits-per-second
BPSCS	coded bits per single carrier for each spatial stream
BPSK	binary phase shift keying
BSS	basic service set
BSSID	BSS identifier
BW	bandwidth
CBPS	coded bits per symbol
CBPSS	coded bits per spatial stream

CBW	channel bandwidth
CCA	clear channel assessment
CCDF	complementary cumulative distribution function
CCK	complementary code keying
CFP	contention free period
CP	contention period
CRC	cyclic redundancy code
CS	carrier sense
CSD	cyclic shift diversity
CSI	channel state information
CSMA	carrier sense multiple access
CSMA/CA	carrier sense multiple access with collision avoidance
CSMA/CD	carrier sense multiple access with collision detection
CTS	clear to send
CW	contention window
DA	destination address
DAC	digital-to-analog converter
dB	decibels
dBc	decibels relative to carrier
dBi	decibels isotropic relative to an antenna
dBm	decibel of measured power referenced to one milliwatt
DBPS	data bits per OFDM symbol
dBr	dB (relative)
DC	direct current
DCF	distributed coordination function
DELBA	delete block acknowledgement
DIFS	DCF inter-frame space
DLS	direct link session
DS	distribution system
DSL	digital subscriber line
DSSS	direct sequence spread spectrum
DTIM	delivery traffic indication message
DVD	digital versatile disc
EDCA	enhanced distributed channel access
EIFS	extended inter-frame space
ERP	enhanced rate PHY
ESS	extended service set
ETSI	European Telecommunications Standards Institute
EVM	error vector magnitude
EWC	Enhanced Wireless Consortium
FCC	Federal Communications Commission
FCS	frame check sequence
FEC	forward error correction
FFT	fast Fourier transform

FHSS	frequency hopped spread spectrum
FS	free space
FTP	file transfer protocol
GF	Greenfield
GF-HT-STF	Greenfield High Throughput Short Training field
GHz	gigahertz
GI	guard interval
GIF	graphics interchange format
GPS	global positioning system
HC	hybrid coordinator
HCCA	HCF controlled channel access
HCF	hybrid coordination function
HEMM	HCCA, EDCA mixed mode
HT	high throughput
HTC	high throughput control
HT-DATA	High Throughput Data field
HT-LTF	High Throughput Long Training field
HTSG	High Throughput Study Group
HT-SIG	High Throughput Signal field
HT-STF	High Throughput Short Training field
HTTP	hypertext transfer protocol
Hz	Hertz
IBSS	independent basic service set
IC	integrated circuit
IDFT	inverse discrete Fourier transform
IEEE	Institute of Electrical and Electronic Engineers
IFFT	inverse fast Fourier transform
IFS	inter-frame space
IP	Internet Protocol
IPv6	Internet Protocol version 6
IR	infrared
ISI	inter-symbol interference
ISM	industrial, scientific, and medical
JPEG	Joint Photographic Experts Group
kHz	kilohertz
km/h	kilometers per hour
LAN	local area networking
LDPC	low density parity check
LLC	logical link control
L-LTF	Non-HT (Legacy) Long Training field
LNA	low noise amplifier
LOS	line-of-sight
LSB	least significant bit
L-SIG	Non-HT (Legacy) Signal field

L-STF	Non-HT (Legacy) Short Training field
LTF	Long Training field
m	meters
MAC	medium access control
MAI	MRQ or ASEL indication
MAN	metropolitan area networking
Mbps	megabit per second
MCS	modulation and coding scheme
MF	mixed format
MFB	MCS feedback
MFSI	MCS feedback sequence indication
MHz	megahertz
MIB	management information base
MIMO	multiple-input multiple-output
ML	maximum likelihood
MMPDU	MAC management protocol data unit
MMSE	minimum mean-square-error
MPDU	MAC protocol data unit
MPEG	Moving Picture Experts Group
MRC	maximal-ratio combining
MRQ	MCS request
Msample/s	mega-samples per second
MSB	most significant bit
MSDU	MAC service data unit
MSE	mean-square-error
MSFI	MCS feedback sequence identifier
MSI	MCS request sequence identifier
NAV	network allocation vector
NDP	null data packet
NF	noise figure
NLOS	non-line-of-sight
nsec	nanosecond
OBO	output back-off
OBSS	overlapping BSS
OFDM	orthogonal frequency division multiplexing
OSI	open systems interconnection
PA	power amplifier
PAR	project authorization request
PAS	power angular spectrum
PC	point coordinator
PCF	point coordination function
PCO	phased coexistence operation
PDU	protocol data unit
PER	packet error rate

PHY	physical layer
PIFS	PCF inter-frame space
PLCP	physical layer convergence procedure
PPDU	PLCP protocol data unit
ppm	parts per million
PSD	power spectral density
PSDU	PLCP service data unit
PSMP	power-save multi-poll
PSMP-DTT	PSMP downlink transmission time
PSMP-UTT	PSMP uplink transmission time
QAM	quadrature amplitude modulation
QoS	quality of service
QPSK	quadrature phase shift keying
R	code rate
RA	receiver address
RD	reverse direction
RDG	reverse direction grant
RF	radio frequency
RIFS	reduced inter-frame space
RMS	root-mean-square
RSSI	received signal strength indication
RTS	request to send
Rx	receive
SA	source address
SAP	service access point
SCP	secure copy protocol
SDM	spatial division multiplexing
SDU	service data unit
SE	spatial expansion
SIFS	short inter-frame space
SIG	Signal field
SIMO	single-input, multiple-output
SISO	single-input, single-output
SMTP	simple mail transfer protocol
SNR	signal-to-noise ratio
SOHO	small-office, home-office
SS	spatial stream
SSC	starting sequence control
SSID	service set identifier
SSN	starting sequence number
STA	station
STBC	space-time block coding
STF	Short Training field
STS	space-time stream

SVD	singular value decomposition
SYM	symbol
TA	transmitter address
TBTT	target beacon transmission time
TC	traffic category
TCLAS	traffic classification
TCM	trellis coded modulation
TCP	transmission control protocol
TDD	time division duplexing
TGn	Task Group n
TGy	Task Group y
TID	traffic identifier
TIFF	tagged image file format
TRQ	training request
TS	traffic stream
TSID	traffic stream identifier
TSPEC	traffic specification
TV	television
Tx	transmit
TxBF	transmit beamforming
TXOP	transmit opportunity
TXTIME	transmit time
UDP	user datagram protocol
USA	United States of America
VoIP	voice over IP
VPN	virtual private network
WEP	wired equivalent privacy
WFA	Wi-Fi Alliance
WLAN	wireless local area network
WM	wireless medium
WNG SC	Wireless Next Generation Standing Committee
WWiSE	world wide spectral efficiency
XOR	exclusive-or
ZF	zero-forcing
ZIP	ZIP file format

1 Introduction

Wireless local area networking has experienced tremendous growth in the last ten years with the proliferation of IEEE 802.11 devices. Its beginnings date back to Hertz's discovery of radio waves in 1888, followed by Marconi's initial experimentation with transmission and reception of radio waves over long distances in 1894. In the following century, radio communication and radar proved to be invaluable to the military, which included the development of spread spectrum technology. The first packet-based wireless network, ALOHANET, was created by researchers at the University of Hawaii in 1971. Seven computers were deployed over four islands communicating with a central computer in a bi-directional star topology.

A milestone event for commercial wireless local area networks (WLANs) came about in 1985 when the United States Federal Communications Commission (FCC) allowed the use of the experimental industrial, scientific, and medical (ISM) radio bands for the commercial application of spread spectrum technology. Several generations of proprietary WLAN devices were developed to use these bands, including WaveLAN by Bell Labs. These initial systems were expensive and deployment was only feasible when running cable was difficult.

Advances in semiconductor technology and WLAN standardization with IEEE 802.11 led to a dramatic reduction in cost and the increased adoption of WLAN technology. With the increasing commercial interest, the Wi-Fi Alliance (WFA) was formed in 1999 to certify interoperability between IEEE 802.11 devices from different manufacturers through rigorous testing. Since 2000, shipments of Wi-Fi certified integrated circuits (IC) reached 200 million per year in 2006 (ABIresearch, 2007). Shipments are expected to exceed a billion units per year by 2012 (ABIresearch, 2007), as illustrated in Figure 1.1.

Such large and sustained growth is due to the benefits WLANs offer over wired networking. In existing homes or enterprises, deploying cables for network access may involve tearing up walls, floors, or ceilings, which is both inconvenient and costly. In contrast, providing wireless network connectivity in these environments is often as simple as installing a single wireless access point. Perhaps more importantly though, the proliferation of laptops and handheld devices has meant that people desire connectivity wherever they are located, not just where the network connection is located. Network connectivity in a conference room or while seated on the sofa in the living room are just two examples of the flexibility afforded by WLANs.

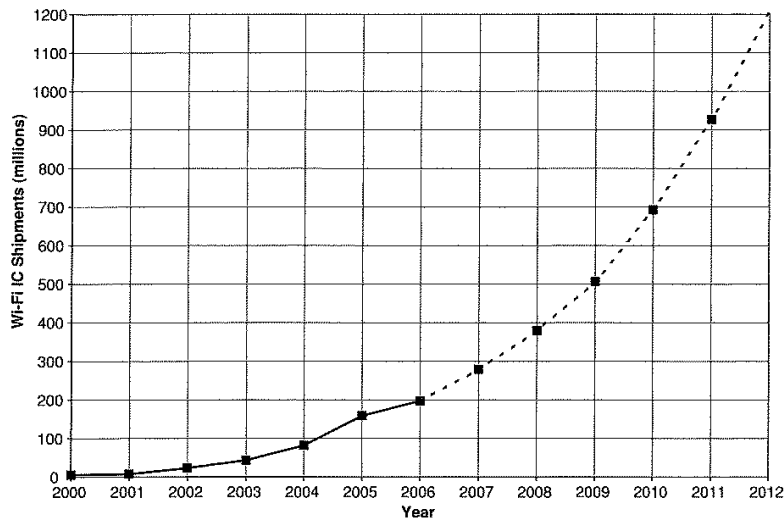


Figure 1.1 Wi-Fi IC shipments. Source: ABIresearch (2007).

Building on the convenience of mobility from the cellular world, WLANs are now enabling Internet access at little or no cost in public wireless networks. In 2005, Google offered to deploy a free Wi-Fi service covering San Francisco at no cost to the city. There has also been a proliferation of small scale deployments providing Internet access in coffee shops, airports, hotels, etc., which have come to be known as hotspots. Additionally, when these networks are used in conjunction with virtual private network (VPN) technology, employees can securely access corporate networks from almost anywhere.

The vast majority of WLAN products and systems today are based on the 802.11b, 802.11g, and 802.11a standard amendments, which provide throughput enhancements over the original 802.11 PHYs. Progress in WLAN technology continues with the development of 802.11n. Increased data rates are achieved with the multiple-input multiple-output (MIMO) concept, with its origins by Foschini (1996) at Bell Labs. In 2004, Atheros demonstrated that 40 MHz devices could be produced at almost the same cost as 20 MHz devices. During a similar time frame, the FCC and ETSI adopted new regulations in the 5 GHz band that added an additional 400 MHz of unlicensed spectrum for use by commercial WLANs. These events paved the way for the broad acceptance of 40 MHz operating modes in 802.11n. When spectrum is free, increasing the channel bandwidth is the most cost effective way to increase the data rate.

Typically product development lags standardization efforts and products are released after the publication of the standard. An interesting event occurred in 2003 when Broadcom released a chipset based on a draft version of the 802.11g amendment, prior to final

publication. This set a precedent for the flurry of “pre-n” or “draft-n” products in 2005 and 2006, as industry players rushed to be first to market. Most of these products were either proprietary implementations of MIMO, or based on draft 1.0 of 802.11n, and thus unlikely to be compliant with the final standard.

Through early 2007, major improvements and clarifications were made to the 802.11n draft resulting in IEEE 802.11n draft 2.0. To continue the market momentum and forestall interoperability issues, the IEEE took the unusual step of releasing 802.11n D2.0 to the public while work continued toward the final standard. This allowed the WFA to begin interoperability testing and certification of devices based on a subset of the 802.11n D2.0 features in May 2007. WFA certified 802.11n D2.0 products provide consumers the assurance of interoperability between manufacturers that was not guaranteed by previous “pre-n” or “draft-n” products. These were major steps in speeding up the standardization and certification process of new technology.

1.1 History of IEEE 802.11

The IEEE 802.11 working group began development of a common medium access control (MAC) layer for multiple physical layers (PHY) to standardize wireless local area networking. As a member of the IEEE 802 family of local area networking (LAN) and metropolitan area networking (MAN) standards, 802.11 interfaces with 802.1 architecture, management, and interworking, and 802.2 logical link control (LLC). The combination of 802.2 LLC and 802.11 MAC and PHY make up the data link and physical layers of the Open Systems Interconnection (OSI) reference model, as described in Table 1.1.

Table 1.1 OSI reference model (Zimmerman, 1980; Teare, 1999)

OSI Reference Model layers	Description	Examples	Layer categories
Application	Interacts with software applications that implement a communicating component	Telnet, FTP, SMTP	Application
Presentation	Coding and conversion functions that are applied to application layer data	QuickTime, MPEG, GIF, JPEG, TIFF	
Session	Establishes, manages, and terminates communication sessions	ZIP, AppleTalk, SCP, DECnet Phase IV	
Transport	Accepts data from the session layer and segments the data for transport across the network	TCP, UDP	Data transport
Network	Defines the network address	IP, IPv6	
Data link	Transit of data across a physical network link	802.2 LLC	
Physical	Electrical, mechanical, procedural, and functional specifications	802.11 PHY	

The initial version of the 802.11 standard was completed in 1997. Influenced by the huge market success of Ethernet (standardized as IEEE 802.3), the 802.11 MAC adopted the same simple distributed access protocol, carrier sense multiple access (CSMA). With CSMA, a station wishing to transmit first listens to the medium for a predetermined period. If the medium is sensed to be "idle" during this period then the station is permitted to transmit. If the medium is sensed to be "busy," the station has to defer its transmission. The original (shared medium) Ethernet used a variation called CSMA/CD or carrier sense multiple access with collision detection. After determining that the medium is "idle" and transmitting, the station is able to receive its own transmission and detect collisions. If a collision is detected, the two colliding stations backoff for a random period before transmitting again. The random backoff period reduces the probability of a second collision.

With wireless it is not possible to detect a collision with one's own transmission directly in this way; thus 802.11 uses a variation called CSMA/CA or carrier sense multiple access with collision avoidance. With CSMA/CA, if the station detects that the medium is busy, it defers its transmission for a random period following the medium going "idle" again. This approach of always backing off for a random period following another station's transmission improves performance since the penalty for a collision is much higher on a wireless LAN than on a wired LAN. On a wired LAN collisions are detected electrically and thus almost immediately, while on wireless LAN collisions are inferred through the lack of an acknowledgement or other response from the remote station once the complete frame has been transmitted.

There is no doubt that the simplicity of this distributed access protocol, which enables consistent implementation across all nodes, significantly contributed to Ethernet's rapid adoption as the industry LAN standard. Likewise, the adoption by the industry of 802.11 as the wireless LAN standard is in no small part due to the simplicity of this access protocol, its similarity to Ethernet, and again the consistent implementation across all nodes that has allowed 802.11 to beat out the more complex, centrally coordinated access protocols of competing WLAN technologies such as HyperLAN.

The original (1997) 802.11 standard included three PHYs: infrared (IR), 2.4 GHz frequency hopped spread spectrum (FHSS), and 2.4 GHz direct sequence spread spectrum (DSSS). This was followed by two standard amendments in 1999: 802.11b built upon DSSS to increase the data rate in 2.4 GHz and 802.11a to create a new PHY in 5 GHz. 802.11b enhanced DSSS with complementary code keying (CCK), increasing the data rate to 11 Mbps. With higher data rates, IEEE 802.11b devices achieved significant market success, and markets for IR and FHSS PHYs did not materialize.

The development of 802.11a introduced orthogonal frequency division multiplexing (OFDM) to 802.11. Even though 802.11a introduced data rates of up to 54 Mbps, it is confined to the 5 GHz band and, as a result, adoption has been slow. New devices wishing to take advantage of the higher rates provided by 802.11a but retain backward compatibility with the huge installed base of 802.11b devices would need to implement two radios, one to operate using 802.11b in the 2.4 GHz band and one to operate using 802.11a in the 5 GHz band. Furthermore, international frequency regulations in the 2.4 GHz band uniformly allowed commercial use, whereas in 1999 and 2000 the non-military use of the 5 GHz band was limited to select channels in the United States.

Table 1.2 Overview of 802.11 PHYs

	802.11	802.11b	802.11a	802.11g	802.11n
PHY technology	DSSS	DSSS/CCK	OFDM	OFDM DSSS/CCK	SDM/OFDM
Data rates	1, 2 Mbps	5.5, 11 Mbps	6–54 Mbps	1–54 Mbps	6–600 Mbps
Frequency band	2.4 GHz	2.4 GHz	5 GHz	2.4 GHz	2.4 and 5 GHz
Channel spacing	25 MHz	25 MHz	20 MHz	25 MHz	20 and 40 MHz

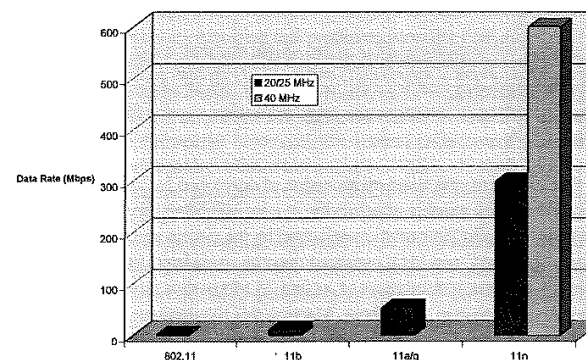


Figure 1.2 Increase in 802.11 PHY data rate.

In 2001, the FCC permitted the use of OFDM in the 2.4 GHz band. Subsequently, the 802.11 working group developed the 802.11g amendment, which incorporates the 802.11a OFDM PHY in the 2.4 GHz band, and adopted it as part of the standard in 2003. In addition, backward compatibility and interoperability is maintained between 802.11g and the older 802.11b devices. This allows for new 802.11g client cards to work in existing 802.11b hotspots, or older 802.11b embedded client devices to connect with a new 802.11g access point (AP). Because of this and new data rates of up to 54 Mbps, 802.11g has experienced large market success. A summary of the high level features of each PHY is given in Table 1.2.

With the adoption of each new PHY, 802.11 has experienced a five-fold increase in data rate. This rate of increase continues with 802.11n with a data rate of 300 Mbps in 20 MHz and 600 Mbps in 40 MHz. The exponential rate of increase in data rate is illustrated in Figure 1.2.

1.2 History of high throughput and 802.11n

1.2.1 The High Throughput Study Group

Interest in a high data rate extension to 802.11a began with a presentation to the Wireless Next Generation Standing Committee (WNG SC) of IEEE 802.11 in January 2002. Market drivers were outlined, such as increasing data rates of wired Ethernet, more data

rate intensive applications, non-standard 100+ Mbps products entering the market, and the need for higher capacity WLAN networks (Jones, 2002). The presentation mentioned techniques such as spatial multiplexing and doubling the bandwidth as potential approaches to study for increasing data rate.

After many additional presentations, the High Throughput Study Group (HTSG) was formed with its first meeting in September 2002. The primary objective of HTSG was to complete two documents necessary for the creation of the High Throughput Task Group (TGn). These are the project authorization request (PAR) form and five criteria form. The PAR defined the scope and purpose of the task group as follows:

The scope of this project is to define an amendment that shall define standardized modifications to both the 802.11 physical layers (PHY) and the 802.11 medium access control layer (MAC) so that modes of operation can be enabled that are capable of much higher throughputs, with a maximum throughput of at least 100 Mbps, as measured at the MAC data service access point (SAP). IEEE (2006)

By this statement, the standard amendment developed by TGn must contain modes of operation that are capable of achieving at least 100 Mbps throughput. Throughput is the measure of "useful" information delivered by the system and by using throughput as the metric, both MAC and PHY overhead must be considered. 802.11a/g systems typically achieve a maximum throughput of around 25 Mbps; thus this statement required at least a four fold increase in throughput. Meeting this requirement would in essence mandate PHY data rates well in excess of 100 Mbps as well as significant enhancements to MAC efficiency.

Additional explanatory notes were included with the PAR outlining many evaluation metrics. These include throughput at the MAC SAP, range, aggregate network capacity, power consumption, spectral flexibility, cost complexity flexibility, backward compatibility, and coexistence (IEEE, 2006).

The five criteria form requires that the study group demonstrate the necessity of creating an amendment to the standard. The criteria include (1) broad market potential, (2) compatibility with existing IEEE 802.1 architecture, (3) distinct identity from other IEEE 802 standards, (4) technical feasibility, and (5) economic feasibility (Rosdahl, 2003). The goal is to create a standard amendment which results in marketable products, but that will also be differentiated from other potentially similar products.

In addition to completing the PAR and five criteria forms, HTSG also began development of new multipath fading MIMO channel models (Erceg *et al.*, 2004) and usage models (Stephens *et al.*, 2004). The channel models and usage models were used to create a common framework for simulations by different participants in the standard development process.

1.2.2 Formation of the High Throughput Task Group (TGn)

The PAR was accepted and approved by the 802 working group, creating Task Group n (TGn) with the first meeting of the task group held in September 2003. The standard amendment developed by the task group would be proposal driven, meaning that

members of the task group would make partial or complete technical proposals, with the complete proposals proceeding through a down-selection process culminating in a single proposal upon which the standard amendment would be based. Partial proposals would be informative and could be incorporated in a complete proposal along the way. To that end, the task group began development of the functional requirements (Stephens, 2005) and comparison criteria (Stephens, 2004) documents. These two documents would provide, respectively, the technical requirements complete proposals must meet and criteria by which complete proposals would be compared.

The task group began with nine functional requirements. One of the functional requirements was a catch-all, requiring that proposals meet the PAR and five criteria. A second requirement was a reiteration of the PAR requirement to achieve 100 Mbps throughput at the top of the MAC. Furthermore, since it was expected that not all regulatory domains would allow a single device to use multiple 20 MHz channels (an easy way to achieve the throughput objective), the second requirement added a restriction that 100 Mbps throughput be achieved in a single 20 MHz channel. To enforce efficient use of spectrum, another requirement was added for a mode of operation with a spectral efficiency of at least 3 bps/Hz.

Four functional requirements addressed operational bands and backward compatibility. One of these requirements was that the protocol should support operation in the 5 GHz band due to the large availability of spectrum there. Another requirement was that at least some modes of operation be backward compatible with 802.11a systems. Noteworthy was the fact that there was no requirement to support operation in the 2.4 GHz band. However, if a proposal did support 2.4 GHz band operation, it was required that there be modes of operation that were backward compatible with 802.11g systems. In this context, some flexibility was given, allowing an 802.11n AP to be configured to accept or reject associations from legacy stations.

The 802.11e amendment to the standard, nearing completion at the time, added many features for improving the quality of service (QoS) in 802.11 systems. Many of the perceived applications for 802.11n involved real time voice and video which necessitate QoS. Therefore a functional requirement was included which mandated that a proposal allow for the implementation of 802.11e features within an 802.11n station.

The comparison criteria in Stephens (2004) outlined metrics and required disclosure of results which would allow for comparison between proposals under the same simulation setup and assumptions. The comparison criteria incorporated the simulation scenarios and usage models defined in Stephens *et al.* (2004). During the development of the comparison criteria, the task group realized that members of the task group did not always share the same definitions for common terms. Therefore definitions for goodput, backward compatibility, and signal-to-noise ratio (SNR) were provided. The comparison criteria covered four main categories: marketability, backward compatibility and coexistence with legacy devices, MAC related criteria, and PHY related criteria.

Under marketability, the proposal must provide goodput results for residential, enterprise, and hotspot simulation scenarios. Goodput is defined by totaling the number of bits in the MAC service data units (MSDU) indicated at the MAC service access point

(SAP), and dividing by the simulation duration (Stephens, 2004). Two optional criteria included describing the PHY and MAC complexity. The PHY complexity was to be given relative to 802.11a.

To ensure backward compatibility and coexistence with legacy devices, a proposal was required to provide a summary of the means used to achieve backward compatibility with 802.11a and, if operating in 2.4 GHz, 802.11g. Simulation results demonstrating interoperability were also required. The goodput of a legacy device in an 802.11n network and the impact of a legacy device on the goodput of 802.11n devices were also to be reported.

The MAC related criteria included performance measurements and changes that were made to the MAC. In the residential, enterprise, and hotspot simulation scenarios a number of different metrics were to be captured and reported. These included the ability to support the service requirements of various applications, including QoS requirements. Measurements of aggregate goodput of the entire simulation scenario were required to indicate network capacity. MAC efficiency was to be provided, which is defined as the aggregate goodput divided by the average PHY data rate. To ensure reasonable range for the new modes of operation, throughput versus range curves were also to be provided.

The PHY related criteria included PHY rates and preambles, channelization, spectral efficiency, PHY performance, and PHY changes. In addition, the comparison criteria also defined PHY impairments to be used in combination with channel models for PHY simulations. Each proposal was required to generate simulation results for both additive white Gaussian noise (AWGN) and non-AWGN channels. Furthermore, simulation conditions to analyze the impact on packet error rate (PER) of carrier frequency offset and symbol clock offset were also defined.

1.2.3 Call for proposals

The TGN call for proposals was issued on May 17, 2004, with the first proposals presented in September 2004. Over the course of the process two main proposal teams emerged, TGN Sync and WWiSE (world wide spectral efficiency). The TGN Sync proposal team was founded by Intel, Cisco, Agere, and Sony with the objective of covering the broad range of markets these companies were involved in, including the personal computer (PC), enterprise, and consumer electronics markets. The WWiSE proposal team was formed by Broadcom, Conexant, and Texas Instruments. These semiconductor companies were interested in a simple upgrade to 802.11a for fast time to market. Many other companies were involved in the proposal process and most ended up joining one of these two proposal teams.

The key features of all the proposals were similar, including spatial division multiplexing and 40 MHz channels for increased data rate, and frame aggregation for improved MAC efficiency. The proposals differed in scope (TGN Sync proposed numerous minor improvements to the MAC while WWiSE proposed limiting changes) and support for advanced features such as transmit beamforming (initially absent from the WWiSE proposal).

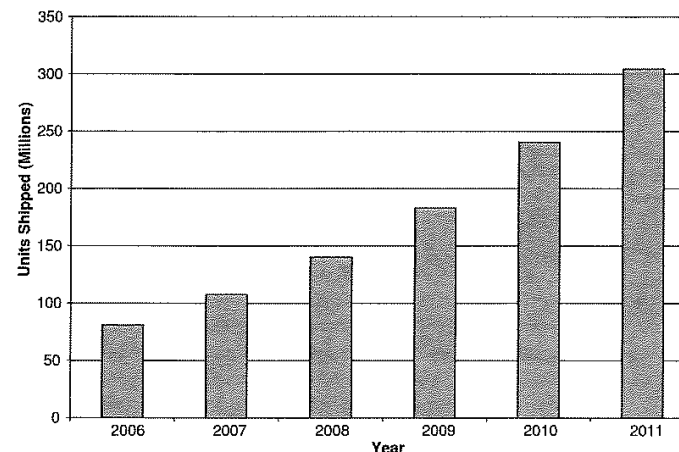


Figure 1.3 Worldwide converged mobile device shipments. Source: IDC (2007).

A series of proposal down-selection and confirmation votes took place between September 2004 and May 2005. During that time, mergers between proposals and enhancements to proposals took place. The TGN Sync proposal won the final down-selection vote between it and WWiSE, but failed the confirmation vote in May 2005.

1.2.4 Handheld devices

During this period interest arose in a new emerging market of converged Wi-Fi and mobile handsets. The shipment of dual mode Wi-Fi/cellular handsets had grown significantly from 2005 to 2006. Some participants in the proposal process believed that handsets would be the dominant Wi-Fi platform within a few years (de Courville *et al.*, 2005). A projected world wide growth of converged mobile devices was given in IDC (2007) and is illustrated in Figure 1.3.

A contentious issue for handheld proponents was the high throughput requirement for 100 Mbps throughput. This, in essence, would force all 802.11n devices to have multiple antennas. This is a difficult requirement for converged mobile devices, since they already contain radios and antennas for cellular 2G, 3G, Bluetooth, and in some cases GPS. Concern was raised that mandating 802.11n devices to have multiple antennas would force handset manufacturers to continue to incorporate single antenna 802.11a/g into handsets and not upgrade to 802.11n. Not only does this diminish the capabilities of the handset device, it burdens all future 802.11n deployments with continued coexistence with 802.11a/g embedded in these new handset devices.

For this reason an ad hoc group was formed to create functional requirements supporting single antenna devices. Two new requirements were added to the functional requirements document in July 2005. The first requirement mandated that a proposal define single antenna modes of operation supporting at least 50 Mbps throughput in a 20 MHz channel. The second requirement dictated that an 802.11n AP or station interoperate with client devices that comply with 802.11n requirements but incorporate only a single antenna. This requirement resulted in 802.11n making mandatory at least two antennas in an AP, but only one antenna in a non-AP device.

1.2.5 Merging of proposals

After the failed confirmation vote, a joint proposal effort was started within the task group to merge the two competing proposals. Due to entrenched positions and the large membership of the group, the joint proposal effort proceeded very slowly. As a result, Intel and Broadcom formed the Enhanced Wireless Consortium (EWC) in October 2005 to produce a specification outside the IEEE that would bring products to market faster. With much of the task group membership ultimately joining the EWC, this effort had the effect of breaking the deadlock within the IEEE, and the EWC specification, which was essentially a merger of the TGn Sync and WWiSE proposals, was adopted as the joint proposal and submitted for confirmation to TGn where it was unanimously passed in January 2006.

1.2.6 802.11n amendment drafts

The joint proposal was converted to a draft 802.11 standard amendment for higher throughput (TGn Draft 1.0), and entered letter ballot. In letter ballot, IEEE 802.11 working group members (not just task group members) vote to either adopt the draft as is or reject it with comments detailing changes needed. The draft requires at least a 75% affirmative vote within the IEEE 802.11 working group in order to proceed to sponsor ballot where it is considered for adoption by the broader IEEE standards association. TGn Draft 1.0 entered letter ballot in March 2006 and, not unusually, failed to achieve the 75% threshold for adoption. Comment resolution began May 2006 on the roughly 6000 unique technical and editorial comments submitted along with the votes.

With resolution of the TGn Draft 1.0 comments, TGn Draft 2.0 went out for letter ballot vote in February 2007 and this time passed with 83% of the votes. However, there were still 3000 unique technical and editorial comments accompanying the letter ballot votes. It is typical for the task group to continue comment resolution until a minimum number of negative votes are received; thus comment resolution for TGn Draft 2.0 continued between March 2007 and September 2007, resulting in TGn Draft 3.0. Since TGn Draft 2.0 passed, TGn Draft 3.0 and possible later drafts only require a recirculation ballot in which comments may only address clauses that changed between the drafts.

At the time this book went to press, the standard amendment was in recirculation ballot and would continue there until a minimum number of negative votes and comments were received. It will then proceed to sponsor ballot. Whereas letter ballot includes only

voting members in IEEE 802.11, the sponsor ballot pool may include members from all of the IEEE 802 standard association, providing a broader review of the draft.

The IEEE 802.11n standard amendment is expected to be completed with final IEEE 802 working group and executive committee approval in March 2009. The IEEE standards board is expected to approve the amendment in June 2009. Publication of the amendment would occur shortly after.¹

The submissions contributed by the task group participants in the development of the standard are publicly available on the World Wide Web.² The drafts of the standard are only available to voting members of 802.11, but, draft 2.0 of 802.11n was released to the public. Draft 3.0 is currently available for purchase from the IEEE.³ Once approved, the final standard amendment would be available there as well.

Of great value to someone investigating the 802.11n PHY is the transmit waveform generator developed by Metalink. The description of the generator, developed in MATLAB®, is given in Anholt and Livshitz (2006). The actual source code is publicly available and is included in Anholt and Livshitz (2007). Most, if not all, transmit waveform features are supported by the generator.

1.3

Environments and applications for 802.11n

The basic service set (BSS) is the basic building block of an 802.11 LAN. Stations that remain within a certain coverage area and form some sort of association form a BSS. The most basic form of association is where stations communicate directly with one another in an ad-hoc network, referred to as an independent BSS or IBSS. This is illustrated as BSS 1 in Figure 1.4.

More typically though, stations associate with a central station dedicated to managing the BSS and referred to as an access point (AP). A BSS built around an AP is called an infrastructure BSS and is illustrated by BSS 2 and BSS 3 in Figure 1.4. Infrastructure BSSs may be interconnected via their APs through a distribution system (DS).

The BSSs interconnected by a DS form an extended service set (ESS). A key concept of the ESS is that stations within the ESS can address each other directly at the MAC layer. The ESS, being an 802.11 concept, encompasses only the 802.11 devices and does not dictate the nature of the DS. In practice, however, the DS is typically an Ethernet (802.3) LAN and the AP functions as an Ethernet bridge. As such, stations in a BSS can also directly address stations on the LAN at the MAC layer.

In the development of 802.11n, three primary environments were considered for studying system performance and capabilities: residential, enterprise, and hotspot. Within each of these environments a different mix of existing and new applications was envisioned. Use cases were defined (Stephens *et al.*, 2004) that describe how an end user uses an

¹ The reader is referred to http://grouper.ieee.org/groups/802/11/Reports/802.11_Timelines.htm for the latest update on the timeline of 802.11n.

² <http://grouper.ieee.org/groups/802/11/>

³ <http://standards.ieee.org/getieee802/>

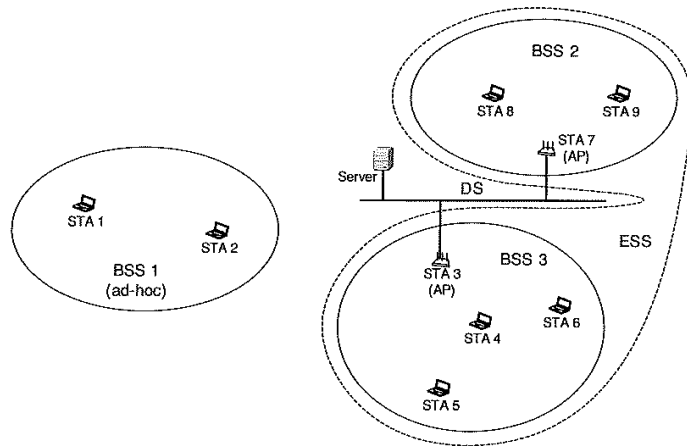


Figure 1.4 BSS, DS, and ESS concepts.

application in a specific WLAN environment. Examples include watching television remotely from the cable or set-top box within the home, or talking on the telephone remotely from one's desk at work. Additionally, usage models were developed for each environment which combined multiple use cases and applications.

Finally, a simulation scenario was created for each usage model and environment. These simulation scenarios were used to stress MAC capability when comparing proposals. Each simulation scenario includes a channel model associated with the particular environment. In addition, the location of the AP and stations were defined, giving a spatial component to the usage model in terms of the distance between the AP and the stations. For each application, system parameters were defined such as packet size, maximum packet loss rate tolerated by the application, maximum delay, network-layer protocol running (e.g. UDP or TCP), and offered load.

The residential usage model consists of a single BSS, as illustrated in Figure 1.5. This model typically includes only one AP and many client stations. In a typical AP-station configuration, applications include Internet access and streaming audio and video. Furthermore, user experience with applications like intra-networking for local file transfer, backups, and printing is enhanced with higher data rates. New applications such as voice over IP (VoIP) and video phones were also incorporated into the residential usage model.

The high throughput task group envisioned an AP that could also take the form of a wireless home media gateway. Such a device would distribute audio and video content throughout the home, such as DVD and standard and high definition TV. Other residential applications benefiting from higher wireless data rates include content download from

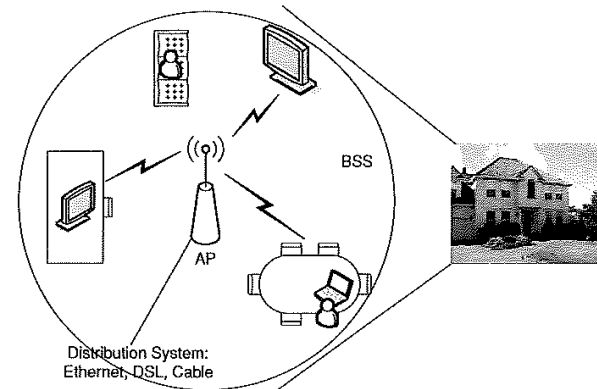


Figure 1.5 Residential usage model.

a video camera or photo camera. Interactive gaming has recently begun to incorporate wireless technology. Gamers benefit from the freedom of not being tethered by wires when the connections between the controller and the console, the console and the display, and the console to internet access are made wireless.

The usage model for an enterprise environment emphasizes network connectivity supported by multiple BSSs to cover larger buildings and floor plans, as illustrated in Figure 1.6. The BSSs are interconnected via the distribution system, typically Ethernet, creating an extended service set (ESS). As in a cellular deployment, each additional AP increases the total network coverage and capacity.

Networking applications such as file transfer and disk backup will benefit greatly from the higher data rates of 802.11n. Higher data rates will increase network capacity providing support for a larger number of clients. Higher throughputs will also enable new applications such as remote display via a wireless connection between a laptop and projector, simplifying presentations in conference rooms. Additionally, wireless video conferencing and VoIP may be supported (Stephens *et al.*, 2004).

The hotspot model envisions locations such as an airport lounge (illustrated in Figure 1.7), coffee shop, library, hotel, or convention center. Some municipalities have also blanketed downtown areas with Wi-Fi coverage. A hotspot could be located either indoors or outdoors and could cover a large open area. Therefore the propagation model could be substantially different from either residential or enterprise. In a hotspot, most traffic goes through the Internet and a session is typically limited to less than two hours (Stephens *et al.*, 2004). Applications include web browsing, Internet file transfer, and email. Also, new hotspot applications are envisioned such as the ability to watch a TV program or movie on a laptop or other display. This would involve the streaming of audio

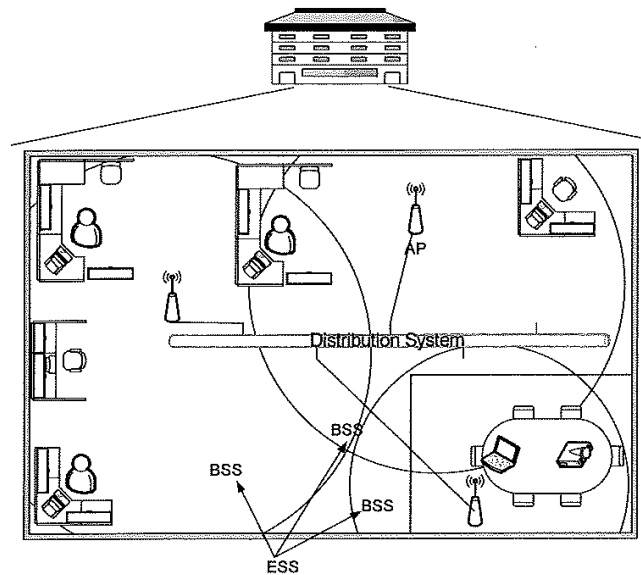


Figure 1.6 Enterprise usage model.

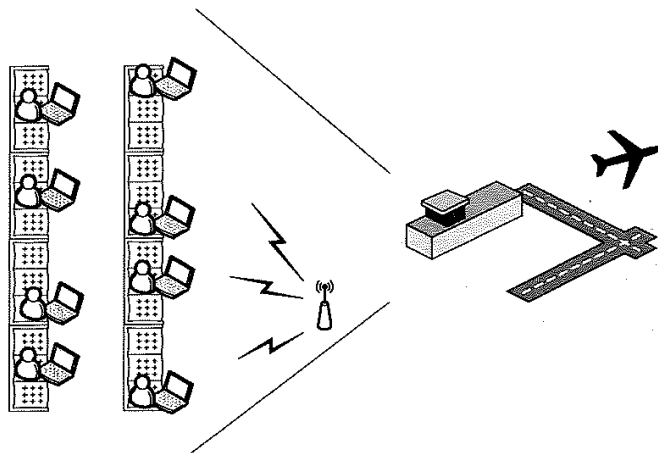


Figure 1.7 Hotspot usage model.

and video content over the Internet or the redistribution of standard or high definition TV signals.

1.4 Major features of 802.11n

PHY data rates in 802.11n are significantly improved over 802.11a and 802.11g primarily through the use of spatial multiplexing using MIMO and 40 MHz operation. To take advantage of the much higher data rates provided by these techniques, MAC efficiency is also improved through the use of frame aggregation and enhancements to the block acknowledgment protocol. These features together provide the bulk of the throughput enhancement over that achievable with 802.11a and 802.11g.

Robustness is improved inherently through the increased spatial diversity provided by the use of multiple antennas. Space-time block coding (STBC) as an option in the PHY further improves robustness, as does fast link adaptation, a mechanism for rapidly tracking changing channel conditions. More robust channel codes are adopted in the form of low density parity check (LDPC) codes. The standard amendment also introduces transmit beamforming, with both PHY and MAC enhancements to further improve robustness.

A number of other enhancements provide further gains. In the PHY, these include a shorter guard interval, which may be used under certain channel conditions. The PHY also includes a Greenfield preamble, which is shorter than the mandatory mixed format preamble. However, unlike the mixed format, it is not backward compatible with existing 802.11a and 802.11g devices without MAC protection. In the MAC, the reverse direction protocol provides a performance improvement for certain traffic patterns, by allowing a station to sublease the otherwise unused portion of its allocated transmit opportunity to its remote peer and thus reducing overall channel access overhead. A reduced interframe space (RIFS) used when transmitting a burst of frames reduces overhead in comparison to the existing short interframe space (SIFS).

An overview of the mandatory and optional features of the 802.11n PHY is given in Figure 1.8. At the time this book went to press, two generations of so called pre-n or draft 2.0 products have been released. The first generation of products typically operate in the 2.4 GHz band only, with up to two spatial streams and 40 MHz channel width. In this book the term spatial streams is used to refer to one or more independent data streams transmitted from the antennas. A device requires at least as many antennas as spatial streams. When using the short guard interval, these initial products are able to achieve a PHY data rate of 300 Mbps. With second generation products, we begin to see dual-band 2.4 GHz and 5 GHz products. These products also achieve 300 Mbps, but several incorporate an extra receive antenna chain for additional receive diversity. Some products also support the Greenfield preamble format. We expect that third generation devices will add another transmit antenna chain to support three spatial streams and

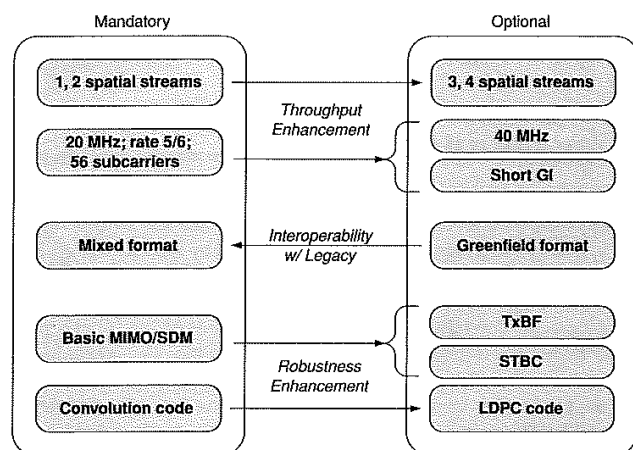


Figure 1.8 Mandatory and optional 802.11n PHY features.

450 Mbps. For robustness, these devices may begin employing STBC and transmit beamforming (TxBF).

An overview of the features added to the MAC in 802.11n is given in Figure 1.9. In addition to the throughput and robustness enhancing features already mentioned, the MAC is extended in a number of other areas.

The numerous optional features in 802.11n mean that extensive signaling of device capability is required to ensure coexistence and interoperability. For example, whether a device supports certain PHY features such as the Greenfield format preamble or MAC features such as the ability to participate in a reverse direction protocol exchange.

The existence of 40 MHz operation also creates a number of coexistence issues. The AP needs to manage the 40 MHz BSS so that 40 MHz and 20 MHz devices, both legacy and high throughput, are able to associate with the BSS and operate. Because 40 MHz operation uses two 20 MHz channels, mechanisms are needed to mitigate the effect this might have on neighboring 20 MHz BSSs operating independently on those two channels. Coexistence is primarily achieved through careful channel selection, i.e. choosing a pair of channels that have little or no active neighborhood traffic. To this end, the amendment adds scanning requirements to detect the presence of active neighborhood BSSs as well as the ability to actively move the BSS to another pair of channels should a neighboring 20 MHz BSS become active. If neighboring BSSs cannot be avoided then a fallback technique called phased coexistence operation (PCO) may be used. This allows the BSS to alternate between 20 MHz and 40 MHz phases of operation, with the 40 MHz phase entered after a frame exchange on the two 20 MHz channels has silenced devices operating there.

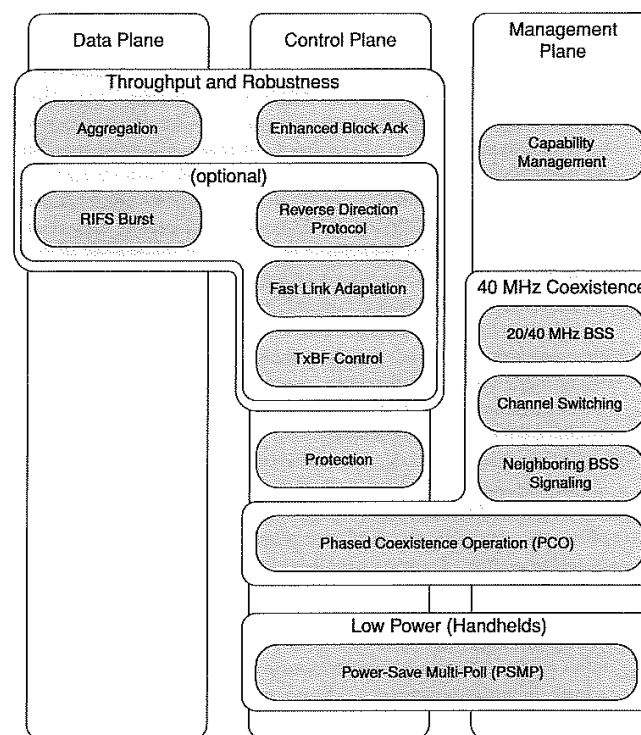


Figure 1.9 Summary of 802.11n MAC enhancements.

Finally, in recognition of the growing importance of handheld devices, a channel access scheduling technique called power-save multi-poll (PSMP) has been added to efficiently support a large number of stations.

1.5

Outline of chapters

The book is divided into three parts. The first part, Chapters 2–6, covers the PHY and provides a comprehensive review of all mandatory and optional PHY features in 802.11n. The second part, Chapters 7–11, covers the MAC, an overview of existing MAC features, and a detailed review of the new features introduced in 802.11n. Transmit beamforming is presented in the last part, Chapter 12, with a description of both PHY and MAC aspects.

Chapter 2 gives a brief overview of orthogonal frequency division multiplexing (OFDM). Chapter 3 begins with a description of multiple-input multiple-output (MIMO) basics and spatial division multiplexing (SDM). This is followed by a discussion of the MIMO environment and the 802.11n MIMO multipath fading channel and propagation models. The chapter concludes with an explanation of linear receiver design and highlights of maximum likelihood estimation. Included in the discussion of MIMO and receiver design are capacity based performance curves.

Chapter 4 details the design of the mixed format (MF) preamble used for interoperability with legacy 802.11a/g OFDM devices. The chapter begins with a review of 802.11a preamble design. Included in the review are illustrations of the waveform. A description of the 802.11a/g packet encoding process and receive procedure is presented, which includes a receiver block diagram. This leads into a discussion of the legacy part of the MF preamble. Next, the high throughput (HT) portion of the preamble is described. Following this, the encoding of the Data field is presented. The chapter ends with a discussion of the receive procedure and a block diagram for basic modes of operation. Tables of parameters of the modulation and coding scheme (MCS) for basic modes of operation are given in an appendix.

Chapter 5 outlines all the PHY techniques employed in the 802.11n specification to increase the data rate. The first section of the chapter details the new 40 MHz channel and waveform. This includes a plot illustrating throughput as a function of range. This is followed by a brief discussion of the extra subcarriers which were added to the 802.11n 20 MHz waveform. The next part of the chapter gives the MCS definitions. This includes several waterfall curves illustrating the packet error rate (PER) versus SNR performance in additive white Gaussian noise (AWGN) and in the 802.11n MIMO multipath fading channel models. A description of the shorter Greenfield (GF) preamble is also provided in this chapter. This also includes discussion on the debate of how much the GF preamble actually improves performance. The last topic covered in this chapter is on the short guard interval (GI).

Chapter 6 covers the subject of improving the robustness of the system. Four techniques are described. The first method is receive diversity, where PER versus SNR waterfall curves are provided along with throughput curves to demonstrate the gain achieved from receive diversity in a MIMO system. The next technique is a straightforward one involving spatial expansion (SE), which provides a small amount of transmit diversity gain. Waterfall curves are provided for SE as well. This is followed by a detailed description of space-time block coding (STBC). Transmit antenna configurations are presented, along with an approach for implementing a receiver and equalizer. Again performance curves are presented that illustrate which system configurations benefit the most from STBC. In the last part of the chapter, low density parity check (LDPC) codes are discussed. The specific characteristics of the LDPC encoding process in the 802.11n standard amendment are detailed. Waterfall curves for LDPC are provided to compare performance with the mandatory binary convolutional code.

The MAC section begins in Chapter 7 with a functional description of the 802.11 MAC as background for the remaining chapters. This chapter covers the basic contention-based

access protocol including the 802.11e quality of service (QoS) extensions, channel access timing, the concept of a transmit opportunity, and the basic acknowledgement and block acknowledgement protocols.

Chapter 8 describes why changes are necessary in the MAC to improve throughput and then details the two key throughput enhancing features: aggregation and enhancements to the block acknowledgment protocol.

Beyond the basic contention-based access protocol, the 802.11 MAC includes additional channel access mechanisms. Chapter 9 provides an overview of these mechanisms, including the point coordination function (PCF) from the original 802.11 specification and the hybrid coordinated channel access (HCCA) function from the 802.11e amendment. The chapter then provides details on the power-save multi-poll (PSMP) channel access technique and the reverse direction protocol, both of which are new in 802.11n.

Coexistence and interoperability is a critical issue with 802.11n and Chapter 10 provides details on this broad topic. The chapter covers capability signaling and BSS control. The chapter then covers 40 MHz operation, managing 40 MHz BSS operation, and maintaining interoperability with legacy 20 MHz devices. The critical topic of 40 MHz coexistence with neighboring 20 MHz BSSs is also discussed. Finally, the chapter covers protection mechanisms.

To round out the MAC section, Chapter 11 provides details on MAC frame formats. This chapter is intended as a reference for the discussions in the other chapters.

The final part of the book deals with the complex topics of transmit beamforming and fast link adaptation. Chapter 12 provides details on both the PHY and MAC aspects of this topic.

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Part I

Physical layer

2 Orthogonal frequency division multiplexing

2.1 Background

The 802.11n physical layer builds upon the 802.11a orthogonal frequency division multiplexing (OFDM) structure. OFDM is well suited to wideband systems in frequency selective fading environments. Only a few subcarriers are impacted by a deep fade or narrow band interference, which can be protected by forward error correction. Important in a high data rate system, OFDM is tolerant of time synchronization errors. In addition, OFDM is bandwidth efficient, since a nearly square power spectrum can be created with narrow subcarriers with each subcarrier supporting a constellation with many bits per symbol.

With frequency division multiplexing, signals are transmitted simultaneously on different subcarriers. Figure 2.1 illustrates an example with four subcarriers. The top two graphs in the figure depict the time domain waveform of each subcarrier individually for the real and imaginary part of each subcarrier. The bottom graph gives the composite time domain waveform.

Figure 2.2 illustrates the power spectrum of each subcarrier individually. If the subcarriers are separated in frequency (Δ_F) by the inverse of the symbol period (T), the nulls of adjacent subcarriers coincide with the peak of the main lobes of the subcarriers. With the construction of the waveform in this manner, the subcarriers are orthogonal.

A baseband OFDM waveform is constructed as an inverse Fourier transform of a set of coefficients X_k ,

$$r(t) = \frac{1}{N} \sum_k X_k \exp(j2\pi k \Delta_F t) \quad 0 \leq t < T \quad (2.1)$$

where Δ_F is the subcarrier frequency spacing and T is the inverse Fourier transform symbol period, with Δ_F equal to $1/T$, and N is the number of samples in the inverse Fourier transform. A set of modulated symbols is transmitted on subcarriers as the coefficients X_k . The inverse Fourier transform is commonly implemented by an IFFT.

In 802.11a, the fundamental sampling rate is 20 MHz, with a 64-point FFT/IFFT. The Fourier transform symbol period, T , is 3.2 μ s in duration and Δ_F is 312.5 kHz. Of the 64 subcarriers in 802.11a, there are 52 populated subcarriers. Numbering the subcarrier locations as $-32, -31, \dots, -1, 0, 1, \dots, 31$, the populated subcarriers are located from $-26, -25, \dots, -2, -1, 1, 2, \dots, 25, 26$. That is, the lowest six subcarriers, the DC subcarrier, and the highest five subcarriers are not used.

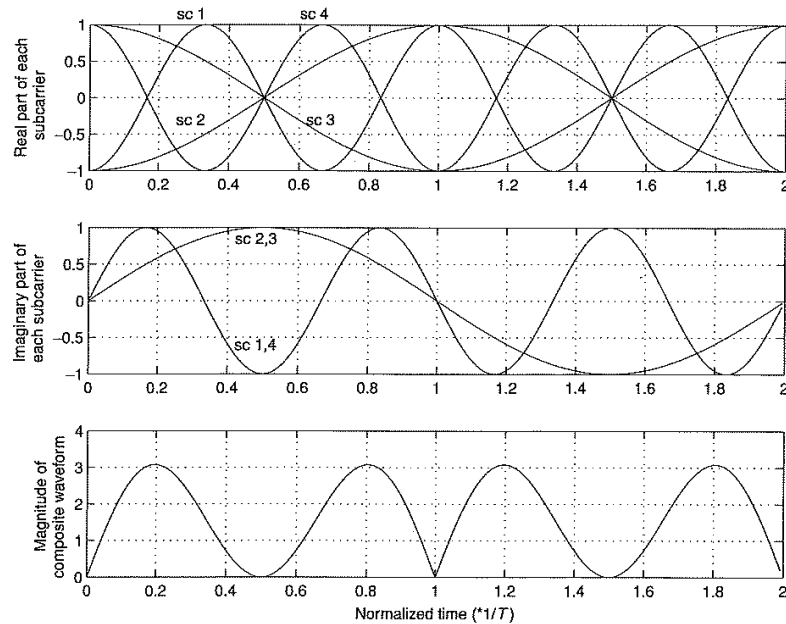


Figure 2.1 Frequency division multiplexing waveform.

Inter-symbol interference (ISI) between OFDM symbols that are adjacent in time degrades the orthogonality between the subcarriers and impairs performance. ISI may be caused by delay spread in the channel and filtering. To minimize the impact of ISI, a guard interval (GI) is added in between adjacent OFDM symbols. Therefore in 802.11a, the total symbol time is 4 μ s. Whereby, the first 0.8 μ s is the GI consisting of an exact copy of the last 0.8 μ s of the OFDM symbol (otherwise known as cyclic extension), followed by the 3.2 μ s OFDM symbol.

To extract the information from the received waveform, a Fourier transform is performed on the received signal, as given in Eq. (2.2):

$$X_k = \sum_n r(n) \exp(-j2\pi k \cdot n/N) \quad (2.2)$$

Note the Fourier transform is represented in the discrete time domain, whereas the inverse Fourier transform in Eq. (2.1) is represented in the continuous time domain. We have assumed that the received waveform has been sampled prior to the Fourier transform.

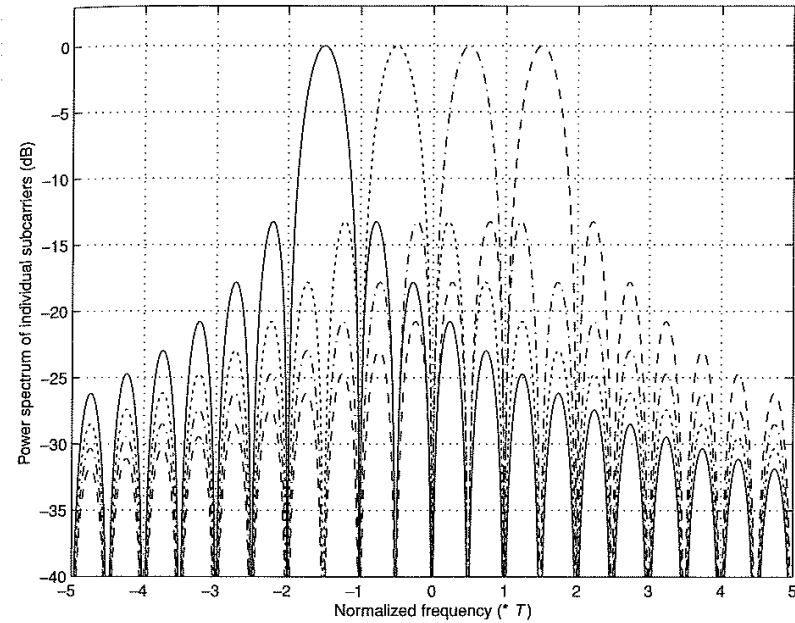


Figure 2.2 Power spectrum of the individual subcarriers of the OFDM waveform.

For an 802.11a OFDM waveform, a 4 μ s block of time domain samples is processed at a time. A 64-point FFT (assuming the data is sampled at a 20 MHz sampling rate) is performed on a 3.2 μ s subset of the time domain samples. The subset is selected to avoid ISI. Section 4.1 details the 802.11a waveform design and receive procedure.

Further general information on OFDM is provided in Halford (2001) and van Nee and Prasad (2000).

2.2

Comparison to single carrier modulation

With single carrier modulation, data bits are modulated and the pulses are transmitted sequentially in time. As described in the previous section, with OFDM, blocks of data bits are modulated onto subcarriers across the frequency band. An illustrative comparison between single carrier and OFDM is given in Figure 2.3.

An OFDM signal experiences large variation in magnitude when the individual subcarriers are combined into a single time domain waveform. The variation results in a high peak-to-average ratio, whereas the magnitude of the signal of a phase modulated

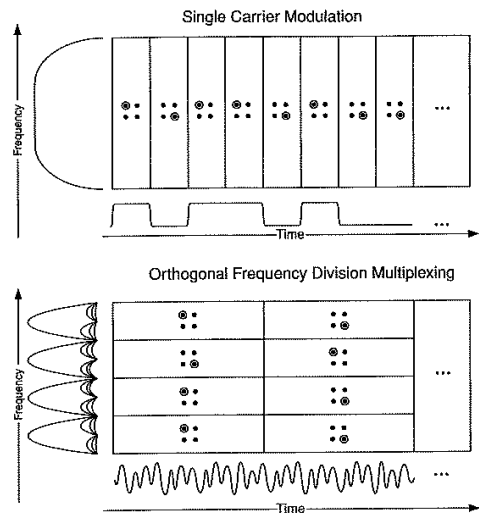


Figure 2.3 Comparison between single carrier modulation and orthogonal frequency division multiplexing.

single carrier waveform theoretically has a constant envelope. A properly designed single carrier system has little to no variation of the magnitude of the signal and a very small peak-to-average ratio.

This is one of the primary limitations of OFDM as compared to single carrier modulation. Due to the non-linear nature of a typical transmitter power amplifier, signals with a high peak-to-average ratio are distorted unless the amplifier is operated with enough power backoff to remain linear and avoid distortion. As a result, OFDM systems may have lower output transmit power or devices may require a larger power amplifier.

However, to achieve high data rates with limited bandwidth, designs of single carrier modulation systems inevitably utilize high order modulation with both amplitude and phase modulation. This results in the peak-to-average ratio of a single carrier waveform approaching that of an OFDM waveform. As such, typically there is only a small difference in the peak-to-average ratio of single carrier systems and OFDM.

The second limitation of OFDM when comparing to single carrier is higher sensitivity to carrier frequency offset and phase noise. In both OFDM and single carrier modulation, carrier frequency offset and phase noise causes unwanted phase variation in the modulated symbols. This requires phase tracking loops in either system to mitigate the degradation.

In addition, in OFDM, carrier frequency offset and phase noise cause the subcarriers to deviate from the $1/T$ spacing required for orthogonality, causing inter-carrier interference. In Pollet *et al.* (1995) an approximation for degradation due to carrier frequency offset (ϵ_f) is given for OFDM and single carrier, as follows:

$$D \approx \begin{cases} \frac{10}{\ln 10} \frac{1}{3} (\pi \epsilon_f T)^2 & \text{single carrier} \\ \frac{10}{\ln 10} \frac{1}{3} (\pi \epsilon_f T)^2 \frac{E_s}{N_0} & \text{OFDM} \end{cases} \quad (2.3)$$

Unlike single carrier, with OFDM the degradation is proportional to the signal-to-noise ratio E_s/N_0 . For phase noise, an approximation for degradation is given in Pollet *et al.* (1995) as a function of the oscillator linewidth (β), as follows:

$$D \approx \begin{cases} \frac{10}{\ln 10} \frac{1}{60} (4\pi\beta T) \frac{E_s}{N_0} & \text{single carrier} \\ \frac{10}{\ln 10} \frac{11}{60} (4\pi\beta T) \frac{E_s}{N_0} & \text{OFDM} \end{cases} \quad (2.4)$$

OFDM may be shown to be several orders of magnitude more sensitive to carrier frequency offset and phase noise than single carrier modulation. This is in part due to the symbol time (represented by T in Eqs. (2.3) and (2.4)) which for OFDM is much longer than for single carrier modulation, as illustrated in Figure 2.3.

The additional sensitivity of OFDM to carrier frequency offset and phase noise may appear to be problematic. However, the proliferation of low cost 802.11 OFDM devices demonstrates that the necessity of a more accurate oscillator does not significantly impact the overall cost of an OFDM device at these frequency bands.

As described in the previous section, OFDM is intrinsically resistant to frequency selective multipath fading. To combat frequency selectivity due to channels with long delay spread, a single carrier system requires an equalizer with a long tap-delay line. The implementation complexity based on number of multiplies, between the FFT in OFDM and the tap-delay line equalizer in single carrier, is shown in van Nee and Prasad (2000) to be ten times more complex for single carrier. In Clark (1998), a reduced complexity frequency domain equalizer for single carrier is proposed, as opposed to a tap-delay line equalizer. This technique requires both an FFT and IFFT as part of the receiver, which has twice the complexity of an OFDM receiver.

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3 MIMO/SDM basics

3.1 SISO (802.11a/g) background

A basic communication system is described as

$$y = \sqrt{\rho} \cdot h \cdot x + z \quad (3.1)$$

where x is the transmitted data with unity mean expected power, h is the channel fading coefficient, z is independent, complex additive white Gaussian noise (AWGN) with zero mean and unit variance, ρ is the average signal-to-noise ratio (SNR), and y is the received signal. Typically h is modeled as Rayleigh fading and is defined as complex, zero mean, unit variance Gaussian distribution: $\text{Normal}(0, 1/\sqrt{2}) + \sqrt{-1} \cdot \text{Normal}(0, 1/\sqrt{2})$. The notation $\text{Normal}(x, y)$ defines a Gaussian distributed random variable with a mean of x and a variance of y .

In this model, both the transmitter and receiver are configured with one antenna. This is termed single-input, single-output (SISO), with reference to a single input to the environment and single output from the environment, as illustrated in Figure 3.1.

The capacity for a general SISO system is given by the Shannon capacity formula in Eq. (3.2) (Foschini and Gans, 1998):

$$C(\text{bps/Hz}) = \log_2(1 + \rho \cdot |h|^2) \quad (3.2)$$

In the frequency domain (after the FFT), each subcarrier may be described by Eq. (3.1). Subsequently, the receiver extracts the information by equalizing the received signal, as follows:

$$\begin{aligned} \hat{x} &= (\sqrt{\rho} \cdot h)^{-1} \cdot y \\ &= x + (\sqrt{\rho} \cdot h)^{-1} \cdot z \end{aligned} \quad (3.3)$$

where $\hat{x}$ is the noisy estimate of the transmitted signal x .

3.2 MIMO basics

Multiple-input, multiple-output (MIMO) describes a system with a transmitter with multiple antennas transmitting through the propagation environment to a receiver with multiple receive antennas, as illustrated in Figure 3.2.

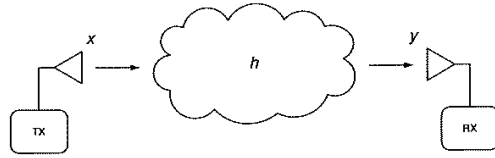


Figure 3.1 SISO system.



Figure 3.2 MIMO system.

In such a system the transmitter may use transmit beamforming (or an adaptive antenna array) to improve the quality of the link (Perahia and Pottie, 1996). An example of a system with a two antenna array transmitter and a single antenna receiver is described mathematically, and in vector notation, in Eq. (3.4):

$$\begin{aligned} y &= \sqrt{\rho} \cdot h_1 \cdot w_1 \cdot x + \sqrt{\rho} \cdot h_2 \cdot w_2 \cdot x + z \\ &= \sqrt{\rho} [h_1 \quad h_2] \begin{bmatrix} w_1 \\ w_2 \end{bmatrix} x + z \end{aligned} \quad (3.4)$$

This example is more accurately described as multiple-input, single-output. In this example, the received signal is the composite of signals transmitted by the two antennas, with the weight w_1 used to manipulate the data x from the first antenna, and w_2 used to manipulate the data x from the second antenna. The beamforming weights, w_1 and w_2 , are applied at the transmitter before transmission. The transmitted signal from the first antenna passes through the environment modeled by h_1 , and the transmitted signal from the second antenna passes through the environment modeled by h_2 .

One may also improve the robustness of the link by adding extra antennas to the receiver with diversity combining, termed single-input, multiple-output. An example of a two receive antenna array system is described mathematically and in vector notation in Eq. (3.5):

$$\begin{aligned} y_1 &= \sqrt{\rho} \cdot h_1 \cdot x + z_1 \\ y_2 &= \sqrt{\rho} \cdot h_2 \cdot x + z_2 \\ \begin{bmatrix} y_1 \\ y_2 \end{bmatrix} &= \sqrt{\rho} \begin{bmatrix} h_1 \\ h_2 \end{bmatrix} x + \begin{bmatrix} z_1 \\ z_2 \end{bmatrix} \end{aligned} \quad (3.5)$$

With this type of system a maximal-ratio combining (MRC) receiver could be used to optimally combine the received signals (Jakes, 1974). The output of the maximal ratio

combiner is given in Eq. (3.6). Note that perfect knowledge of the channel has been assumed.

$$\begin{aligned} r &= [h_1^* \quad h_2^*] \begin{bmatrix} y_1 \\ y_2 \end{bmatrix} \\ &= \sqrt{\rho} \cdot (|h_1|^2 + |h_2|^2) \cdot x + [h_1^* \quad h_2^*] \begin{bmatrix} z_1 \\ z_2 \end{bmatrix} \end{aligned} \quad (3.6)$$

After the MRC function, the received signal may be equalized as follows:

$$\begin{aligned} \hat{x} &= \frac{1}{\sqrt{\rho}} \cdot \frac{1}{|h_1|^2 + |h_2|^2} \cdot r \\ &= x + \frac{1}{\sqrt{\rho}} \cdot \frac{1}{|h_1|^2 + |h_2|^2} \cdot [h_1^* \quad h_2^*] \begin{bmatrix} z_1 \\ z_2 \end{bmatrix} \end{aligned} \quad (3.7)$$

These examples of transmit beamforming and receive diversity combining illustrate two antennas. However, these techniques are not limited to two antennas and extension to three, four, or more antennas is straightforward.

Using the combination of transmit beamforming and receive diversity combining improves robustness of the link in what some call a MIMO system. These techniques have been used with IEEE 802.11a/g to increase the range of a given data rate. However, they do not increase the maximum data rate beyond 54 Mbps, the maximum data rate in 802.11a/g.

3.3

SDM basics

A MIMO system may be used to transmit independent data streams (or spatial streams) on different antennas. We define spatial streams as streams of bits transmitted over separate spatial dimensions. When multiple spatial streams are used with MIMO this is termed spatial division multiplexing (SDM), and is illustrated in Figure 3.3. In the previous example of Figure 3.2 note that each antenna transmits the same sequence x , whereas in Figure 3.3 x_1 and x_2 represent independent streams.

Thus, with MIMO/SDM, the maximum data rate of the system increases as a function of the number of independent data streams. The system must contain the same number or more transmit (Tx) antennas as data streams. With a linear receiver, the system must

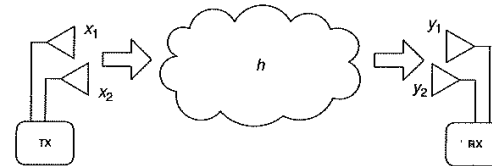


Figure 3.3 MIMO/SDM system.

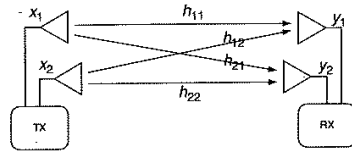


Figure 3.4 Mathematical model of the MIMO/SDM system.

contain the same number or more receive (Rx) antennas as data streams. In other words the data rate of the system increases by $\min(\text{Tx antennas, Rx antennas, data streams})$.

A more detailed graphical view of a MIMO/SDM system with two transmit antennas (and two spatial streams) and two receive antennas is given in Figure 3.4. The notation used to describe such systems is “number of Tx antennas” $\times$ “number of Rx antennas;” thus Figures 3.3 and 3.4 represent a 2×2 MIMO/SDM system.

An $M \times N$ MIMO/SDM system is represented more generally by Eq. (3.8). Here it has been assumed that the total transmit power is equally divided over the M transmit antennas:

$$\begin{aligned} y_1 &= \sqrt{\rho/M} \cdot h_{11} \cdot x_1 + \sqrt{\rho/M} \cdot h_{12} \cdot x_2 + \cdots + \sqrt{\rho/M} \cdot h_{1M} \cdot x_M + z_1 \\ y_2 &= \sqrt{\rho/M} \cdot h_{21} \cdot x_1 + \sqrt{\rho/M} \cdot h_{22} \cdot x_2 + \cdots + \sqrt{\rho/M} \cdot h_{2M} \cdot x_M + z_2 \\ &\vdots \\ y_N &= \sqrt{\rho/M} \cdot h_{N1} \cdot x_1 + \sqrt{\rho/M} \cdot h_{N2} \cdot x_2 + \cdots + \sqrt{\rho/M} \cdot h_{NM} \cdot x_M + z_N \end{aligned} \quad (3.8)$$

The set of equations in Eq. (3.8) may be converted to vector form, as shown in Eq. (3.9):

$$\begin{bmatrix} y_1 \\ y_2 \\ \vdots \\ y_N \end{bmatrix} = \sqrt{\rho/M} \begin{bmatrix} h_{11} & h_{12} & \cdots & h_{1M} \\ h_{21} & h_{22} & \cdots & h_{2M} \\ \vdots & \vdots & \ddots & \vdots \\ h_{N1} & h_{N2} & \cdots & h_{NM} \end{bmatrix} \begin{bmatrix} x_1 \\ x_2 \\ \vdots \\ x_M \end{bmatrix} + \begin{bmatrix} z_1 \\ z_2 \\ \vdots \\ z_N \end{bmatrix} \quad (3.9)$$

We further simplified Eq. (3.9) with matrix notation, as given in Eq. (3.10):

$$Y_N = \sqrt{\rho/M} \cdot H_{N \times M} X_M + Z_N \quad (3.10)$$

where

$$Y_N = \begin{bmatrix} y_1 \\ y_2 \\ \vdots \\ y_N \end{bmatrix}, \quad X_M = \begin{bmatrix} x_1 \\ x_2 \\ \vdots \\ x_M \end{bmatrix}, \quad Z_N = \begin{bmatrix} z_1 \\ z_2 \\ \vdots \\ z_N \end{bmatrix}, \quad H_{N \times M} = \begin{bmatrix} h_{11} & h_{12} & \cdots & h_{1M} \\ h_{21} & h_{22} & \cdots & h_{2M} \\ \vdots & \vdots & \ddots & \vdots \\ h_{N1} & h_{N2} & \cdots & h_{NM} \end{bmatrix}$$

With OFDM, each subcarrier is described by Eq. (3.10).

A generalization of the Shannon capacity formula for M transmit antennas and N receive antennas is given by Eq. (3.11) (Foschini, 1996):

$$C(\text{bps/Hz}) = \log_2[\det(I_N + \rho/M \cdot H H^*)] \quad (3.11)$$

where H^* is the conjugate transpose of H .

In matrix notation, the MIMO/SDM system described in Eq. (3.10) has a similar mathematical structure to the SISO system given in Eq. (3.1). Conceptually, the equalizer for the MIMO/SDM system may also be designed with a similar structure as the SISO equalizer in Eq. (3.3). In matrix form, the received signal is divided by the channel to extract the estimate of the transmitted data, given in Eq. (3.12):

$$\begin{aligned} \hat{X} &= (\sqrt{\rho/M} \cdot H)^{-1} Y \\ &= X + (\sqrt{\rho/M} \cdot H)^{-1} Z \end{aligned} \quad (3.12)$$

The equalizer is given in Eq. (3.12) as a matrix inversion for simplicity, but is actually only valid where $M = N$. The representation of the data estimate is generalized for $M = N$ and $M \neq N$ as follows:

$$\begin{aligned} \hat{X} &= \frac{H^*}{\sqrt{\rho/M} \cdot H^* H} Y \\ &= X + \frac{H^*}{\sqrt{\rho/M} \cdot H^* H} Z \end{aligned}$$

It almost seems too good to be true that with such simplicity one may increase the data rate of a system by merely adding additional transmit and receive antennas. Through the use of an example it is shown that there are bounds on this increase. Consider the SISO system of Eq. (3.1), where now the antenna port of the transmitting device is directly connected to the antenna port of the receiving device. Such is a typical setup for conductive testing. Mathematically, h is equal to 1, and $y = \sqrt{\rho} \cdot x + z$. With enough transmit power, the system is very reliable. Now consider a 2×2 MIMO/SDM system with each antenna port of the transmitter device connected to each antenna port of the receiver device. The channel H for such system is equal to $\begin{bmatrix} 1 & 1 \\ 1 & 1 \end{bmatrix}$. To extract the transmitted data from the receive channel, we must perform an inverse of the channel matrix. In this case, the matrix is singular and cannot be inverted. The receiver fails, even in high SNR. As we see, the channel matrix must be well conditioned for the MIMO system to achieve large increases in data rate from spatial multiplexing.

3.4

MIMO environment

If the terms in the channel matrix are randomly distributed and uncorrelated with each other, statistically the channel matrix is well conditioned and invertible the vast majority of the time. Fortunately, a fading multipath channel creates such an environment.

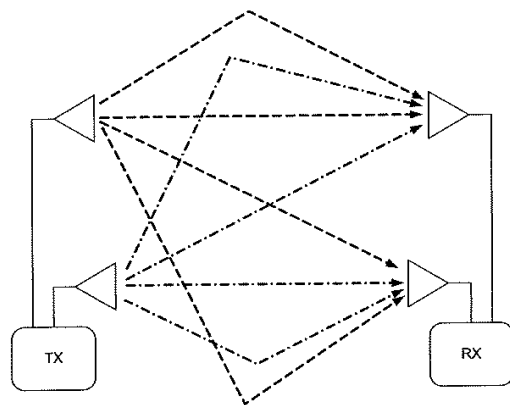


Figure 3.5 Multipath fading environment.

In an indoor environment, rays bounce off floors, ceilings, walls, furniture, etc. when propagating between the transmitter and the receiver (Rappaport, 1996). The paths of the rays are different as a function of the location of the transmit and receive antennas, as illustrated in Figure 3.5. How the paths propagate determines the amount of fading due to cancellation, delay, and correlation.

With a channel matrix having terms which are complex Gaussian distributed, uncorrelated with each other, and the same channel matrix across all subcarriers, a quantitative benefit of MIMO is realized. In a wideband OFDM system such as 802.11n, such a model represents a flat Rayleigh fading across the band. This is not very realistic, but allows for a simple analytical comparison between SISO and MIMO capacity. A more realistic channel model for 802.11n is provided in Section 3.5.

In Figure 3.6 a 2×2 MIMO system is compared with a SISO system and demonstrates the large gain that is obtained with MIMO systems. The complementary cumulative distribution function (CCDF) is plotted versus capacity to show how often the capacity exceeds a certain level. Capacity of each system is simulated by Eq. (3.11) at an SNR equal to 20 dB. For reference the maximum data rate of 802.11a is 54 Mbps in a 20 MHz channel, resulting in a spectral efficiency of 2.7 bps/Hz. In a flat fading environment, with ideal conditions (i.e. no hardware impairments, synchronization loss, etc.) such a capacity is achieved 95% of the time. At a comparable percentage, a 2×2 MIMO system achieves three times the capacity. In reality, various impairments and realistic channel environments inhibit the ability to increase the capacity by that amount in all situations. But in subsequent chapters it is shown that a two stream 802.11n system has a maximum data rate which is 2.8 times that of 802.11a in a 20 MHz channel, so this three fold increase can be approached in some circumstances.

A significant impact on MIMO capacity is antenna correlation. As the correlation increases, the condition of the channel matrix degrades. To ensure that the channel is

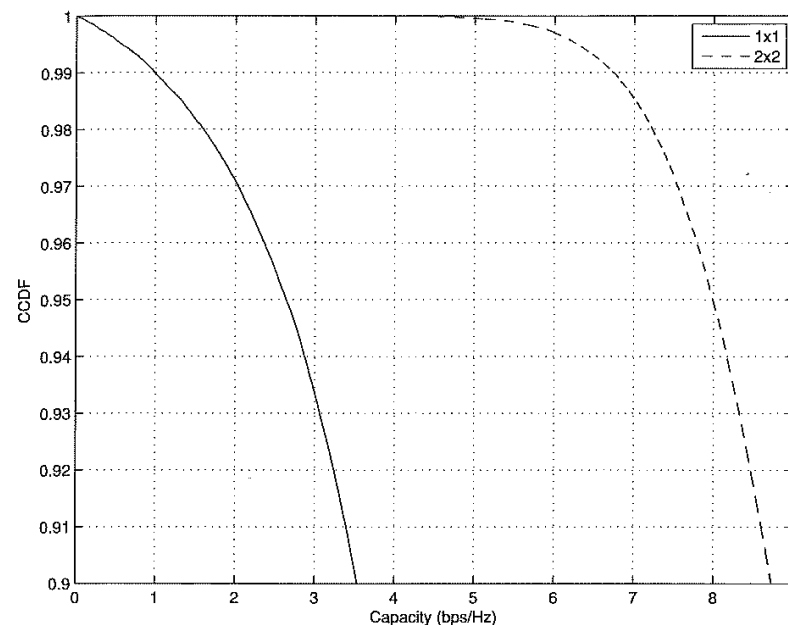


Figure 3.6 Capacity comparison between 2×2 MIMO and SISO with SNR = 20 dB.

uncorrelated “enough,” an antenna spacing of at least half a wavelength is generally used. In addition, the antennas should be designed with low mutual coupling between the elements to minimize channel correlation. A non-line-of-sight (NLOS) environment may also reduce correlation since highly correlated direct paths between the transmit and receive devices are not present. However, this typically means a reduction in SNR at the receiver.

Performance of a MIMO/OFDM system varies with many environmental conditions: LOS or NLOS based on obstructions, delay spread from delayed responses, antenna correlation, and Doppler from mobility. These aspects of the environment are addressed in Section 3.5 where the 802.11n channel model is presented.

3.5

802.11n propagation model

A set of channel models and path loss models were created during the development of the 802.11n standard (Erceg *et al.*, 2004). Due to the nature of typical WLAN deployments, the propagation models were developed based mostly on indoor measurements. A key

Table 3.1 Channel models

RMS delay			
Model	Spread (ns)	Environment	Example
A	0	N/A	N/A
B	15	Residential	Intra-room, room-to-room
C	30	Residential/small office	Conference room, classroom
D	50	Typical office	Sea of cubes, large conference room
E	100	Large office	Multi-story office, campus small hotspot
F	150	Large space (indoors/outdoors)	Large hotspot, industrial, city square

component of the channel models which affects MIMO system performance is the correlation matrix.

The channel models are composed of impulse responses for a range of indoor environments. Measurement data from both the 2.4 GHz and 5 GHz frequency bands were incorporated to develop a set of impulse responses and antenna correlation matrices applicable to both bands. The channel model is a function of frequency, as well as Doppler.

3.5.1 Impulse response

There are six channel models, each with a different impulse response corresponding to different indoor environments (Stephens *et al.*, 2004). An overview of the models is given in Table 3.1.

Channel model A is flat fading Rayleigh, and is not applicable to a wideband system like 802.11. This channel model was only included for its utility in analytic modeling. The most commonly used channel models are B, D, and E. These three models were largely used to compare proposals during the development of the 802.11n standard as outlined in Stephens (2004). These were also used for physical layer (PHY) packet error rate (PER) performance. In addition, the three simulation scenarios selected for MAC network simulations in Stephens (2004), residential, large enterprise, and hot spot, also use channel models B, D, and E, respectively, as described in Stephens *et al.* (2004).

The designs of the impulse responses are based on the cluster model initially developed by Saleh and Valenzuela (1987). In Figure 3.7, the vertical bars indicate a delayed response. These are located on 10 ns intervals. The lines indicate the extent of each cluster. Notice that the clusters overlap, and the delayed responses during the overlap are a composite of the power from the overlapping clusters. In addition, the power of the delayed responses of each cluster decays linearly on a log-scale. The power delay profile, separated into clusters, is given in Appendix 3.1 for each channel model. To compute the power for each tap at each delay, the powers of the taps in overlapping clusters are summed at each delay.

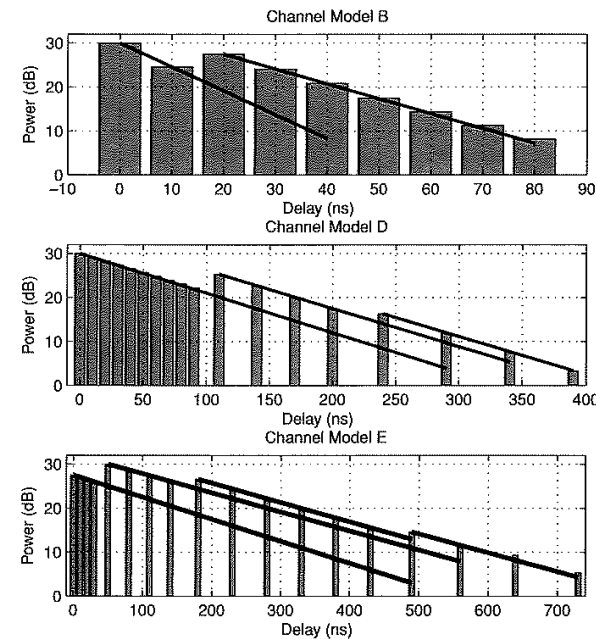


Figure 3.7 Impulse response of selected channel models.

Each tap of the impulse response, h , is Ricean distributed and comprises a fixed and random component. The fixed component is due to a constant LOS path between the transmitter and receiver. The NLOS component is random and Rayleigh distributed:

$$h = \sqrt{P} \left(\sqrt{\frac{K}{K+1}} e^{j\phi} + \sqrt{\frac{1}{K+1}} X \right) \quad (3.13)$$

where X is a complex Gaussian random variable with zero mean and unit variance, ϕ is derived from the angle of arrival/departure of the LOS component, K is the Ricean K -factor, and P is the power of the tap. To compute the total power for each tap at each delay, the powers of the taps in overlapping clusters, given in Appendix 3.1, are summed at each delay.

The K -factor values for each channel model are given in Table 3.2. The K -factor only applies to the first tap of the impulse response; all other taps have a K -factor of 0. For a NLOS channel, the K -factor for all taps is 0. The distance between the transmitter and receiver determines whether the channel should be modeled as LOS or NLOS.

Table 3.2 K-factor

Channel	LOS		NLOS	
	K-factor Tap index 1	K-factor All other taps	K-factor Tap index 1	K-factor All other taps
A	1	0	0	0
B	1	0	0	0
C	1	0	0	0
D	2	0	0	0
E	4	0	0	0
F	4	0	0	0

This is discussed further in Section 3.5.5. For the 802.11n channel model the angle of arrival/departure of the LOS component is fixed at 45° .

3.5.2 Antenna correlation

The model for the impulse response outlined in Section 3.5.1 and detailed in Eq. (3.13) represents a SISO model. For a MIMO system, the taps are created for each element of the H matrix in Eq. (3.9). Therefore, h in Eq. (3.13) is modified for the i th receive antenna and j th transmit antenna pair, as follows:

$$h_{ij} = \sqrt{P} \left(\sqrt{\frac{K}{K+1}} e^{j\phi_{ij}} + \sqrt{\frac{1}{K+1}} X_{ij} \right) \quad (3.14)$$

There is an H matrix for each tap of the impulse response in a MIMO channel:

$$H = \sqrt{P} \left(\sqrt{\frac{K}{K+1}} \begin{bmatrix} e^{j\phi_{11}} & e^{j\phi_{12}} & \dots & e^{j\phi_{1M}} \\ e^{j\phi_{21}} & e^{j\phi_{22}} & \dots & e^{j\phi_{2M}} \\ \vdots & \vdots & \ddots & \vdots \\ e^{j\phi_{N1}} & e^{j\phi_{N2}} & \dots & e^{j\phi_{NM}} \end{bmatrix} + \sqrt{\frac{1}{K+1}} \begin{bmatrix} X_{11} & X_{12} & \dots & X_{1M} \\ X_{21} & X_{22} & \dots & X_{2M} \\ \vdots & \vdots & \ddots & \vdots \\ X_{N1} & X_{N2} & \dots & X_{NM} \end{bmatrix} \right) \quad (3.15)$$

Correlation is then applied to the random elements X_{ij} to incorporate antenna correlation into the channel model, as follows:

$$[X] = [R_{Rx}]^{1/2} [\tilde{X}] ([R_{Tx}]^{1/2})^T \quad (3.16)$$

where R_{Rx} and R_{Tx} are the receive and transmit correlation matrices, respectively. $[R]^{1/2}$ is defined as a matrix square root, where $[R]^{1/2} \cdot [R]^{1/2} = R$. $\tilde{X}$ is an independent, complex Gaussian random variable with zero mean and unit variance. The correlation

matrices, R_{Rx} and R_{Tx} , are defined as follows:

$$R_{Tx} = \begin{bmatrix} 1 & \rho_{Tx12} & \rho_{Tx13} & \dots & \rho_{Tx1M} \\ \rho_{Tx21} & 1 & \rho_{Tx23} & \dots & \rho_{Tx2M} \\ \rho_{Tx31} & \rho_{Tx32} & 1 & \dots & \vdots \\ \vdots & \vdots & \vdots & \ddots & \vdots \\ \rho_{TxN1} & \rho_{TxN2} & \dots & \rho_{TxN(M-1)} & 1 \end{bmatrix} \quad (3.17)$$

$$R_{Rx} = \begin{bmatrix} 1 & \rho_{Rx12} & \rho_{Rx13} & \dots & \rho_{Rx1M} \\ \rho_{Rx21} & 1 & \rho_{Rx23} & \dots & \rho_{Rx2M} \\ \rho_{Rx31} & \rho_{Rx32} & 1 & \dots & \vdots \\ \vdots & \vdots & \vdots & \ddots & \vdots \\ \rho_{RxN1} & \rho_{RxN2} & \dots & \rho_{RxN(M-1)} & 1 \end{bmatrix}$$

where ρ_{Txij} are the complex correlation coefficients between the i th and j th transmitting antennas, and ρ_{Rxij} are the complex correlation coefficients between the i th and j th receiving antennas.

3.5.2.1 Correlation coefficient

In the 802.11n channel model, a complex correlation coefficient is derived based on the power angular spectrum (PAS) formulation. This section follows the description given in Erceg *et al.* (2004), Salz and Winters (1994), and Schumacher *et al.* (2002).

The PAS for each tap is a function of the angular spread (AS) and angle of incidence (angle of arrival (AoA) or angle of departure (AoD), depending on Tx or Rx) of each cluster. The angular spread and angle of incidence for each cluster (the 802.11n model assumes that all taps in a cluster have the same angular spread and angle of incidence) is given for each channel model in Appendix 3.1. The shape of the PAS distribution commonly used for 802.11n is truncated Laplacian. The PAS distribution over the angle for each tap is given by

$$\text{PAS}(\phi) = \frac{1}{A} \sum_{k=1}^{N_C} \frac{p_k}{\sigma_k} \exp \left[\frac{-\sqrt{2} |\phi - \psi_k|}{\sigma_k} \right] \quad (3.18)$$

where N_C is the number of clusters, and for each cluster k , p_k is the tap power, σ_k is the tap AS, and ψ_k is the tap angle of incidence. Since the PAS is a probability density function, it must fulfill the requirement that $\int_{-\pi}^{\pi} \text{PAS}(\phi) d\phi = 1$. Therefore A is equal to $\int_{-\pi}^{\pi} \sum_{k=1}^{N_C} (p_k/\sigma_k) \exp[\sqrt{2} |\phi - \psi_k|/\sigma_k] d\phi$. Figure 3.8 illustrates the distribution function for the third tap of channel model B for each cluster for Rx (using parameters from Table 3.5). The sum over the clusters at each angle results in $\text{PAS}(\phi)$.

For a uniform linear antenna array, the correlation of the fading between two antennas spaced D apart is described by Lee (1973). The correlation functions are given in Erceg *et al.* (2004), as follows:

$$R_{XX}(D) = \int_{-\pi}^{\pi} \cos \left(\frac{2\pi D}{\lambda} \sin \phi \right) \text{PAS}(\phi) d\phi \quad (3.19)$$

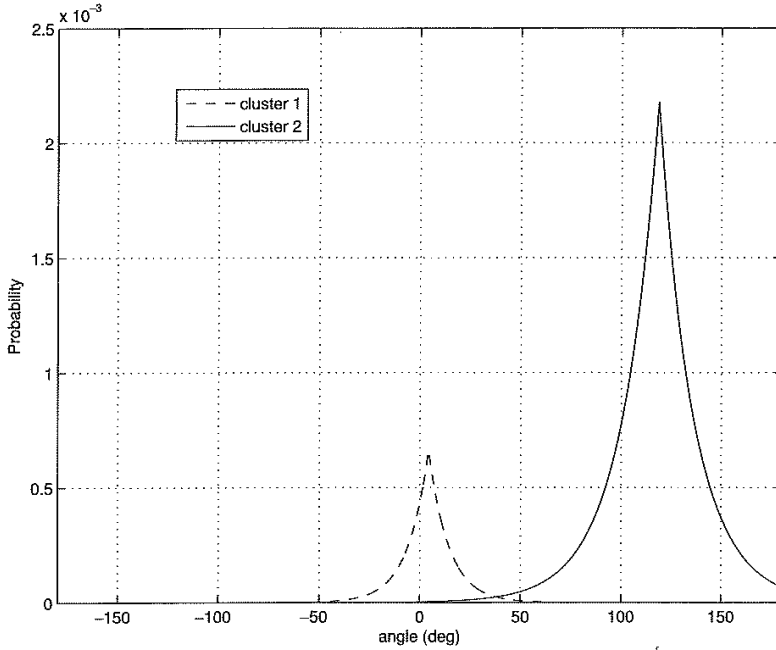


Figure 3.8 Distribution function for each cluster.

and

$$R_{XY}(D) = \int_{-\pi}^{\pi} \sin\left(\frac{2\pi D}{\lambda} \sin \phi\right) \text{PAS}(\phi) d\phi \quad (3.20)$$

where R_{XX} is the correlation function between the real parts of the fading (or the imaginary parts), R_{XY} is the correlation function between the real and imaginary parts of the fading, and λ is the wavelength.

As an example, for the receiver and tap index 3 of channel model B (using parameters from Table 3.5), R_{XX} is equal to -0.52 and R_{XY} is equal to 0.38 with D equal to $\lambda/2$.

The complex correlation coefficients ρ_{TXij} between the i th and j th transmitting antennas, and ρ_{RXij} between the i th and j th receiving antennas in Eq. (3.17) is described by Eq. (3.21) (Erceg *et al.*, 2004):

$$\rho = R_{XX}(D) + jR_{XY}(D) \quad (3.21)$$

To continue the third tap example, the antenna correlation matrix R_{RX} is equal to

$$\begin{bmatrix} 1 & -0.52 + j \cdot 0.38 \\ -0.52 - j \cdot 0.38 & 1 \end{bmatrix}$$

When comparing proposals in the development of 802.11n, the antenna spacing was set to $\lambda/2$ in Stephens (2004).

3.5.3 Doppler model

Typically in an indoor WLAN model the transmitter and the receiver are stationary. However, people moving in between the transmitter and receiver could cause the channel to change. If the total transmission time of the packet or packet exchange is long enough to be affected by this motion, a Doppler model may be included in the fading characteristics of the impulse response. Due to the nature of the indoor environment, the Doppler spectrum is quite different from the typical models for mobile cellular channels. A bell shaped Doppler spectrum is used in the 802.11n channel model (Erceg *et al.*, 2004), as follows:

$$S(f) = \frac{1}{1 + A \left(\frac{f}{f_d}\right)^2} \quad |f| \leq f_{\max} \quad (3.22)$$

where f_d is the Doppler spread, $f_{\max}$ is the maximum frequency component of the Doppler power spectrum, and A is a constant used to set the value of $S(f)$.

In the 802.11n channel model the value of $S(f_d)$ is set to 0.1, therefore A is equal to 9. The Doppler spread f_d is defined as v_0/λ , where v_0 is the environmental speed. The value for v_0 in 802.11n is equal to 1.2 km/h. Therefore the values for Doppler spread in the 5 GHz band is approximately 6 Hz and in the 2.4 GHz band is approximately 3 Hz. $f_{\max}$ limits the range of frequencies, and in the 802.11n channel model, $f_{\max}$ is set to five times f_d .

The 802.11n channel model Doppler spectrum for 2.4 GHz and 5.25 GHz is illustrated in Figure 3.9. As a note, the relative bandwidth of the Doppler spectrum is independent of the absolute carrier frequency, which is seen in Eq. (3.22).

The autocorrelation function of the bell shaped spectrum is given by $R(\Delta t) = (\pi f_d / \sqrt{A}) \cdot \exp(-(2\pi f_d / \sqrt{A}) \cdot \Delta t)$ and the coherence time is given by $T_c = (\sqrt{A} / 2\pi f_d) \cdot \ln(2)$.

3.5.3.1 Modified Doppler model for channel model F

Channel model F is used for large indoor or outdoor environments for large hotspot or industrial conditions, as in Table 3.1. The Doppler model for channel model F includes the scenario of reflections off of a moving object, such as a vehicle. Therefore for channel model F, an extra Doppler component was added to the third tap to account for

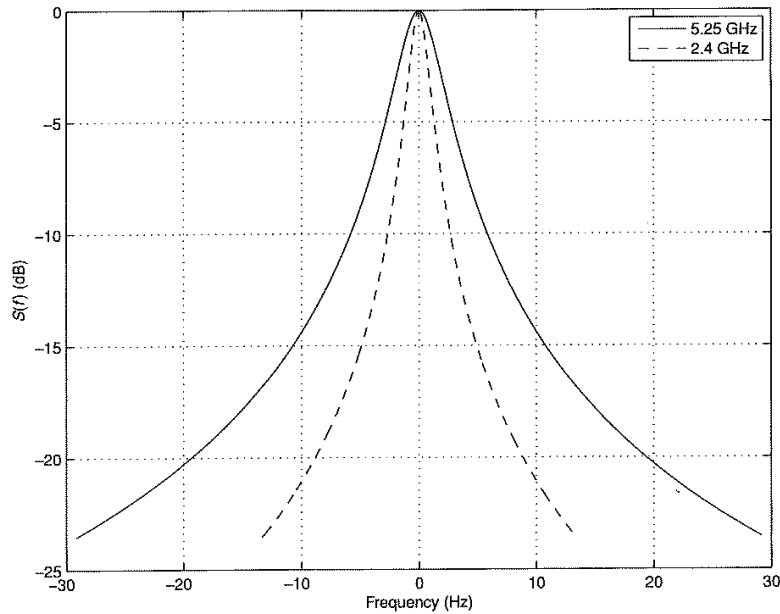


Figure 3.9 Doppler spectrum. Reproduced with permission from Erceg *et al.* (2004) © IEEE.

this higher speed component, as in Eq. (3.23):

$$S(f) = \frac{1}{1 + 9 \left(\frac{f}{f_d} \right)^2} + \frac{1}{2} \frac{1}{1 + B \left(\frac{f - f_{veh}}{f_{veh}} \right)^2} \quad |f| \leq f_{max} \quad (3.23)$$

where f_{veh} is defined as v_1/λ , where v_1 is the speed of the moving vehicle in the environment. The value for v_1 in the 802.11n channel model is equal to 40 km/h. The term B is used to set the bandwidth of the extra Doppler component. This extra Doppler component causes a spike at a frequency f_{veh} . The bandwidth of this spike is defined to be $0.02 \cdot f_{veh}$ when the spectrum is 10 dB below the peak of the spike. Solving for B , by setting $S(f_{veh}(1 \pm \frac{0.02}{2}))/S(f_{veh}) = 0.1$, results in B equal to 90 000. We define f_{max} to be $f_{veh} \cdot (1 + 5 \cdot (0.02/2)) + 5 \cdot f_d$.

The Doppler spectrum for the third tap of channel model F at a center frequency of 2.4 GHz and 5.25 GHz is illustrated in Figure 3.10. Note that the relative bandwidth and location of the spike in the Doppler spectrum is independent of the absolute carrier frequency, which is seen in Eq. (3.23).

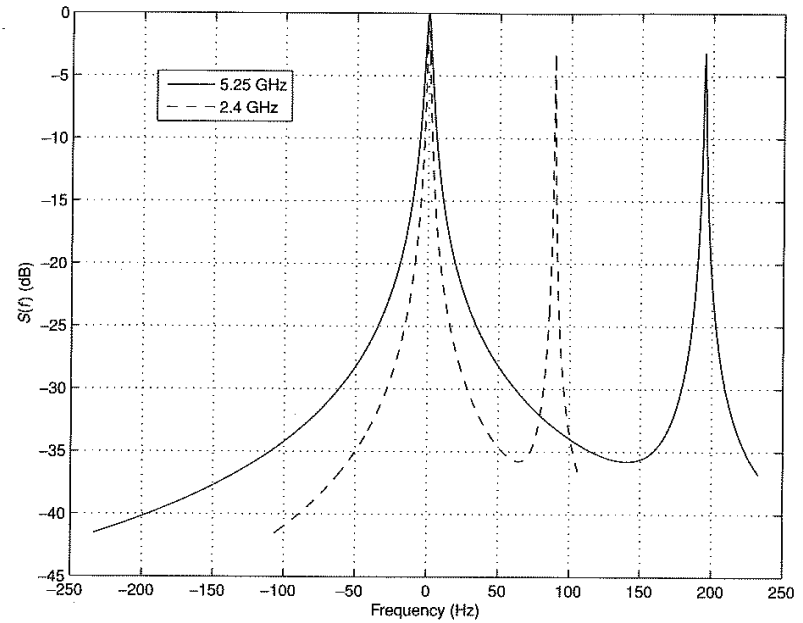


Figure 3.10 Doppler spectrum for third tap of channel model F. Reproduced with permission from Erceg (2004) © IEEE.

3.5.4 Physical layer impairments

Several physical layer (PHY) impairments were added to the channel model. These include Tx and Rx phase noise, power amplifier (PA) non-linearity, and carrier frequency and clock symbol offset. The noise figure is addressed in Section 3.5.5.

When comparing proposals in the development of 802.11n, the carrier frequency offset was set to -13.675 ppm at the receiver, with the clock symbol having the same relative offset.

3.5.4.1 Phase noise

Systems with high order modulation are sensitive to phase noise. Therefore, when comparing proposals for 802.11n, phase noise was added at both the transmitter and receiver in the PHY layer simulations. A single-pole, single-zero phase noise model was

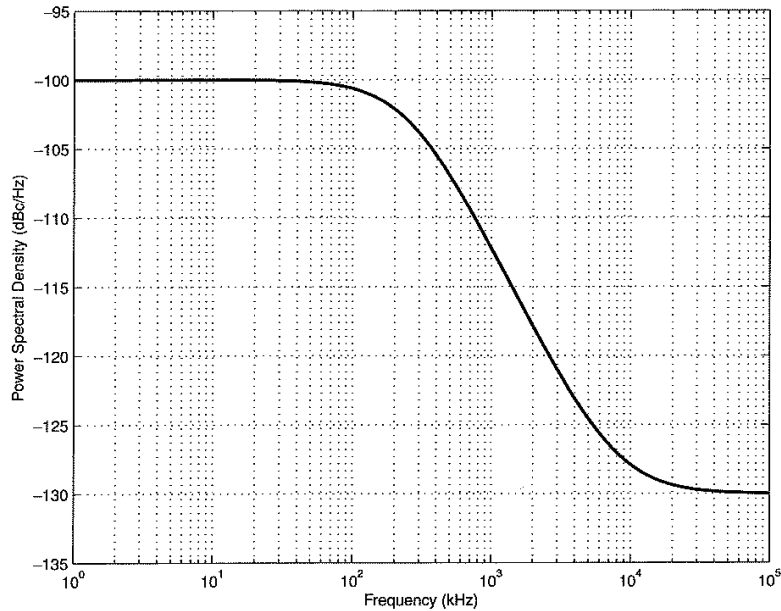


Figure 3.11 Phase noise power spectral density.

utilized with the following specifications (Stephens *et al.*, 2004):

$$\begin{aligned} \text{PSD}(f) &= \text{PSD}(0) \frac{[1 + (f/f_z)^2]}{[1 + (f/f_p)^2]} \\ \text{PSD}(0) &= -100 \text{ dBc/Hz} \\ f_p &= 250 \text{ kHz} \\ f_z &= 7905.7 \text{ kHz} \end{aligned} \quad (3.24)$$

where PSD is the phase noise power spectral density, f_p is the pole frequency, and f_z is the zero frequency. This model results in $\text{PSD}(\infty)$ equal to -130 dBc/Hz . The phase noise power spectral density is illustrated in Figure 3.11. Some people believe this to be a lower phase noise than could typically be achieved in practice.

3.5.4.2 Power amplifier non-linearity

A power amplifier (PA) of a certain size, class, and drive current is most efficiently used when driven to saturation for maximum output power. However, when operated at saturation, the PA exhibits non-linear behavior when amplifying the input signal. The distortion caused by the PA on the transmitted waveform causes spectral re-growth,

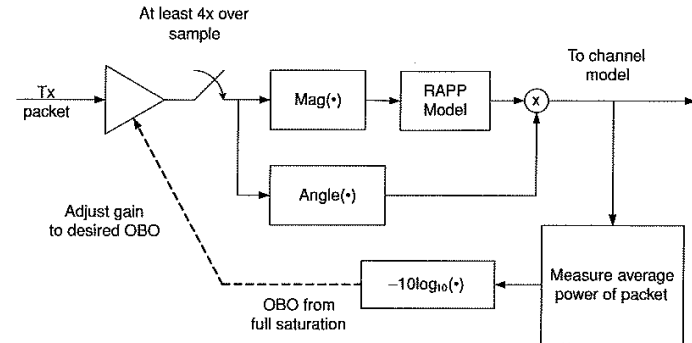


Figure 3.12 Power amplifier model.

impacting the transmitter's ability to meet the specified spectral mask. In addition, the distortion impacts the Tx error vector magnitude (EVM), increasing the packet error rate at the receiver.

The 802.11n MIMO/OFDM system is especially sensitive to PA non-linearity, since the transmitted waveform has high dynamic range and uses high order modulation. Therefore in order to properly model the PHY, especially with high order QAM modulation, a model for PA non-linearity was included in the PHY simulations when comparing proposals.

Figure 3.12 illustrates the selected model for the power amplifier adapted from Webster (2000) to be used in simulations during the development of 802.11n. The gain on the transmitted packet is adjusted to the desired output backoff (OBO). With the waveform over-sampled by at least a factor of four, a non-linear distortion is applied to the amplitude of the signal. The Rapp PA model was selected as the function to model this distortion, and is given in Eq. (3.25):

$$A_{\text{out}} = \frac{A_{\text{in}}}{(1 + A_{\text{in}}^{2p})^{\frac{1}{2p}}} \quad (3.25)$$

For the simulation of different proposals for the 802.11n development, p is set to 3 (Stephens, 2004). The PA distortion between the input amplitude and output amplitude is illustrated in Figure 3.13.

For simulations, the recommended transmitted power, at full saturation, was 25 dBm. The total transmit power was limited to no more than 17 dBm. Therefore, with the recommended settings, the OBO from full saturation is 8 dB.

Due to the non-linear structure of the power amplifier model, one way to achieve the desired OBO of the transmitted waveform is to iteratively adjust the gain. This is depicted in Figure 3.12 by the dashed line between the measure of the OBO and the gain applied to the transmitted packet.

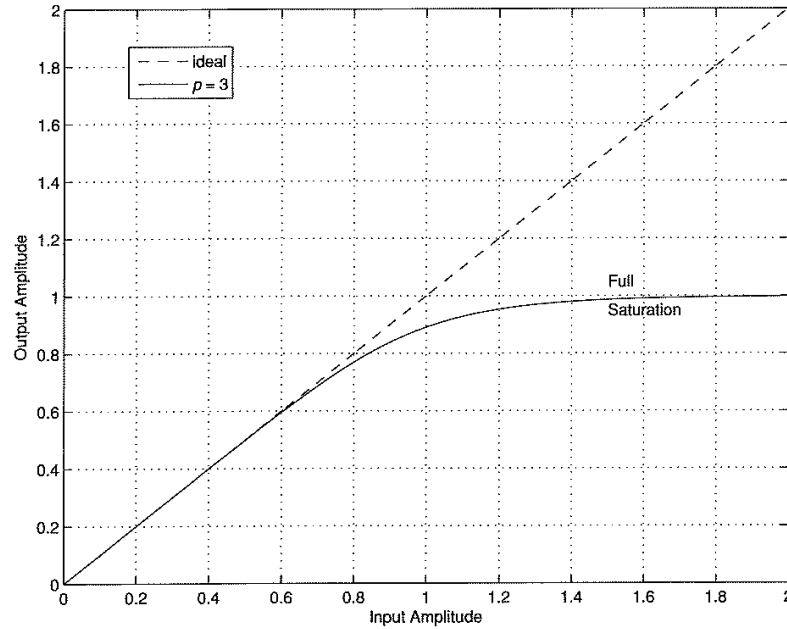


Figure 3.13 Rapp model.

3.5.5 Path loss

To determine the achievable range between the transmitter and receiver, a path loss model and the noise figure of the receiver are required. For 802.11n proposal comparisons, the input referenced total noise figure from the antenna to the output of the ADC was set to 10 dB in Stephens (2004). Real hardware implementations typically achieve a better noise figure.

The path loss model used for indoor propagation was defined in Erceg *et al.* (2004) and consists of the free space (FS) loss (slope of 2) up to a breakpoint distance, and a slope of 3.5 after the breakpoint distance. The breakpoint distances for the channel models are given in Table 3.3. Also included in the path loss model is the shadow fading loss due to large scale obstructions. The path loss model is given as follows:

$$\begin{aligned} L(d) &= L_{FS}(d) + SF & d \leq d_{BP} \\ L(d) &= L_{FS}(d_{BP}) + 35 \log_{10} \left(\frac{d}{d_{BP}} \right) + SF & d > d_{BP} \end{aligned} \quad (3.26)$$

Table 3.3 Path loss model parameters

Channel model	Breakpoint distance d_{BP} (m)	Path loss slope		Shadow fading std. dev. (dB)		Channel conditions	
		Before d_{BP}	After d_{BP}	Before d_{BP}	After d_{BP}	Before d_{BP}	After d_{BP}
A	5	2	3.5	3	4	LOS	NLOS
B	5	2	3.5	3	4	LOS	NLOS
C	5	2	3.5	3	5	LOS	NLOS
D	10	2	3.5	3	5	LOS	NLOS
E	20	2	3.5	3	6	LOS	NLOS
F	30	2	3.5	3	6	LOS	NLOS

where d is the separation distance between the transmitter and receiver in meters, L_{FS} is the free space path loss in dB, d_{BP} is the breakpoint distance in meters, and SF is the shadow fading loss in dB. The definition for free space path loss is

$$L_{FS}(d) = 20 \log_{10}(d) + 20 \log_{10}(f) - 147.5 \quad (3.27)$$

where $L_{FS}(d)$ is in dB and f is the frequency in Hz.

The shadow fading loss is modeled by a log-normal distribution (Gaussian in dB) with zero mean, as follows:

$$p_{SF}(x) = \frac{1}{\sqrt{2\pi}\sigma_{SF}} \exp\left(-\frac{x^2}{2\sigma_{SF}^2}\right) \quad (3.28)$$

where σ_{SF} is the standard deviation of the shadow fading. The breakpoint distance is also the transition for the shadow fading standard deviation. Table 3.3 gives the shadow fading standard deviation before and after the breakpoint.

As discussed in Section 3.5.1 and in Table 3.2, the Ricean K-factor for the taps of the impulse response are a function of whether the channel should be modeled as LOS or NLOS. For all channel models, LOS conditions are used before the breakpoint and NLOS after the breakpoint, using the parameters outlined in Table 3.3.

3.6

Linear receiver design

Equations (3.3) and (3.12) demonstrate a linear equalization technique to extract data from the received signal commonly known as the zero-forcing (ZF) algorithm, which is described in detail in Proakis (1989). During a large flat fade (h approaches 0), or a deep notch in a frequency selective fading channel, noise is highly amplified in the estimate of the data using a ZF equalizer.

An alternative approach is to select equalizer weights (W) based on minimizing the mean-square-error (MSE). The estimate of x is given in Eq. (3.29), based on the received

signal (Y) given in Eq. (3.10):

$$\begin{aligned} Y &= \sqrt{\rho/M} \cdot H \cdot X + Z \\ \hat{X} &= W \cdot Y \\ &= W \cdot \sqrt{\rho/M} \cdot H \cdot X + W \cdot Z \end{aligned} \quad (3.29)$$

The minimum mean-square-error (MMSE) estimate minimizes the mean-square value of the error vector $e = \hat{X} - X$. With the definition of $\hat{X}$ and Y in Eq. (3.29), the expression for MSE is given by

$$J_{M \times M} = \frac{\rho}{M} W H H^* W^* + W \Phi_Z W^* - 2 \sqrt{\frac{\rho}{M}} \text{Re}(W H) + I \quad (3.30)$$

where $J = E[e \cdot e^*]$, H^* is the conjugate transpose of H , Φ_Z is the noise covariance matrix, and I is the identity matrix from the signal covariance. By minimizing the MSE expression with respect to W , we arrive at a solution for W :

$$W = \sqrt{\frac{\rho}{M}} H^* \left(\frac{\rho}{M} H H^* + \Phi_Z \right)^{-1} \quad (3.31)$$

The diagonal terms of the MSE in Eq. (3.30) are between 0 and 1 (Proakis, 1989) with MMSE weights. The output SNR for the i th data stream for MMSE is given by

$$\text{SNR}_i = \frac{1 - J_i}{J_i} \quad (3.32)$$

where J_i is the i th diagonal element of the MSE matrix given in Eq. (3.30).

On the other hand, with ZF the MSE is unbounded. With ZF (and the simplification of $M = N$ in Eq. (3.10)), the weights are equal to $\left(\sqrt{\rho/M} \cdot H \right)^{-1}$. By replacing W in Eq. (3.30) and setting $\Phi_Z = I$ (since we defined the noise term Z as Normal(0,1)), the MSE for ZF is $((\rho/M) \cdot H^* H)^{-1}$. The output SNR for ZF is the inverse of the diagonal terms of the MSE:

$$\text{SNR}_i = \frac{1}{\text{diag}_i \left(\left(\frac{\rho}{M} \cdot H^* H \right)^{-1} \right)} \quad (3.33)$$

In Figure 3.14, MMSE (dashed line) and ZF (solid line) are compared for three input SNRs, 0, 10, and 20 dB, with two transmit and two receive antennas. Each element of the channel matrix H is modeled as independent, identically distributed Rayleigh fading. At higher input SNR, the output SNR between MMSE and ZF is similar. This is seen by equivalently expressing the weights for ZF as $\sqrt{(\rho/M)} H^* ((\rho/M) H H^*)^{-1}$. Comparing this to the MMSE weights in Eq. (3.31), as the input SNR increases the contribution of Φ_Z goes to zero, the weights for ZF and MMSE converge. However, at lower input SNR levels, the improvement of MMSE over ZF exceeds 5 dB as illustrated in Figure 3.14 at the CCDF level of 90% or more.

The advantage of MMSE over ZF is further demonstrated by examining capacity. The formula for capacity based on output SNR is given by Eq. (3.34):

$$C = \sum_{i=1}^M \log_2 (1 + \text{SNR}_i) \quad (3.34)$$

where M is the number of data streams.

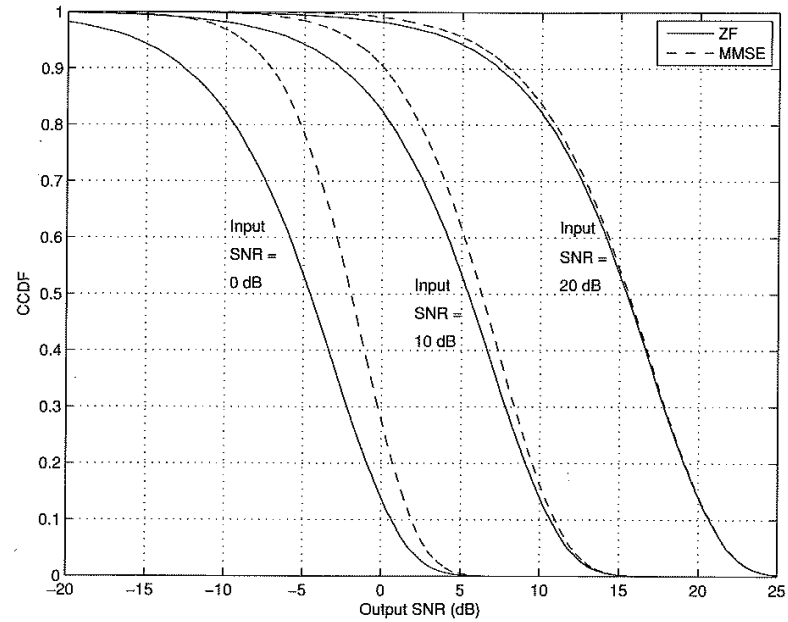


Figure 3.14 Output SNR comparison between MMSE and ZF.

To simulate capacity, the output SNR data provided in Figure 3.14 is substituted into Eq. (3.34). The capacity results for ZF (solid line) and MMSE receiver (dashed line) are illustrated in Figure 3.15 for an input SNR of 0, 10, and 20 dB. The results are compared to the Shannon capacity (dotted line) given by Eq. (3.11). With an input SNR of 0 dB, MMSE capacity is comparable to Shannon capacity at a probability of 90%. With a ZF receiver, the capacity is substantially lower. As the SNR increases, the capacity of MMSE converges to the capacity of ZF. Furthermore, the capacity of MMSE diverges from the Shannon capacity limit.

3.7

Maximum likelihood estimation

An alternative to a linear receiver design is the use of maximum likelihood (ML) estimation. Referring to the communication system defined by Eq. (3.10), an ML estimator maximizes the probability that a signal X_i was transmitted given the received signal Y , as shown below (Proakis, 1989):

$$P(X_i \text{ was transmitted} | Y) = \frac{p(Y | X_i) \cdot P(X_i)}{p(Y)} \quad i = 1, \dots, M \quad (3.35)$$

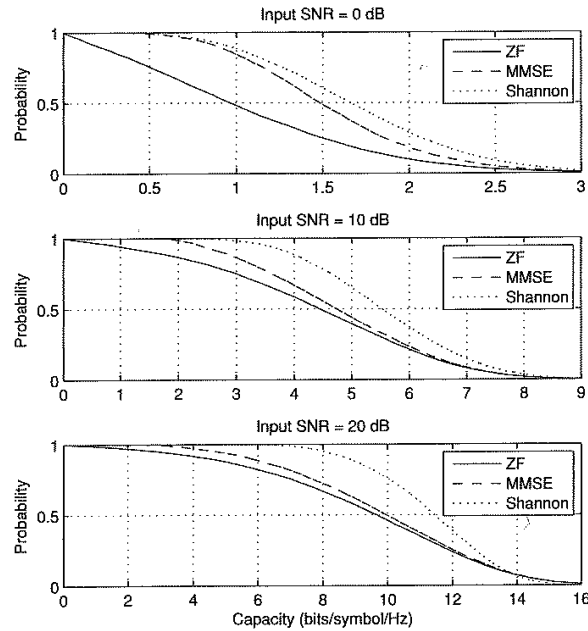


Figure 3.15 Comparison between MMSE and ZF capacity and Shannon capacity.

where X is a set of M equally likely transmitted signals. $p(Y)$ is identical for all signals, and for equally likely signals $P(X_i) = 1/M$ for all i , so neither term affects the decision. Therefore the maximum likelihood criterion becomes (Proakis, 1989)

$$\hat{X} = \arg \max_i p(Y | X_i) \quad i = 1, \dots, M \quad (3.36)$$

With an estimate of the channel at the receiver, $\hat{H}$, this is expanded as follows (van Nee *et al.*, 2000):

$$\hat{X} = \arg \min_i \|Y - \hat{H} \cdot X_i\| \quad (3.37)$$

where the maximum likelihood estimate is a function of the minimum Euclidean distance.

An advantage of ML decoding is in the fact that it obtains a diversity order equal to the number of receive antennas (van Nee *et al.*, 2000). In comparison, a MMSE receiver only obtains a diversity order equal to the number of receive antennas minus the number of spatial streams plus one. For example, a three antenna MMSE receiver receives a two spatial stream transmission with diversity order of two. However, an ML decoder

achieves a diversity order of two with only two receive antennas, thereby reducing the RF implementation complexity and cost. Or conversely, when comparing a two antenna MMSE receiver and a two antenna ML receiver, the ML receiver provides an additional diversity order. Being independent of the number of spatial streams is a very important property, since an ML receiver maintains its diversity order as the number of spatial streams grow with MIMO/SDM.

The main disadvantage of ML is that the estimator complexity grows exponentially with the constellation size and number of spatial streams. In this respect, ML is much more complex than MMSE. For example, with 64-QAM and two spatial streams, ML requires calculating and sorting $64^2 = 4096$ Euclidean norms. With three spatial streams this increases to $64^3 = 262144$.

Sub-optimal implementations of ML for complexity reduction is an on-going area of research. Two areas in particular include lattice reduction and spherical decoding. The basic approach with lattice reduction is to perform the initial detection with a basis change of the channel matrix. The basis change allows for a ZF-like receiver structure, but avoiding the issues of channel matrix singularities which degrade a ZF receiver. Assuming the streams are independent after the receiver greatly reduces the number of Euclidean norms which must be computed. Degradation to theoretical ML performance arises from the lattice reduction not resulting in completely independent streams. A lattice reduction technique for MIMO systems first appeared in Yao and Warnell (2002). An extension of the reduction technique is provided in Berenguer *et al.* (2004). A comprehensive survey of closest point searches in lattices is given in Argrell *et al.* (2002).

With spherical decoding, only Euclidean norms which fall within a specified sphere are considered in the search for X_i in Eq. (3.37). As the radius of the sphere is decreased, the complexity of the estimator is reduced. However, if the radius of the sphere is too small, the algorithm could fail to find any point inside the sphere (Hochwald and ten Brink, 2003). Further discussion (including performance curves) of list sphere decoding is given in Hochwald and ten Brink (2003).

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Appendix 3.1: 802.11n channel models

Tables 3.4–3.11 give the parameters for the 802.11n channel models from Erceg et al. (2004). Empty entries in the tables indicate that there is no channel tap at the corresponding time for the particular cluster number.

Table 3.4 Channel model A (Erceg et al., 2004)

Tap index	Cluster 1					
	Excess delay [ns]	Power [dB]	AoA [°]	AS Rx [°]	AoD [°]	AS Tx [°]
1	0	0	45	40	45	40

Table 3.5 Channel model B (Erceg et al., 2004)

Tap index	Excess delay [ns]	Cluster 1					Cluster 2				
		Power [dB]	AoA [°]	AS Rx [°]	AoD [°]	AS Tx [°]	Power [dB]	AoA [°]	AS Rx [°]	AoD [°]	AS Tx [°]
1	0	0	4.3	14.4	225.1	14.4					
2	10	−5.4	4.3	14.4	225.1	14.4					
3	20	−10.8	4.3	14.4	225.1	14.4	−3.2	118.4	25.2	106.5	25.4
4	30	−16.2	4.3	14.4	225.1	14.4	−6.3	118.4	25.2	106.5	25.4
5	40	−21.7	4.3	14.4	225.1	14.4	−9.4	118.4	25.2	106.5	25.4
6	50						−12.5	118.4	25.2	106.5	25.4
7	60						−15.6	118.4	25.2	106.5	25.4
8	70						−18.7	118.4	25.2	106.5	25.4
9	80						−21.8	118.4	25.2	106.5	25.4

Table 3.6 Channel model C (Erceg et al., 2004)

Tap index	Excess delay [ns]	Cluster 1					Cluster 2				
		Power [dB]	AoA [°]	AS Rx [°]	AoD [°]	AS Tx [°]	Power [dB]	AoA [°]	AS Rx [°]	AoD [°]	AS Tx [°]
1	0	0	290.3	24.6	13.5	24.7					
2	10	−2.1	290.3	24.6	13.5	24.7					
3	20	−4.3	290.3	24.6	13.5	24.7					
4	30	−6.5	290.3	24.6	13.5	24.7					
5	40	−8.6	290.3	24.6	13.5	24.7					
6	50	−10.8	290.3	24.6	13.5	24.7					
7	60	−13.0	290.3	24.6	13.5	24.7	−5.0	332.3	22.4	56.4	22.5
8	70	−15.2	290.3	24.6	13.5	24.7	−7.2	332.3	22.4	56.4	22.5
9	80	−17.3	290.3	24.6	13.5	24.7	−9.3	332.3	22.4	56.4	22.5
10	90	−19.5	290.3	24.6	13.5	24.7	−11.5	332.3	22.4	56.4	22.5
11	110						−13.7	332.3	22.4	56.4	22.5
12	140						−15.8	332.3	22.4	56.4	22.5
13	170						−18.0	332.3	22.4	56.4	22.5
14	200						−20.2	332.3	22.4	56.4	22.5

Table 3.7 Channel model D (Erceg *et al.*, 2004)

Tap index	Excess delay [ns]	Cluster 1					Cluster 2					Cluster 3				
		Power [dB]	AoA [°]	AS Rx [°]	AoD [°]	AS Tx [°]	Power [dB]	AoA [°]	AS Rx [°]	AoD [°]	AS Tx [°]	Power [dB]	AoA [°]	AS Rx [°]	AoD [°]	AS Tx [°]
1	0	0	158.9	27.7	332.1	27.4										
2	10	-0.9	158.9	27.7	332.1	27.4										
3	20	-1.7	158.9	27.7	332.1	27.4										
4	30	-2.6	158.9	27.7	332.1	27.4										
5	40	-3.5	158.9	27.7	332.1	27.4										
6	50	-4.3	158.9	27.7	332.1	27.4										
7	60	-5.2	158.9	27.7	332.1	27.4										
8	70	-6.1	158.9	27.7	332.1	27.4										
9	80	-6.9	158.9	27.7	332.1	27.4										
10	90	-7.8	158.9	27.7	332.1	27.4										
11	110	-9.0	158.9	27.7	332.1	27.4										
12	140	-11.1	158.9	27.7	332.1	27.4										
13	170	-13.7	158.9	27.7	332.1	27.4										
14	200	-16.3	158.9	27.7	332.1	27.4										
15	240	-19.3	158.9	27.7	332.1	27.4										
16	290	-23.2	158.9	27.7	332.1	27.4										
17	340															
18	390															

Table 3.8 Channel model E, clusters 1 and 2 (Erceg *et al.*, 2004)

Tap index	Excess delay [ns]	Cluster 1					Cluster 2				
		Power [dB]	AoA [°]	AS Rx [°]	AoD [°]	AS Tx [°]	Power [dB]	AoA [°]	AS Rx [°]	AoD [°]	AS Tx [°]
1	0	-2.6	163.7	35.8	105.6	36.1					
2	10	-3.0	163.7	35.8	105.6	36.1					
3	20	-3.5	163.7	35.8	105.6	36.1					
4	30	-3.9	163.7	35.8	105.6	36.1					
5	50	-4.5	163.7	35.8	105.6	36.1	-1.8	251.8	41.6	293.1	42.5
6	80	-5.6	163.7	35.8	105.6	36.1	-3.2	251.8	41.6	293.1	42.5
7	110	-6.9	163.7	35.8	105.6	36.1	-4.5	251.8	41.6	293.1	42.5
8	140	-8.2	163.7	35.8	105.6	36.1	-5.8	251.8	41.6	293.1	42.5
9	180	-9.8	163.7	35.8	105.6	36.1	-7.1	251.8	41.6	293.1	42.5
10	230	-11.7	163.7	35.8	105.6	36.1	-9.9	251.8	41.6	293.1	42.5
11	280	-13.9	163.7	35.8	105.6	36.1	-10.3	251.8	41.6	293.1	42.5
12	330	-16.1	163.7	35.8	105.6	36.1	-14.3	251.8	41.6	293.1	42.5
13	380	-18.3	163.7	35.8	105.6	36.1	-14.7	251.8	41.6	293.1	42.5
14	430	-20.5	163.7	35.8	105.6	36.1	-18.7	251.8	41.6	293.1	42.5
15	490	-22.9	163.7	35.8	105.6	36.1	-19.9	251.8	41.6	293.1	42.5
16	560						-22.4	251.8	41.6	293.1	42.5

Table 3.9 Channel model E, clusters 3 and 4 (Erceg *et al.*, 2004)

Tap index	Excess delay [ns]	Cluster 3					Cluster 4				
		Power [dB]	AoA [°]	AS Rx [°]	AoD [°]	AS Tx [°]	Power [dB]	AoA [°]	AS Rx [°]	AoD [°]	AS Tx [°]
9	180	-7.9	80.0	37.4	61.9	38.0					
10	230	-9.6	80.0	37.4	61.9	38.0					
11	280	-14.2	80.0	37.4	61.9	38.0					
12	330	-13.8	80.0	37.4	61.9	38.0					
13	380	-18.6	80.0	37.4	61.9	38.0					
14	430	-18.1	80.0	37.4	61.9	38.0					
15	490	-22.8	80.0	37.4	61.9	38.0	-20.6	182.0	40.3	275.7	38.7
16	560						-20.5	182.0	40.3	275.7	38.7
17	640						-20.7	182.0	40.3	275.7	38.7
18	730						-24.6	182.0	40.3	275.7	38.7

Table 3.10 Channel model F, clusters 1, 2, and 3 [Erceg et al., 2004]

Tap index	Excess delay [ns]	Cluster 1					Cluster 2					Cluster 3				
		Power [dB]	AoA [°]	AS Rx [°]	AS Rx [°]	AS Tx [°]	Power [dB]	AoA [°]	AS Rx [°]	AoD [°]	AS Tx [°]	Power [dB]	AoA [°]	AS Rx [°]	AoD [°]	AS Tx [°]
1	0	-3.3	315.1	48.0	56.2	41.6										
2	10	-3.6	315.1	48.0	56.2	41.6										
3	20	-3.9	315.1	48.0	56.2	41.6										
4	30	-4.2	315.1	48.0	56.2	41.6										
5	50	-4.6	315.1	48.0	56.2	41.6										
6	80	-5.3	315.1	48.0	56.2	41.6	-1.8	180.4	55.0	183.7	55.2					
7	110	-6.2	315.1	48.0	56.2	41.6	-2.8	180.4	55.0	183.7	55.2					
8	140	-7.1	315.1	48.0	56.2	41.6	-3.5	180.4	55.0	183.7	55.2					
9	180	-8.2	315.1	48.0	56.2	41.6	-4.4	180.4	55.0	183.7	55.2					
10	230	-9.5	315.1	48.0	56.2	41.6	-5.3	180.4	55.0	183.7	55.2	-5.7	74.7	42.0	153.0	47.4
11	280	-11.0	315.1	48.0	56.2	41.6	-7.4	180.4	55.0	183.7	55.2	-6.7	74.7	42.0	153.0	47.4
12	330	-12.5	315.1	48.0	56.2	41.6	-7.0	180.4	55.0	183.7	55.2	-10.4	74.7	42.0	153.0	47.4
13	400	-14.3	315.1	48.0	56.2	41.6	-10.3	180.4	55.0	183.7	55.2	-9.6	74.7	42.0	153.0	47.4
14	490	-16.7	315.1	48.0	56.2	41.6	-10.4	180.4	55.0	183.7	55.2	-14.1	74.7	42.0	153.0	47.4
15	600	-19.9	315.1	48.0	56.2	41.6	-13.8	180.4	55.0	183.7	55.2	-12.7	74.7	42.0	153.0	47.4
16	730						-15.7	180.4	55.0	183.7	55.2	-18.5	74.7	42.0	153.0	47.4

Table 3.11 Channel model F, clusters 4, 5, and 6 [Erceg et al., 2004]

Tap index	Excess delay [ns]	Cluster 4				Cluster 5				Cluster 6						
		Power [dB]	AoA [°]	AS Rx [°]	AS Tx AoD [°]	Power [dB]	AoA [°]	AS Rx [°]	AS Tx AoD [°]	Power [dB]	AoA [°]	AS Rx [°]	AS Tx AoD [°]			
13	400	-8.8	251.5	28.6	112.5	27.2										
14	490	-13.3	251.5	28.6	112.5	27.2										
15	600	-18.7	251.5	28.6	112.5	27.2	-12.9	68.5	30.7	291.0	33.0					
16	730						-14.2	68.5	30.7	291.0	33.0					
17	880											-16.3	246.2	38.2	62.3	38.0
18	1050											-21.2	246.2	38.2	62.3	38.0

4 PHY interoperability with 11a/g legacy OFDM devices

One of the functional requirements in the development of the 802.11n standard was that some modes of operation must be backward compatible with 802.11a (and 802.11g if 2.4 GHz was supported) as described in Stephens (2005). Furthermore, the 802.11n standard development group also decided that interoperability should occur at the physical layer. This led to the definition of a mandatory mixed format (MF) preamble in 802.11n. In this chapter, we first review the 802.11a packet structure, transmit procedures, and receive procedures to fully understand the issues in creating a preamble that is interoperable between 802.11a and 802.11n devices. For further details regarding 802.11a beyond this review, refer to clause 17 in IEEE (2007a). Following this overview, the mixed format preamble, which is part of 802.11n, is discussed.

4.1 11a packet structure review

The 802.11a packet structure is illustrated in Figure 4.1.

The Short Training field (STF) is used for start-of-packet detection and automatic gain control (AGC) setting. In addition, the STF is also used for initial frequency offset estimation and initial time synchronization. This is followed by the Long Training field (LTF), which is used for channel estimation and for more accurate frequency offset estimation and time synchronization. Following the LTF is the Signal field (SIG), which contains the rate and length information for the packet. Example rates are BPSK, rate $\frac{1}{2}$ encoding and 64-QAM, rate $\frac{3}{4}$ encoding. Following this is the Data field. The first 16 bits of the Data field contain the Service field. An example of an 802.11a transmit waveform is given in Figure 4.2.

4.1.1 Short Training field

The STF is 8 μ s in length. In the time domain, the STF contains ten repetitions of a 0.8 μ s symbol. The STF is defined based on the frequency domain sequence given in Eq. (4.1) (IEEE, 2007a):

$$S_{-26,26} = \sqrt{13/6} \{0, 0, 1+j, 0, 0, 0, -1-j, 0, 0, 0, 1+j, 0, 0, 0, -1-j, 0, 0, 0, -1-j, 0, 0, 0, 1+j, 0, 0, 0, 0, 0, 0, -1-j, 0, 0, 0, -1-j, 0, 0, 0, 1+j, 0, 0, 0, 1+j, 0, 0, 0, 1+j, 0, 0, 0, 1+j, 0\} \quad (4.1)$$

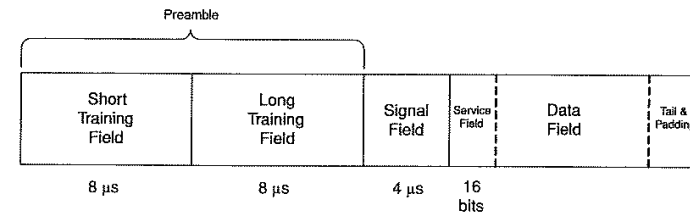


Figure 4.1 802.11a packet structure.

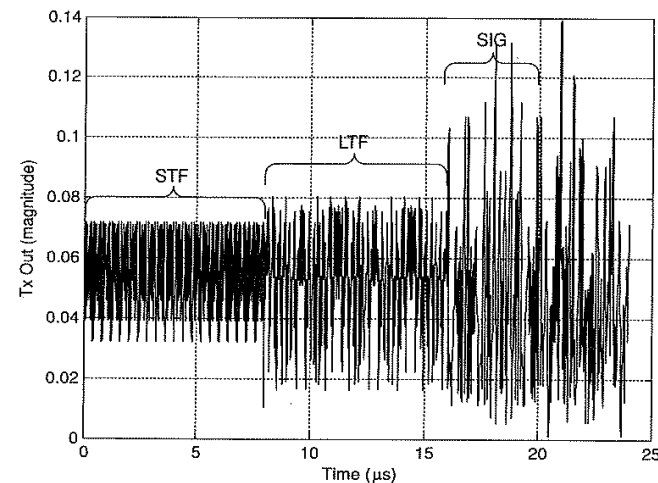


Figure 4.2 802.11a transmit waveform.

The sequence uses 12 of the 52 subcarriers. A 64-point IFFT creates a 3.2 μ s time domain sequence with a pattern which repeats four times (resulting in 0.8 μ s periodicity). This is illustrated in Figure 4.3, with a sampling rate of 20 Msamples/s. This sequence may then be repeated two and half times to create ten short symbol repetitions.

The sequence was chosen to have good correlation properties and a low peak-to-average power so that its properties are preserved even after clipping or compression by an overloaded analog front end. The cross-correlation between the STF and one short training symbol is illustrated in Figure 4.4. There is a separation of over 10 dB between a correlation peak and a correlation sidelobe.

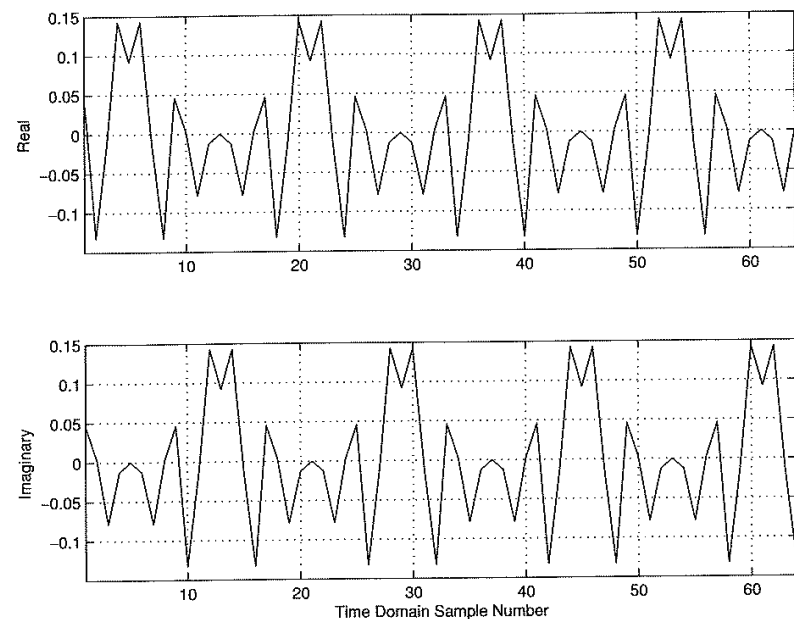


Figure 4.3 Construction of short training symbols.

The correlation peaks may be used to derive an initial time estimate in a time domain based timing acquisition method. It is important to note that the cross-correlation properties of the STF are degraded by delayed responses in the channel. In Figure 4.5, we give an example of the cross-correlation with a channel modeled by two taps. The curve represented by the solid line depicts the channel with the second tap delayed by 50 ns. The curve represented by the dashed line depicts the channel with the second tap delayed by 400 ns. With 50 ns, the cross-correlation peak is a bit wider than the peak illustrated in Figure 4.4. However, with 400 ns, extra correlation peaks arise which may degrade timing acquisition. This issue had a major impact on the design of the MF preamble in 802.11n and is further discussed in Section 4.2.1.

The repetitive nature of the STF can also be used by correlating a $0.8 \mu\text{s}$ short symbol with the previous symbol (commonly termed auto-correlation since the received signal is correlated with a delayed version of itself). This technique may be used for packet detection, whereby the correlation value exceeding a threshold indicates a packet detect. Additionally, this correlation value may be used to set the AGC. This approach proved less sensitive to longer delay spreads as compared to the cross-correlation; however, it is potentially less effective in noisier conditions.

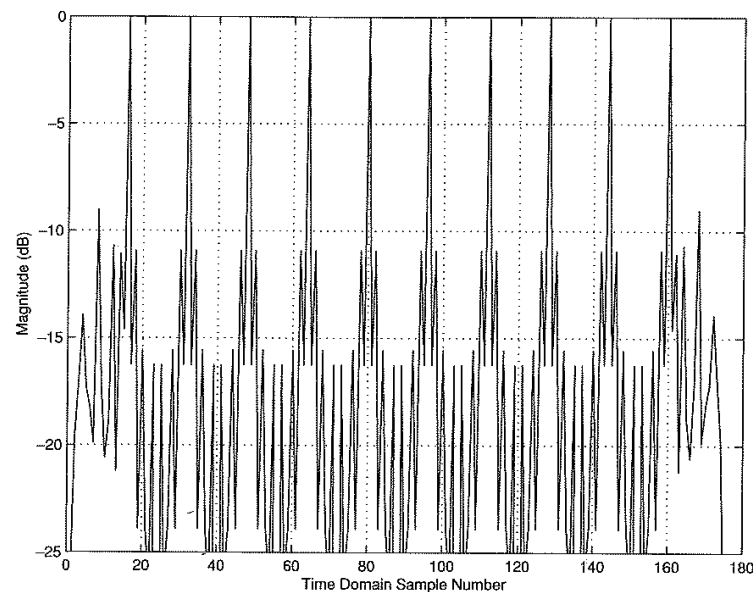


Figure 4.4 Short Training field correlation.

The STF is also used for initial frequency offset estimation. Due to the symbol repetition, the difference in phase between two samples in the STF separated by $0.8 \mu\text{s}$ (or 16 samples if sampling at 20 MHz) allows an estimate of the frequency offset. This operation can resolve a frequency offset between the transmitting device and receiving device of up to $\pm 625 \text{ kHz}$ ($\pm 1/0.8 \mu\text{s}/2$).

4.1.2 Long Training field

The LTF is also $8 \mu\text{s}$ in length, and is composed of two $3.2 \mu\text{s}$ long training symbols prepended by a $1.6 \mu\text{s}$ cyclic prefix. The cyclic prefix is comprised of the second half of the long training symbol. The long training symbol is a 64-point IFFT of the frequency domain sequence given in Eq. (4.2) (IEEE, 2007a). Numbering the subcarrier locations ranging from $-32, -31, \dots, -1, 0, 1, \dots, 31$, the populated subcarriers are located from $-26, -25, \dots, -2, -1, 1, 2, \dots, 25, 26$. All the populated subcarriers have the value of $+1$ or -1 . The subcarrier at DC, L_0 , is not populated.

$$L_{-26,26} = \{1, 1, -1, -1, 1, 1, -1, 1, -1, 1, 1, 1, 1, 1, -1, -1, 1, 1, -1, 1, -1, 1, 1, 1, 0, 1, -1, -1, 1, 1, -1, 1, -1, 1, -1, -1, -1, -1, -1, 1, 1, -1, -1, 1, -1, 1, 1, 1, 1\} \quad (4.2)$$

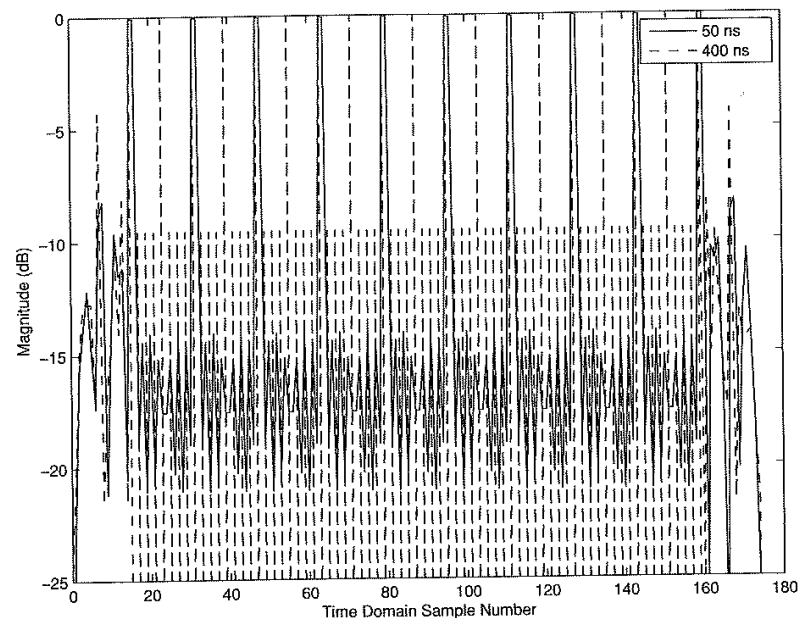


Figure 4.5 STF cross-correlation with a delayed response.

The cross-correlation between the entire preamble and the long training symbol is illustrated in Figure 4.6, with a sampling rate of 20 Msamples/s. The correlation peaks may be used to derive a more accurate time estimate in a time domain based timing acquisition method. Alternatively, the 1.6 μ s cyclic prefix of the LTF may be used for correlation, with either cross-correlation or auto-correlation. Additionally, the transition between the STF and LTF may be used for more accurate time acquisition, e.g. by determining when the STF auto-correlation subsides.

Due to the symbol repetition, the difference in phase between two samples in the LTF separated by 3.2 μ s (or 64 samples if sampling at 20 MHz) allows a more accurate estimate of the carrier frequency offset. This operation can resolve a frequency offset of up to ± 156.25 kHz ($\pm 1/3.2 \mu\text{s}/2$), so it should be used in conjunction with the STF frequency offset estimate.

The other main purpose of the LTF is for channel estimation. The receiver extracts the two long training symbols from the LTF. An FFT is performed on the symbols and the training subcarriers are extracted. The subcarriers from the first long training symbol are averaged with the subcarriers from the second symbol to reduce the effect of noise by 3 dB. Subsequently for each subcarrier k we have the basic communication system

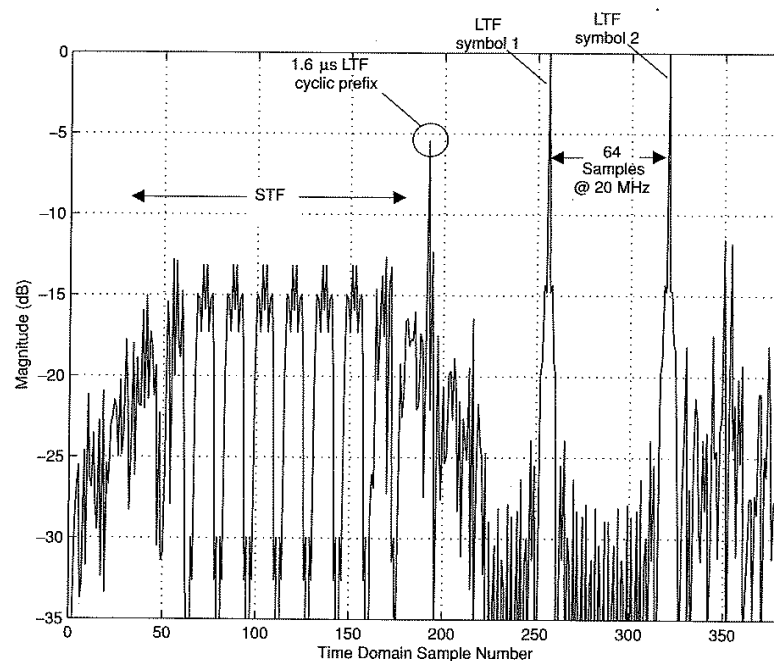


Figure 4.6 Correlation of the preamble with long training symbol.

equation, $y_k = h_k \cdot L_k + z_k$. The values for L_k are given in Eq. (4.2). The known training symbol information is divided out of the received signal leaving the channel estimate for each subcarrier k , $\hat{h}_k = y_k/L_k$.

Since the channel estimate is used to equalize the subsequent OFDM symbols, noise on the channel estimate propagates through the packet during data detection. To reduce the noise on the channel estimate, subcarrier smoothing may be employed. A simple approach is to perform a weighted average of the channel estimate at subcarrier k with its adjacent neighbors, as shown in Eq. (4.3):

$$\tilde{h}_k = \frac{a \cdot \hat{h}_{k-1} + b \cdot \hat{h}_k + a \cdot \hat{h}_{k+1}}{2 \cdot a + b} \quad (4.3)$$

The nature of the channel must be taken into consideration with subcarrier smoothing. In a low delay spread, flat fading channel, channel taps on adjacent subcarriers are highly correlated and subcarrier smoothing provides a significant noise reduction benefit. However, in a highly frequency selective fading channel, adjacent channel taps may not

Table 4.1 Rate information

Rate bit field	Data rate (Mbps)	Modulation	Code rate
1101	6	BPSK	$\frac{1}{2}$
1111	9	BPSK	$\frac{3}{4}$
0101	12	QPSK	$\frac{1}{2}$
0111	18	QPSK	$\frac{3}{4}$
1001	24	16-QAM	$\frac{1}{2}$
1011	36	16-QAM	$\frac{3}{4}$
0001	48	64-QAM	$\frac{2}{3}$
0011	54	64-QAM	$\frac{3}{4}$

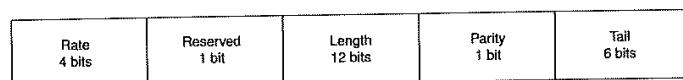


Figure 4.7 Signal field.

be correlated and more sophisticated smoothing algorithms must be applied otherwise improper averaging degrades the channel estimate.

4.1.3 Signal field

The Signal field (SIG) consists of 24 information bits, illustrated in Figure 4.7. The SIG itself is transmitted using BPSK modulation and rate $\frac{1}{2}$ binary convolution code (BCC) (followed by interleaving) to maximize the probability of reception. The BCC is described in Section 4.1.5. Not only is it important for the intended receiver to correctly decode the SIG, but also nearby stations need to correctly decode the SIG in order to properly defer the channel access.

The SIG consists of a single 4 μ s symbol (a 3.2 μ s OFDM symbol prepended by a 0.8 μ s cyclic prefix). The waveform uses 52 subcarriers, as in the LTF. The 48 coded bits are BPSK modulated on to 48 subcarriers. Four additional subcarriers are used as pilots for phase and frequency tracking and training. Numbering the subcarrier locations ranging from $-32, -31, \dots, -1, 0, 1, \dots, 31$, the populated subcarriers are located at $-26, -25, \dots, -2, -1, 1, 2, \dots, 25, 26$. The pilot subcarriers are $-21, -7, 7, 21$. The remaining 48 subcarriers are populated with the coded SIG bits.

The SIG contains packet information to configure the receiver: rate (modulation and coding) and length (amount of data being transmitted in octets). The details of the rate field are given in Table 4.1. The reserved bit was intended for future use. The parity bit gives the even parity over the first 17 bits. The tail bits are set to zero and used to flush the encoder and decoder, since the SIG is separately encoded from the data field.

The parity and reserve bits played a large role in the development of the 11n mixed format (MF) preamble structure. A single parity bit in the SIG proved to be problematic in lower SNR conditions. In very noisy conditions, the probability of a false positive

with a single parity bit approaches 50%. This happens especially when no signal is present at the receiver. A false positive incorrectly indicates a valid SIG, which leads the receiver to use the length and rate values to defer transmission. In a noisy environment or when no packet is present, these fields are random bits, and the device cannot receive or transmit a valid packet for potentially a long period of time. This significantly reduces throughput.

This problem is partially mitigated by first measuring the received signal level and only processing signals which exceed a threshold. However, some manufacturers additionally used the reserve bit as extra parity. To maintain backward compatibility with devices in the field, it was decided not to use the reserve bit in the SIG to indicate a new 802.11n MF packet. This then required an auto-detection between 802.11a and 802.11n MF packets, which is discussed further in Section 4.2.2.1.

4.1.4 Data field

The Data field consists of the Service field, data bits, tail bits, and, if necessary, pad bits. The first 16 bits of the Data field contain the Service field. The first 7 bits of the Service field are the scrambler initialization bits. They are used to synchronize the descrambler and are set to zero to enable estimation of the initial state of the scrambler in the receiver. The remaining 9 bits are reserved and also set to zero.

Following the Service field are the data bits, which in turn are followed by six tail bits which are set to zero. Pad bits (also zeros) are added to the end of the packet so the resulting length fills up the last data symbol.

A length-127 frame synchronous scrambler is used in 802.11a which uses the generator polynomial $G(D) = D^7 + D^4 + 1$. The 127-bit sequence generated repeatedly by the scrambler is (leftmost used first), 00001110 11110010 11001001 00000010 00100110 00101110 10110110 00001100 11010100 11100111 10110100 00101010 11111010 01010001 10111000 11111111, if the all ones initial state is used (IEEE, 2007a). When transmitting, the initial state of the scrambler is actually set to a pseudo random non-zero state for each packet. Scrambling with a different sequence provides peak-to-average protection via retransmission. After the scrambling operation on the Data field, the six tail bits are replaced by six unscrambled zero bits. Pad bits are stripped off at the receiver.

The Data field consists of a stream of symbols, each data symbol 4 μ s in length. A data symbol is comprised of a 3.2 μ s OFDM symbol prepended by a 0.8 μ s cyclic prefix. The waveform uses 52 subcarriers, as in the LTF and SIG. The data bits are encoded and modulated on to 48 subcarriers, as per the rate information in the SIG. Four additional subcarriers are used as pilots for phase and frequency tracking and training. Numbering the subcarrier locations $-32, -31, \dots, -1, 0, 1, \dots, 31$, the populated subcarriers are located at $-26, -25, \dots, -2, -1, 1, 2, \dots, 25, 26$. The pilot subcarriers are located on subcarriers $-21, -7, 7, 21$ with values as follows:

$$P_k = \begin{cases} 1 & k = -21, -7, 7 \\ -1 & k = 21 \\ 0 & \text{otherwise} \end{cases} \quad (4.4)$$

Lastly, each of the OFDM symbols is appended one after another. The digital baseband signal is converted to analog with a DAC, which is then followed by an up-converter to RF and a power amplifier.

Further details on the encoding process are given in IEEE (2007a, Clause 17) (otherwise known as 802.11a), van Nee and Prasad (2000) and Halford (2001). An example of data bits encoded into an 802.11a packet is given in IEEE (2007a, Annex G).

4.1.6 Receive procedure

Typically, the receiver waits for a signal to exceed a threshold before initiating the packet reception procedure. The minimum requirement in 802.11a is to detect the start of a valid transmission received at a power level of at least -82 dBm within $4 \mu\text{s}$ with a probability greater than 90%. Due to competitive pressures, most products exceed this specification by 5–10 dB.

The short training symbols are processed to set the AGC, initial frequency correction, and initial timing acquisition. This is followed by processing the long training symbols for fine frequency correction and fine timing adjustment. The LTF is also used to generate the channel estimate.

The next $4 \mu\text{s}$ symbol is the SIG. The receiver follows the same symbol processing as the data, except for the de-scrambler. The bits in the SIG are not scrambled. Once the SIG bits are extracted, the receiver evaluates the parity bit for errors in the SIG. If the parity indicates an error, the receiver returns to an idle state. Upon no errors being detected by the parity bit, the receiver is configured based on rate and length information. The rate determines the modulation type and coding rate. The length determines the duration of the data stream in octets.

Following the SIG is the Data field, for which a receiver block diagram is illustrated in Figure 4.9. For each $4 \mu\text{s}$ symbol, the $0.8 \mu\text{s}$ GI is removed and an FFT is performed. Using the channel estimate from the LTF, the subcarrier data is equalized as described in Section 3.6. The four pilot subcarriers are used for phase rotation correction, and may be used for further frequency offset and symbol timing correction.

Trellis coded modulation (TCM), which combines modulation and coding, was generally accepted as the best approach for improved performance. However, with the commercial availability and popularity of Viterbi decoders, researchers began investigating schemes for maximizing performance with separate modulation and coding entities to enable reuse of the Viterbi decoder. In Zehavi (1992) it was shown that in a fading channel the code diversity is improved by bit-wise interleaving at the encoder output. The analysis in Claire *et al.* (1998) goes further and demonstrates that on some channels there is a downside to combining demodulation and decoding. For this reason, the 802.11a system is based on bit interleaved coded modulation (BICM). Therefore, the subcarrier data is then demapped by finding bit probabilities from the closest constellation points to the equalized signal. This is followed by the deinterleaver, which is the inverse of the interleaver function. A comprehensive presentation of BICM is given in Caire *et al.* (1998), and further information on low-complexity demapping of BICM is given in van Nee and Prasad (2000) and Akay and Ayanoglu, 2004a).

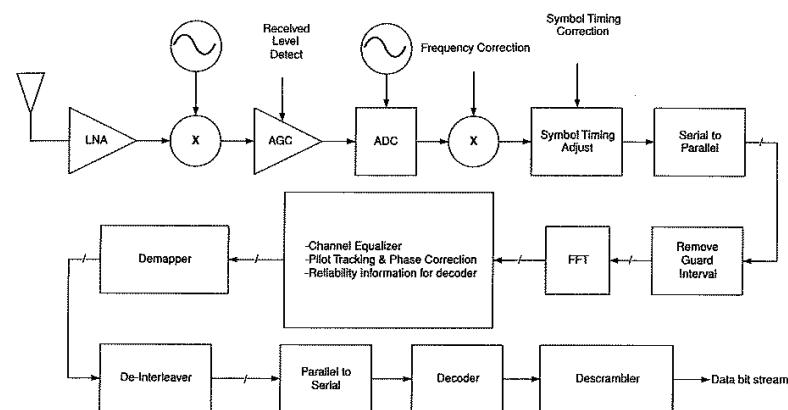


Figure 4.9 Receiver block diagram for Data field.

After the parallel to serial conversion (see Figure 4.9), the data is treated as a stream rather than symbol blocks. In a conventional receiver with ZF or MMSE, the output of the demapper and deinterleaver is passed directly to the decoder, typically a Viterbi decoder. If the code rate is higher than $1/2$, zero metrics are inserted into the decoder in place of the puncture bits.

However, with a frequency selective channel, decoding performance substantially improves by using the channel estimates to weight the decoder metrics (Akay and Ayanoglu, 2004b). The bits associated with subcarriers that are faded are de-weighted (less reliable). This forces the portions of the decoder trellis for which the SNR is low due to the channel fade to have lower branch metrics and have a smaller contribution to the path metric. The subcarriers with high SNR are larger branch metrics and are emphasized in the path metric. Based on the maximum likelihood estimator, we can mathematically demonstrate the benefit of applying channel estimates to weight the decoder metrics.

The maximum likelihood estimation was discussed in Section 3.7 and given in Eq. (3.37). For soft-decision decoding in an AWGN channel with fading, the probability density function of the demodulator output is given by Eq. (4.8) (Proakis, 1989):

$$p(y|x_i) = \frac{1}{\sqrt{2\pi}} \exp \left[-\frac{(y - h \cdot x_i)^2}{2\sigma^2} \right] \quad (4.8)$$

Extrapolating from this, the maximum likelihood estimate of the transmitted data is given by Eq. (4.9) (Akay and Ayanoglu, 2004b):

$$\hat{X} = \min_i \left| \frac{Y - H \cdot X_i}{\sigma} \right|^2 \quad (4.9)$$

where Y is the received signal, σ^2 is the variance of the noise, H and $\hat{H}$ are the channel tap and channel estimate, respectively, and X_i is the set of possible transmitted data. By factoring, we arrive at the expression for the ML estimator given in Eq. (4.10) (Akay and Ayanoglu, 2004b):

$$\hat{X} = \min_i \frac{|\hat{H}|^2}{\sigma^2} \left| \frac{Y}{\hat{H}} - X_i \right|^2 \quad (4.10)$$

The $Y/\hat{H}$ term in Eq. (4.10) is the expression for a ZF receiver. The demapping operation is given by the $|Y/\hat{H} - X_i|$ term. Therefore, to approach ML performance with a ZF receiver, we simply need to include the extra weighting step, given by the $|\hat{H}|^2/\sigma^2$ term. Since with BPSK there is no coupling between bits in the constellation, whose information is lost in a low-complexity BICM demapping operation, Eq. (4.10) provides ML with a ZF receiver.

As illustrated in Figure 4.9, the final step in the receive procedure is descrambling the Data field. The first seven bits of the Service field give the initial state of the descrambler. The same scrambler is used to descramble the data stream. Bits received after the data length indicated by the SIG are discarded.

Further details on the 802.11a packet reception process are given in IEEE (2007a, Clause 17) (otherwise known as 802.11a), van Nee and Prasad (2000), and Halford (2001).

4.2 Mixed format high throughput packet structure

New high throughput (HT) packet formats are used to transmit multiple spatial streams. The mixed format (MF) HT preamble begins with legacy (802.11a) fields to ensure backward compatibility with 802.11a and OFDM 802.11g. These are followed by high throughput training fields. An overview of the fields is given in Figure 4.10. The MF preamble may be used for either 20 MHz or 40 MHz channel width. However, this section focuses on the general functionality of the MF preamble, and only describes the mandatory 20 MHz waveform. The optional 40 MHz waveform is covered in Section 5.1. In addition, other optional features such as sounding packets, short guard interval, STBC, and LDPC and their impact on the waveform design are addressed in Sections 12.6, 5.5, 6.3, and 6.4, respectively.

A plot of a full 802.11n MF waveform with two spatial streams is given in Figure 4.11. The legacy part of the preamble comprises the first 20 μ s. The HT part of the preamble comprises the next 20 μ s, with two spatial streams. This is then followed by the HT data field.

The following sections describe the fields of the MF preamble in detail.

4.2.1 Non-HT portion of the MF preamble

The first part of the MF preamble consists of non-HT (legacy) training fields identical to 802.11a, as illustrated in Figure 4.10. It contains a Non-HT Short Training field (L-STF),

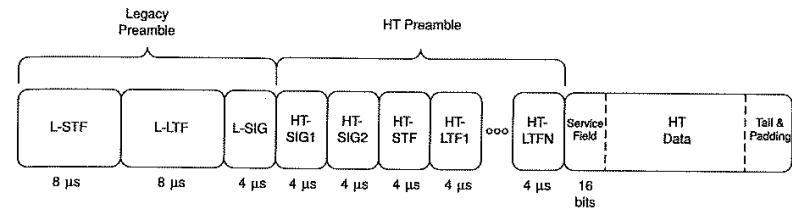


Figure 4.10 Mixed format preamble.

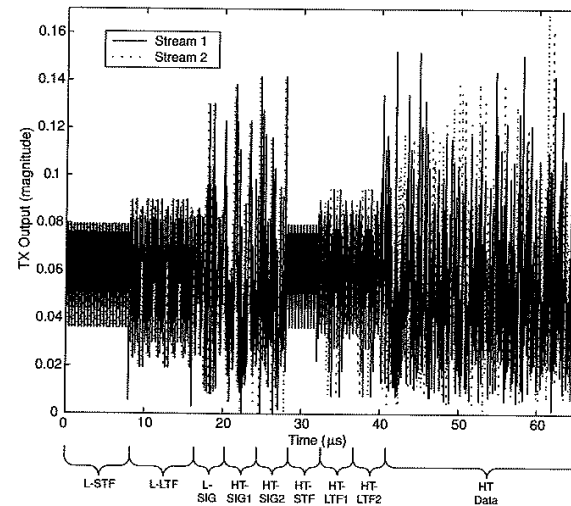


Figure 4.11 802.11n MF with two spatial streams.

non-HT Long Training field (L-LTF), and Non-HT Signal field (L-SIG) such that legacy devices can detect the preamble. However, as discussed below, an 802.11n HT device also needs to use these fields for training and mandatory demodulation of the L-SIG.

Referring to Section 4.1.1, the L-STF is used for start-of-packet detection, AGC setting, initial frequency offset estimation, and initial time synchronization by legacy devices. The L-LTF is used for further fine frequency offset estimation and time synchronization. The L-LTF is also used to generate channel estimates for receiving and equalizing the L-SIG. With proper decoding of the L-SIG, legacy 802.11a/g devices extract the rate and length information and are able to determine the proper time to remain off air. This is discussed in more detail in Section 10.6.9.

Since HT devices perform similar acquisition functions as legacy devices using the L-STF and L-LTF, the issue arises of how to transmit the legacy part of the MF preamble that is a single spatial stream with 802.11n devices that in most cases have multiple transmit antennas. To understand the issue, consider a single antenna legacy device with 17 dBm output transmit power. To maintain similar cost and complexity, a comparable two transmit antenna HT device has two power amplifiers each with 14 dBm output power for the same total output power. If the legacy part of the MF preamble is transmitted from a single antenna of the HT device, the maximum power and range during legacy training is reduced by 3 dB. Moreover, with four antennas, the power on a single transmit RF chain and antenna drops by 6 dB.

To alleviate this transmit power disparity, it is necessary to transmit the legacy training from all antennas of an HT device. However, since the legacy preamble is single stream, transmitting the same signal from each antenna may cause unintentional and undesirable beamforming effects. Since the L-STF is used for AGC setting, large variations in the power of the L-STF also results in saturation or quantization errors.

4.2.1.1 Cyclic shifts

During development of the 802.11n standard, several approaches for how to transmit the legacy training from all antennas were investigated. One approach that was considered was transmitting a subset of subcarriers on each antenna. However, with four transmit antennas, the L-STF, which consists of only twelve active subcarriers, would consist of only three active subcarriers per antenna. This is a very small number of subcarriers per antenna and this approach also suffers from legacy compatibility issues.

Hence, to decorrelate the signals transmitted from different antennas with a single spatial stream and alleviate undesirable beamforming effects, different cyclic shifts (T_{CS}) are applied to each of the signals on the different antennas. Consider an OFDM symbol $s(t)$ over a symbol period T . Replace $s(t)$ with $s(t - T_{CS})$ when $0 \leq t < T + T_{CS}$, and with $s(t - T_{CS} - T)$ when $T + T_{CS} \leq t \leq T$. The cyclic shifted signal is defined as

$$s_{CS}(t; T_{CS})|_{T_{CS} < 0} = \begin{cases} s(t - T_{CS}) & 0 \leq t < T + T_{CS} \\ s(t - T_{CS} - T) & T + T_{CS} \leq t \leq T \end{cases} \quad (4.11)$$

The cyclic shift is applied to each OFDM symbol in the packet separately. Equivalently, in the frequency domain, where $S(f)$ is the Fourier transform of $s(t)$, the cyclic shifted symbol can be described as

$$S_{CS}(f) = S(f)e^{-j2\pi f T_{CS}} \quad (4.12)$$

A two antenna example is given in Figure 4.12, with a -200 ns cyclic shift applied to the signal on the second antenna.

The L-STF is used for AGC setting. Thus, in order to properly adjust the gain setting in the receiver, the power of the STF needs to match the power of the Data field. The ratio between the power of the L-STF and the power of the Data field as measured by

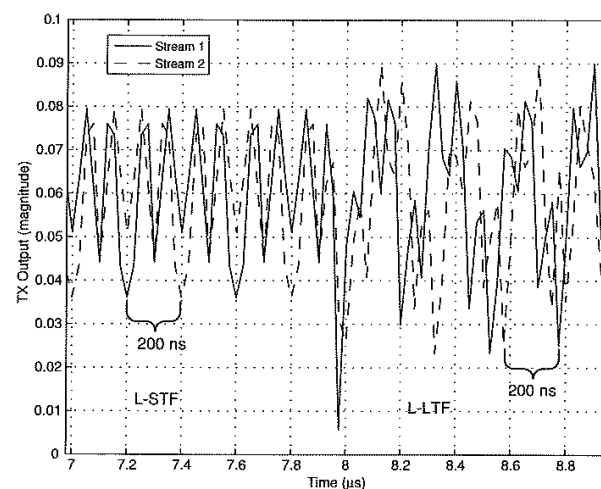


Figure 4.12 Example of cyclic shift applied to second antenna.

the receiver with the propagation modeled by channel model D is given in Figure 4.13. The top graph demonstrates two transmit antennas, and the bottom graph demonstrates four transmit antennas. Several combinations of cyclic shifts are given. For both graphs, the legend gives the cyclic shift on the signals transmitted on each antenna.

With both two and four transmit antennas, having no cyclic shifts on any of the signals results in large variation between the power of the L-STF and Data field, as illustrated by the dash-dotted lines. When the power of the L-STF is smaller than the Data field, the gain is set too high causing saturation or clipping of the data. When the power of the L-STF is larger than the Data field, the gain is set too low causing quantization of the data. With two antennas, the power variation between L-STF and Data field diminishes as the cyclic shift exceeds -200 ns. However with four antennas, cyclic shifts of up to -600 ns are required to minimize variation.

4.2.1.2 Legacy compatibility

Legacy OFDM devices receive the HT packet with MF preamble and process the L-STF, L-LTF, and L-SIG. As discussed in Section 4.1.1 and illustrated in Figure 4.5, the performance of legacy OFDM devices that employ a cross-correlation receiver is degraded by delayed responses. The MF preamble with cyclic shifts applied to the signals on the additional antennas appears to the legacy OFDM devices as a signal convolved with a channel with a delayed response. The larger the cyclic shift, the larger the delay spread, and the more the legacy receiver is degraded.

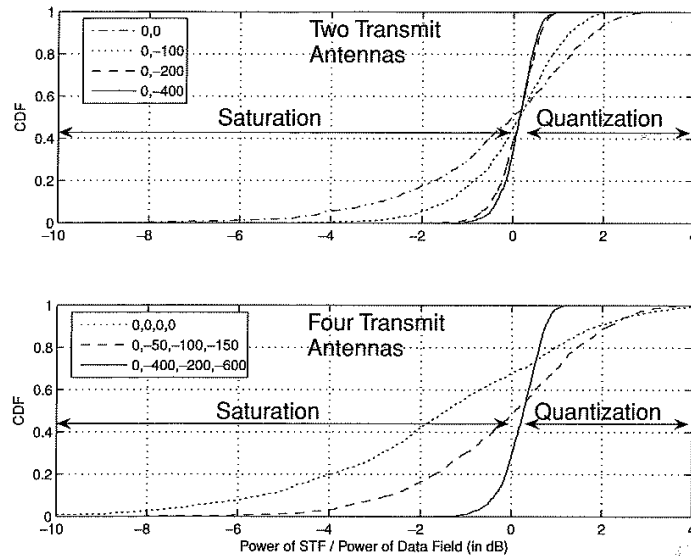


Figure 4.13 Power of the L-STF relative to power of the Data field for channel model D.

During development of the standard, measurements were made by Aoki and Takeda (2005) with 802.11a devices with cross-correlation receivers to test the sensitivity to cyclic shifts. Figure 4.14 illustrates the results of their experiment. Since the legacy device is only able to process up to the L-SIG in the MF preamble, only the SIG error rate was measured. The line marked with diamonds (◆) is the performance of the receiver when receiving an 802.11a transmission, to which we are comparing. The line marked with squares (■) demonstrates the legacy performance when receiving a two antenna transmission with 50 ns cyclic shift on the second antenna. The line marked with triangles (▲) demonstrates the legacy performance when receiving a two antenna transmission with 400 ns cyclic shift on the second antenna. As seen from the results, with 50 ns cyclic shift the SIG error rate is comparable to 802.11a performance. However, with 400 ns cyclic shift the SIG error rate degrades substantially. A compromise was reached to define a maximum magnitude of 200 ns cyclic shift on the non-HT part of the MF preamble.

Table 4.2 gives the values of the cyclic shifts that are applied to the L-STF, L-LTF, L-SIG, and the HT-SIG of the MF preamble.

With more than four transmit antenna chains, specific values for the cyclic shifts on the additional transmit chains are not given. However, the standard states that the shifts on the additional antennas must be between -200 ns and 0 ns.

Table 4.2 Cyclic shift for non-HT portion of the packet (IEEE, 2007b)

Number of Tx chains	Cyclic shift for Tx chain 1 (ns)	Cyclic shift for Tx chain 2 (ns)	Cyclic shift for Tx chain 3 (ns)	Cyclic shift for Tx chain 4 (ns)
1	0	—	—	—
2	0	-200	—	—
3	0	-100	-200	—
4	0	-50	-100	-150

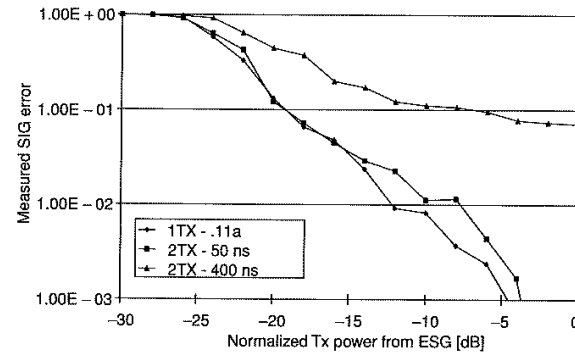


Figure 4.14 Preamble performance with cross-correlation 802.11a receiver in channel model D. Reproduced with permission from Aoki and Takeda (2005) © IEEE.

4.2.1.3 Non-HT Short Training field

In 802.11a, the scaling of the Short Training field was set to make the power of the symbol equal to that of the Long Training field. In 802.11n, the scaling of the symbol is set to make the total power of the time domain signal (summed over all transmit chains) equal to one. Thus, the L-STF frequency domain sequence S_k for the 20 MHz channel is identical to that of 802.11a, given by Eq. (4.1), except as above, the scaling of the sequence for L-STF is $\sqrt{1/2}$ rather than $\sqrt{13/6}$. The time domain L-STF waveform for the 20 MHz channel bandwidth for each transmit chain i_{TX} is defined by the equation below:

$$r_{L-STF}^{(i_{TX})}(t) = \frac{1}{\sqrt{N_{TX} \cdot 12}} \sum_{k=-26}^{26} S_k \exp(j2\pi k \Delta_F (t - T_{CS}^{i_{TX}})) \quad (4.13)$$

where $\Delta_F = 312.5$ kHz (20 MHz/64); $T_{CS}^{i_{TX}}$ represents the cyclic shift for transmit chain i_{TX} with values given by Table 4.2; and N_{TX} is the number of transmit chains.

As described in Section 4.1.1, the L-STF may be created by performing a 64-point IFFT over the frequency domain sequence resulting in a 3.2 μ s time domain sequence, which is repeated two and a half times to create the 8 μ s L-STF waveform. A windowing

function may be applied to the L-STF waveform to smooth transitions between it and the L-LTF (refer to Section 4.1.5, and Eqs. (4.6) and (4.7)).

4.2.1.4 Non-HT Long Training field

The L-LTF frequency domain sequence for the 20 MHz channel is identical to that of 802.11a, given by Eq. (4.2). The time domain L-LTF waveform for the 20 MHz channel bandwidth for each transmit chain i_{TX} is defined by the equation below:

$$r_{L-LTF}^{(i_{TX})}(t) = \frac{1}{\sqrt{N_{TX} \cdot 52}} \sum_{k=-26}^{26} L_k \exp(j2\pi k \Delta_F (t - T_{GI2} - T_{CS}^{i_{TX}})) \quad (4.14)$$

where $\Delta_F = 312.5$ kHz (20 MHz/64); $T_{CS}^{i_{TX}}$ represents the cyclic shift for transmit chain i_{TX} with values given by Table 4.2; T_{GI2} is the 1.6 μ s guard interval; and N_{TX} is the number of transmit chains.

As described in Section 4.1.2, the resulting long training symbol created by performing a 64-point IFFT over the frequency domain sequence is a 3.2 μ s time domain sequence. The 8 μ s L-LTF waveform may be created by repeating the long training symbol twice and prepended by a 1.6 μ s cyclic prefix. A windowing function may be applied to the L-LTF waveform to smooth transitions between it and the L-STF and L-SIG (refer to Section 4.1.5, and Eqs. (4.6) and (4.7)).

4.2.1.5 Non-HT Signal field

The construction of the Non-HT Signal field is identical to the Signal field in 802.11a and described in Section 4.1.3. Although some fields have been co-opted to protect 802.11n from 802.11a devices, e.g. the rate field is always set to 6 Mbps and the length field is set so that 802.11a devices correctly defer transmission for at least the entire duration of the HT packet.

Furthermore, the waveform derived from the 48 complex data d_k , $k = 0, \dots, 47$, may need to be transmitted over multiple antennas. Legacy cyclic shifts are applied to each signal on each transmit antenna chain i_{TX} as described in Section 4.2.1.1. The L-SIG waveform for each transmit antenna chain i_{TX} is given below:

$$r_{L-SIG}^{(i_{TX})}(t) = \frac{1}{\sqrt{N_{TX} \cdot 52}} \sum_{k=-26}^{26} (D_k + p_0 P_k) \exp(j2\pi k \Delta_F (t - T_{GI} - T_{CS}^{i_{TX}})) \quad (4.15)$$

where D_k and $M'(k)$, given by

$$D_k = \begin{cases} 0 & k = 0, \pm 7, \pm 21 \\ d_{M'(k)} & \text{otherwise} \end{cases}$$

$$M'(k) = \begin{cases} k + 26 & -26 \leq k \leq -22 \\ k + 25 & -20 \leq k \leq -8 \\ k + 24 & -6 \leq k \leq -1 \\ k + 23 & 1 \leq k \leq 6 \\ k + 22 & 8 \leq k \leq 20 \\ k + 21 & 22 \leq k \leq 26 \end{cases}$$

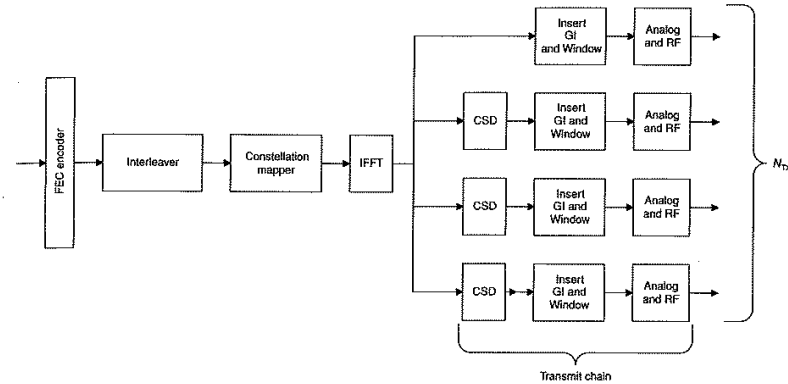


Figure 4.15 Transmitter block diagram for L-SIG and HT-SIG of the HT mixed format packet. Reproduced with permission from IEEE (2007b) © IEEE.

are the mapping of complex data values to subcarrier locations while leaving space for pilots, DC, and guard band; $T_{CS}^{i_{TX}}$ represents the non-HT cyclic shift for transmit chain i_{TX} with values given by Table 4.2; T_{GI} is the 0.8 μ s guard interval; P_k is the same as legacy OFDM and is given by Eq. (4.4); and p_0 is the same as legacy OFDM and is given by Eq. (4.5).

As described in Section 4.1.3, the resulting waveform is a 3.2 μ s time domain OFDM symbol. A 0.8 μ s cyclic prefix is prepended, resulting in a 4.0 μ s symbol. A windowing function may be applied to the L-SIG waveform to smooth transitions between it and the L-LTF and HT-SIG (refer to Section 4.1.5, and Eqs. (4.6) and (4.7)). The transmitter block diagram for L-SIG is illustrated in Figure 4.15.

4.2.2 HT portion of the MF preamble

As illustrated in Figure 4.10, the HT portion of the MF preamble consists of the High Throughput Signal field (HT-SIG), High Throughput Short Training field (HT-STF), and High Throughput Long Training field (HT-LTF). The two OFDM symbols of the HT-SIG, HT-SIG₁ and HT-SIG₂, contain new signaling information and are also used for auto-detection between HT MF and legacy OFDM packets. The HT-STF is used to redo the AGC setting. The HT-LTFs are used for MIMO channel estimation. The HT Data field follows the HT-LTFs.

4.2.2.1 High Throughput Signal field

The HT-SIG contains 48 information bits, divided into two symbols, HT-SIG₁ and HT-SIG₂, each containing 24 information bits. We require 32 bits to signal the new HT

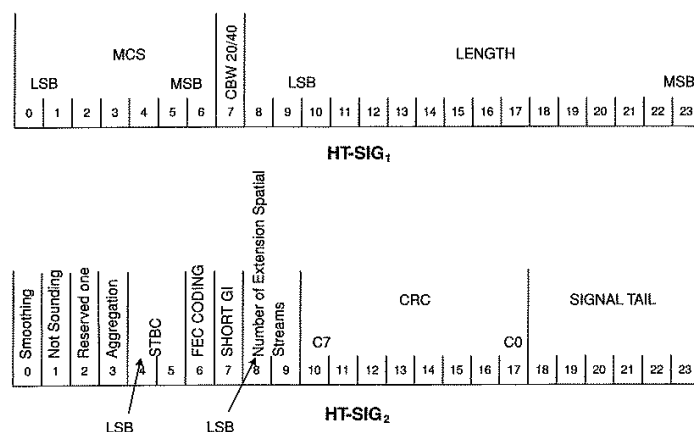


Figure 4.16 Format of HT-SIG₁ and HT-SIG₂. Reproduced with permission from IEEE (2007b)
© IEEE.

information and modes. In addition, the HT-SIG includes an 8-bit CRC for strong error detection, as opposed to the single parity in the L-SIG. As in the L-SIG, the last 6 bits are tail bits (set to zero) to terminate the trellis of the convolution code since the HT-SIG is also encoded with a rate $\frac{1}{2}$ BCC. The HT-SIG is illustrated in Figure 4.16.

In HT-SIG1, the first 7 bits are used to signal which of the modulation coding schemes (MCS) 0 to 76 is used for transmission of the Data field. The MCSs consist of BPSK, QPSK, 16-QAM, and 64-QAM, identical to 802.11a. The coding rates include 1/2, 2/3, and 3/4 the same as 802.11a, and a new rate 5/6. MCS 0 to 7 consist of single stream, MCS 8 to 15 consist of two streams, MCS 16 to 23 consist of three streams, and MCS 24 to 31 consist of four streams. MCS 32 to 31 have equal modulation on each stream. MCS 32 only applies to 40 MHz operation and is discussed in Chapter 5. MCS 33 to 76 have unequal modulation on each stream and are discussed in Chapter 12. The MCS parameters and data rates for 20 MHz, equal modulation, and 800 ns guard interval are defined in Tables 4.7–4.10. Values 77 to 127 are reserved.

The CBW 20/40 bit is used to indicate whether the transmission has 20 MHz or 40 MHz bandwidth. This information is useful to allow a 20 MHz receiver to determine that it is not capable of receiving the transmission and to terminate reception early, reducing power consumption. Sixteen bits are used for the Length field to allow for long packets in the range of 0 to 65 535 bytes. Longer packets than 802.11a are needed to improve efficiency at higher data rates.

In HT-SIG₂, the Smoothing bit, Not Sounding bit, and Number of Extension Spatial Streams bits are all related to Tx beamforming. With Tx beamforming or spatial expansion, the resulting channel between the transmitter and receiver could exceed 800 ns.

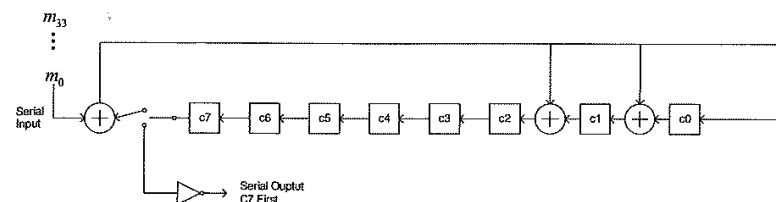


Figure 4.17 HT-SIG CRC calculation. Reproduced with permission from IEEE (2007b) © IEEE.

High delay spread results in the decorrelation of adjacent subcarriers. In addition, with certain methods of Tx beamforming, a phase discontinuity could exist between adjacent subcarriers. Both of these conditions degrade the channel estimate when subcarrier smoothing is used (refer to Section 4.1.2 for a discussion on subcarrier smoothing). Under these conditions, the transmitter should set the Smoothing bit to 0 to inform the receiver that only per-carrier (unsmoothed) channel estimation should be employed.

When the Not Sounding bit is set to 0, the transmission is a sounding packet. A sounding packet is used to gather channel state information for Tx beamforming and link adaptation. To sound extra spatial dimensions, beyond that of the Data field, the Number of Extension Spatial Streams field is set to greater than zero. Further detail on sounding packets is found in Section 12.6.

The aggregate bit indicates whether the payload contains a single MPDU (aggregate bit set to 0) or an aggregate of MPDUs (A-MPDU) (aggregate bit set to 1). An A-MPDU contains delimiters between the individual MPDUs. The PHY reports to the MAC that the packet contains an A-MPDU immediately after decoding the HT-SIG. The MAC requires this information in order to extract the individual MPDUs; refer to Section 8.2.2.

The two STBC bits indicate the dimensionality of the STBC operation. When the bits have a value of 0, there is no STBC. The value of 3 is reserved. Other values are discussed in detail in Section 6.3.

The remaining three informational bits include a bit for FEC Coding that is set to 0 to indicate that the data is BCC encoded, or set to 1 to indicate LDPC encoding. The other bit is the Short GI bit, which is set to 1 to indicate that a 400 ns short guard interval is used, rather than the standard 800 ns length guard interval. These features are presented in Chapter 6. The last bit is a reserved bit that is always set to 1.

The CRC protects bits 0 to 33 of the HT-SIG (bits 0 to 23 of HT-SIG₁ and bits 0 to 9 of HT-SIG₂) (IEEE, 2007b). The generator polynomial is $G(D) = D^8 + D^2 + D + 1$. Figure 4.17 shows the operation of the CRC. First, the shift register is reset to all ones. The HT-SIG bits are then passed sequentially through the exclusive-or operation (XOR) at the input. When the last bit has entered, the output is generated by shifting the bits out of the shift register, c7 first, through an inverter.

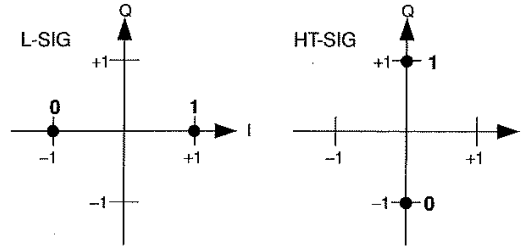


Figure 4.18 Data subcarrier constellations of L-SIG and HT-SIG. Reproduced with permission from IEEE (2007b) © IEEE.

As an example, if bits $\{m_0 \dots m_{33}\}$ are given by

$$\{1111000100100110000000001110000000\}$$

the resulting CRC bits $\{c_7 \dots c_0\}$ are $\{10101000\}$.

The 48 HT-SIG bits are BCC encoded with rate $1/2$ resulting in 96 bits. These are split into two symbols. Each symbol is interleaved, BPSK mapped, and has pilot subcarriers inserted as described in Section 4.1.3. The complex numbers generated by these steps are denoted as d_k , $k = 0, \dots, 47$. The resulting time domain waveform for the two HT-SIG OFDM symbols is given in Eq. (4.16). Cyclic shifts are applied to the additional transmit antennas as with the L-SIG. To complete the waveform, a cyclic prefix is prepended to each OFDM symbol.

$$r_{\text{HT-SIG}}^{i_{\text{TX}}}(t) = \frac{1}{\sqrt{N_{\text{TX}} \cdot 52}} \sum_{n=0}^1 \sum_{k=-26}^{26} (j D_{k,n} + p_{n+1} P_k) \times \exp(j 2\pi k \Delta_F (t - n T_{\text{SYM}} - T_{\text{GI}} - T_{\text{CS}}^{i_{\text{TX}}})) \quad (4.16)$$

where D_k and $M'(k)$ are the mapping of complex data values to subcarrier locations, as for L-SIG defined by Eq. (4.15); $T_{\text{SYM}} = 4 \mu\text{s}$ symbol time composed of $3.2 \mu\text{s}$ OFDM symbol time and $0.8 \mu\text{s}$ guard interval; $T_{\text{CS}}^{i_{\text{TX}}}$ is the non-HT cyclic shift for transmit chain i_{TX} with values given by Table 4.2; $T_{\text{GI}} = 0.8 \mu\text{s}$ guard interval; P_k denotes the pilot subcarriers defined to be the same as legacy OFDM and given by Eq. (4.4); and p_{n+1} is the pilot polarity defined to be the same as legacy OFDM and given by Eq. (4.5) – since the first value of the sequence is used for L-SIG, HT-SIG uses the next two values.

A windowing function may be applied to each of the HT-SIG symbols to smooth transitions between it and adjacent symbols (refer to Section 4.1.5, and Eqs. (4.6) and (4.7)). The transmitter block diagram for the HT-SIG symbols is illustrated in Figure 4.15.

Note that in Eq. (4.16) the data D_k are rotated by 90 degrees (multiplication by j). Since both the L-SIG and HT-SIG are BPSK modulated, the HT-SIG is rotated by 90 degrees relative to L-SIG, as illustrated in Figure 4.18.

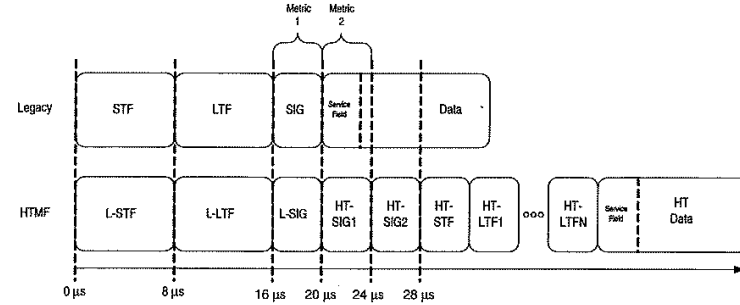


Figure 4.19 Timing of the preamble fields in legacy and HT MF.

The rotation allows for robust auto-detection between L-SIG and HT-SIG₁ to differentiate between legacy OFDM packets and HT MF packets.

An example auto-detection algorithm is as follows. For each symbol that is processed, compute a metric based on the power of the real part of the data subcarriers averaged over the 48 data subcarriers minus the power of the imaginary part of the data subcarriers averaged over the 48 data subcarriers. Figure 4.19 illustrates the relative timing of the preamble fields between legacy and HT MF. During the time spanning $16 \mu\text{s}$ to $20 \mu\text{s}$, the symbol corresponds to the SIG in a legacy OFDM packet or the L-SIG in the HT MF packet. For both packet formats, the metric is a large positive number. During the time spanning $20 \mu\text{s}$ to $24 \mu\text{s}$, the symbol corresponds to the first data symbol in a legacy OFDM packet or HT-SIG₁ in the HT MF packet. For the first data symbol in a legacy OFDM packet, the metric ranges from large positive with BPSK modulation to 0 for QAM modulation. (Moreover, in the case of QAM modulation the packet cannot be HT, yet due to pipelining this knowledge is not yet available.) For the HT-SIG₁ of the HT MF packet, the metric is a large negative number. Finally, the metric from the time spanning $16 \mu\text{s}$ to $20 \mu\text{s}$ is compared with the metric from the time spanning $20 \mu\text{s}$ to $24 \mu\text{s}$. With an appropriate threshold, we can differentiate between a legacy OFDM packet and an HT MF packet.

4.2.2.2 High Throughput Short Training field

As was discussed in Section 4.2.1.1, large cyclic shifts are required to decorrelate the signals when the same spatial stream is sent on the different transmit antenna chains. This was illustrated in Figure 4.13, where, with four transmit antennas, cyclic shifts with a magnitude of 600 ns are required for the STF to have comparable power to the Data field. However, for backward compatibility with legacy OFDM devices, a cyclic shift with only a maximum magnitude of 200 ns is needed for L-STF as discussed in Section 4.2.1.2. Therefore, an additional field is added after HT-SIG to enable accurate power estimates for AGC (automatic gain control) setting required for MIMO. The HT-STF allows for a more accurate AGC setting than L-STF and reduces the dynamic range

Table 4.3 Cyclic shift for values for HT-STF, HT-LTF, and Data field (IEEE, 2007b)

Number of spatial streams	Cyclic shift for spatial stream 1 (ns)	Cyclic shift for spatial stream 2 (ns)	Cyclic shift for spatial stream 3 (ns)	Cyclic shift for spatial stream 4 (ns)
1	0	—	—	—
2	0	−400	—	—
3	0	−400	−200	—
4	0	−400	−200	−600

requirement on the ADC. The values of the cyclic shifts for different numbers of spatial streams are given in Table 4.3.

The time domain HT-STF waveform for 20 MHz channel bandwidth for each spatial stream i_{SS} is defined by Eq. (4.17). The subcarrier sequence S_k for the HT-STF is identical to that of the L-STF, defined in Eq. (4.13). For simplicity, we assume that the number of spatial streams is equal to the number of transmit antennas (direct mapping). Spatial mapping, which maps fewer spatial streams to more transmit antennas, is discussed in Section 4.2.3.8.

$$r_{HT-STF}^{i_{SS}}(t) = \frac{1}{\sqrt{N_{SS} \cdot 12}} \sum_{k=-26}^{26} S_k \exp(j 2\pi k \Delta_F (t - T_{CS}^{i_{SS}})) \quad (4.17)$$

where $\Delta_F = 312.5$ kHz (20 MHz/64); $T_{CS}^{i_{SS}}$ represents the cyclic shift for spatial stream i_{SS} with values given by Table 4.3; and N_{SS} is the number of spatial streams.

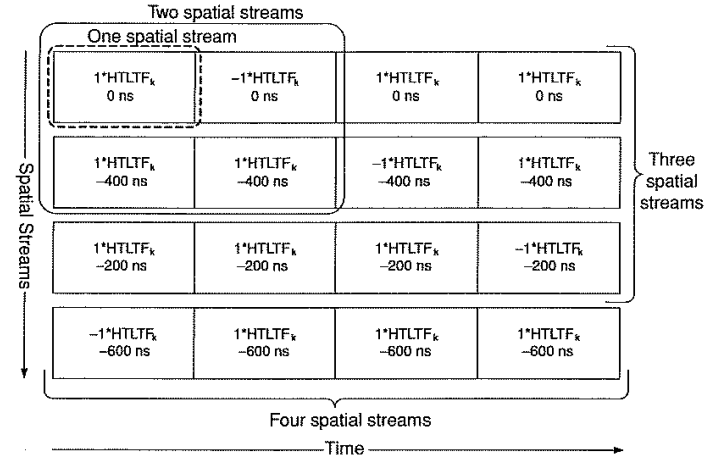
The HT-STF symbol is 4.0 μ s in length. A windowing function may be applied to the HT-STF symbol to smooth transitions between it and adjacent symbols (refer to Section 4.1.5, and Eqs. (4.6) and (4.7)).

4.2.2.3 High Throughput Long Training field

A new HT Long Training field is provided in the HT MF preamble for MIMO channel estimation. A 4 μ s HT long training symbol is transmitted for each spatial stream indicated by the selected MCS, except in the case of three spatial streams where four long training symbols are transmitted. An orthogonal mapping matrix is used to generate the HT-LTF. In order to permit non-complex mapping matrix for three spatial streams, the extra long training symbol was added. The same cyclic shifts as used for HT-STF, defined in Table 4.3, are applied to the long training symbols to accurately match the power variation of the Data field.

The HT-LTF is illustrated in Figure 4.20. The HT long training frequency sequence is multiplied by a value (+1 or −1) from the orthogonal mapping matrix. Each long training symbol is cyclic shifted with values given in Table 4.3.

The HT long training symbol for 20 MHz operation is defined based on the frequency domain sequence given in Eq. (4.18) (IEEE, 2007b). This sequence is an extension of the legacy long training symbol (defined by Eq. (4.2)), where two additional subcarriers,

**Figure 4.20** Construction of the HT-LTF.

−28 and −27, are filled with +1 and two additional subcarriers, 27 and 28, are filled with −1:

$$HTLTF_{-28:28} = \begin{bmatrix} 1, 1, 1, 1, -1, -1, 1, 1, -1, 1, -1, 1, 1, 1, 1, 1, -1, -1, 1, 1, -1, 1, -1, \\ 1, 1, 1, 1, 0, 1, -1, -1, 1, 1, -1, 1, -1, 1, -1, -1, -1, -1, -1, 1, 1, -1, \\ -1, 1, -1, 1, -1, 1, 1, 1, 1, -1, -1 \end{bmatrix} \quad (4.18)$$

The time domain HT-LTF waveform for 20 MHz channel bandwidth for each spatial stream i_{SS} and for each long training symbol n , $1 \leq n \leq N_{LTF}$, is defined by Eq. (4.19). For simplicity, we assume that the number of spatial streams is equal to the number of transmit antennas (direct mapping). Spatial mapping, which maps fewer spatial streams to more transmit antennas, is discussed in Section 4.2.3.8.

$$r_{HT-LTF}^{n, i_{SS}}(t) = \sqrt{\frac{1}{N_{SS} \cdot 56}} \cdot \sum_{k=-28}^{28} [P_{HTLTF}]_{i_{SS}, n} HTLTF_k \times \exp(j 2\pi k \Delta_F (t - T_{GI} - T_{CS}^{i_{SS}})) \quad (4.19)$$

where T_{GI} is the 0.8 μ s guard interval; $\Delta_F = 312.5$ kHz (20 MHz/64); $T_{CS}^{i_{SS}}$ represents the cyclic shift for spatial stream i_{SS} with values given by Table 4.3; and N_{SS} is the number of spatial streams.

The orthogonal mapping matrix is given in Eq. (4.20) (IEEE, 2007b):

$$P_{HTLTF} = \begin{pmatrix} 1 & -1 & 1 & 1 \\ 1 & 1 & -1 & 1 \\ 1 & 1 & 1 & -1 \\ -1 & 1 & 1 & 1 \end{pmatrix} \quad (4.20)$$

The number of long training symbols is given in Eq. (4.21):

$$N_{LTF} = \begin{cases} N_{SS} & \text{if } N_{SS} = 1, 2, 4 \\ 4 & \text{if } N_{SS} = 3 \end{cases} \quad (4.21)$$

To complete the waveform, a $0.8 \mu\text{s}$ cyclic prefix is prepended to each $3.2 \mu\text{s}$ OFDM symbol. A windowing function may be applied to each long training symbol to smooth transitions between it and adjacent symbols (refer to Section 4.1.5, and Eqs. (4.6) and (4.7)).

4.2.3 Data field

Many of the steps in the encoding process of the Data field are the same as 802.11a, a description of which will not be repeated. For an explanation of the 802.11a functions, refer to Section 4.1.5. The transmitter block diagram for the HT Data field is given in Figure 4.21.

The following sections describe the encoding process when using the mandatory binary convolutional code (BCC). Refer to Section 6.4 for the encoding steps when using the optional low density parity check (LDPC) code. In addition, refer to Section 6.3 for the waveform design when using the optional space-time block code (STBC).

4.2.3.1 Bit string

As in 802.11a, the Service field consisting of 16 zero bits is prepended to the data bits. With BCC, six (all zeros) tail bits are appended to the data bits. If necessary, the bit string is extended with pad bits (all zeros) such that the resulting length can be divided equally into an integer multiple of OFDM symbols. The encoding process for the optional LDPC code is described in Section 6.4.

The number of OFDM symbols in the Data field is as follows:

$$N_{SYM} = \left\lceil \frac{8 \cdot \text{length} + 16 + 6}{N_{DBPS}} \right\rceil \quad (4.22)$$

where length is the value of the Length field in octets in the HT-SIG field illustrated in Figure 4.16; and N_{DBPS} is the number of data bits per OFDM symbol determined by the selected MCS.

The symbol $\lceil x \rceil$ denotes the smallest integer greater than or equal to x , corresponding to the ceiling function. The number of pad bits is therefore $N_{SYM} \cdot N_{DBPS} - 8 \cdot \text{length} - 16 - 6$.

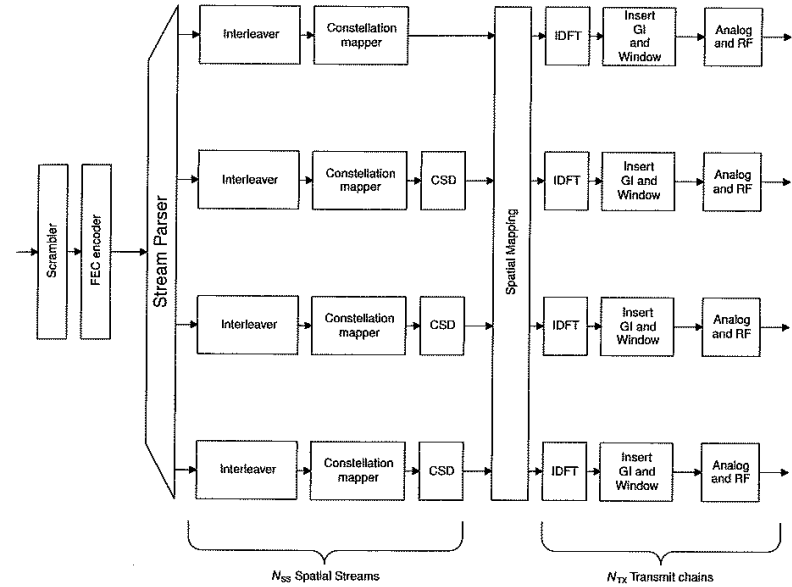


Figure 4.21 Transmitter block diagram for HT Data field. Reproduced with permission from IEEE (2007b) © IEEE.

4.2.3.2 Scrambling and encoding

The resulting data bit string is scrambled with the same scrambler as in 802.11a, described in Section 4.1.4. The six scrambled zero bits that served as tail bits following the data are replaced with the same number of non-scrambled zero bits. This returns the convolutional encoder to the zero state, and is described further in IEEE (2007a, Clause 17.3.5.2).

The scrambled bit string is encoded with the rate $\frac{1}{2}$ BCC. After encoding, the bit string is punctured to the selected data rate. Both encoding and puncturing are the same as in 802.11a and are described in Section 4.1.5.

4.2.3.3 Stream parsing

After encoding and puncturing, the bit string is re-arranged into a new set of bit strings equal to the number of spatial streams (N_{SS}). To simplify the description, this section only describes the stream parsing function for equal MCSs. For unequal MCSs, refer to Section 12.4. The output of the stream parser is N_{SS} bit strings, each of length N_{CBPS} (coded bits per spatial stream).

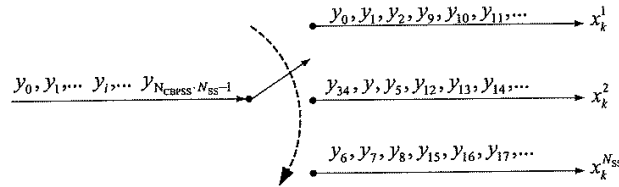


Figure 4.22 Stream parser for single encoder and equal MCS.

A block of s bits is alternately assigned to different spatial streams in a round robin fashion. Figure 4.22 illustrates an example of 64-QAM ($s = 3$) and three spatial streams.

The size of a block of s bits is determined from a single axis (real or imaginary) of a constellation point, defined in the equation below:

$$s = \max \left\{ 1, \frac{N_{BPSCS}}{2} \right\} \quad (4.23)$$

where N_{BPSCS} is the number of coded bits per single subcarrier for each spatial stream, or equivalently the modulation order. The output of the stream parser, x_k^{iss} , is a function of the input, y_i , and is defined as follows:

$$x_k^{iss} = y_i \quad (4.24)$$

$$i = (i_{SS} - 1) \cdot s + s \cdot N_{SS} \cdot \left\lfloor \frac{k}{s} \right\rfloor + k \bmod s$$

where $1 \leq i_{SS} \leq N_{SS}$ and $k = 0, 1, \dots, N_{CBPSS} - 1$.

The symbol $\lfloor x \rfloor$ denotes the largest integer less than or equal to x , corresponding to the floor function. $k \bmod s$ is defined as the remainder from the division of the integer k by the integer s .

4.2.3.4 Interleaving

After the stream parser, each of the N_{SS} bit strings is interleaved using three permutations if BCC encoding was used. The first two interleaver permutations are based on the interleaver for 802.11a. The first permutation ensures that adjacent coded bits are mapped onto non-adjacent subcarriers (IEEE, 2007a). The second permutation ensures that adjacent coded bits are mapped alternately onto less and more significant bits of the constellation, avoiding long runs of low reliability bits (IEEE, 2007a). If more than one spatial stream exists, the third operation is applied, which performs a frequency rotation to the additional spatial streams.

In this section, we limit the description of the interleaver to 20 MHz operation and equal MCS to simplify the explanation. Interleaver design for 40 MHz is addressed in Section 5.1.4 and unequal MCS is addressed in Section 12.4.

The output of the first interleaver permutation, w_k^{iss} , is a function of the output of the stream parser, x_i^{iss} , and is defined as follows:

$$w_k^{iss} = x_i^{iss} \quad (4.25)$$

$$i = 4 \cdot N_{BPSCS} \cdot (k \bmod 13) + \left\lfloor \frac{k}{13} \right\rfloor$$

where $1 \leq i_{SS} \leq N_{SS}$, $k = 0, 1, \dots, N_{CBPSS} - 1$.

Comparing the first permutation to that of 802.11a, the modulus and floor function use a factor of 16 because 802.11a consists of 48 data subcarriers, which is divisible by 16. In 802.11n 20 MHz operation, there are 52 data subcarriers, which is divisible by 13.

The output of the second interleaver permutation, y_k^{iss} , is a function of the output of the first interleaver permutation, w_i^{iss} , and is defined as follows:

$$y_k^{iss} = w_j^{iss} \quad (4.26)$$

$$j = s \cdot \left\lfloor \frac{k}{s} \right\rfloor + \left(k + N_{CBPSS} - \left\lfloor \frac{13 \cdot k}{N_{CBPSS}} \right\rfloor \right) \bmod s$$

where $1 \leq i_{SS} \leq N_{SS}$, $k = 0, 1, \dots, N_{CBPSS} - 1$, and s is the same block size as with the stream parser, defined by Eq. (4.23).

If there are more than one spatial stream, the third permutation is performed. The output of the third interleaver permutation, z_k^{iss} , is a function of the output of the second interleaver permutation, y_r^{iss} , and is defined as follows:

$$z_k^{iss} = y_r^{iss} \quad (4.27)$$

$$r = \left(k - \left(((i_{SS} - 1) \cdot 2) \bmod 3 + 3 \cdot \left\lfloor \frac{i_{SS} - 1}{3} \right\rfloor \right) \cdot 11 \cdot N_{BPSCS} \right) \bmod N_{CBPSS}$$

where $1 \leq i_{SS} \leq N_{SS}$, and $k = 0, 1, \dots, N_{CBPSS} - 1$.

4.2.3.5 Modulation mapping

The subcarriers in 802.11n are BPSK, QPSK, 16-QAM, or 64-QAM modulated based on the selected MCS in the same method as in 802.11a (refer to IEEE (2007a, Clause 17.3.5.7) or van Nee and Prasad (2000) for further details). The bit streams at the output of the interleaver are accordingly mapped to complex constellation points, with the notation shown in Eq. (4.28) for the data streams. Each 20 MHz HT OFDM symbol consists of 52 complex data values for each stream indexed by k .

$$d_{k,l,n}, \quad k = 0, 1, \dots, 51, \quad l = 1, \dots, N_{SS}, \quad n = 0, 1, \dots, N_{SYM} - 1 \quad (4.28)$$

where N_{SYM} is the number of OFDM symbols defined in Eq. (4.22), and N_{SS} is the number of spatial streams.

However, if such a signal was transmitted, simultaneous transmission of the same signals has undesired beamforming effects. To decorrelate the signals we multiply the signals by a cyclic shift. An example is given below:

$$\frac{1}{\sqrt{2}} \begin{bmatrix} d_1 & \times \exp(-j 2\pi k \Delta_F (0 \text{ ns})) \\ d_2 \exp(-j 2\pi k \Delta_F (-400 \text{ ns})) & \times \exp(-j 2\pi k \Delta_F (0 \text{ ns})) \\ d_1 & \times \exp(-j 2\pi k \Delta_F (-200 \text{ ns})) \\ d_2 \exp(-j 2\pi k \Delta_F (-400 \text{ ns})) & \times \exp(-j 2\pi k \Delta_F (-200 \text{ ns})) \end{bmatrix}$$

$$= \frac{1}{\sqrt{2}} \begin{bmatrix} d_1 \exp(-j 2\pi k \Delta_F (0 \text{ ns})) \\ d_2 \exp(-j 2\pi k \Delta_F (-400 \text{ ns})) \\ d_1 \exp(-j 2\pi k \Delta_F (-200 \text{ ns})) \\ d_2 \exp(-j 2\pi k \Delta_F (-600 \text{ ns})) \end{bmatrix}$$

With this selection of cyclic shifts we only attain a 200 ns separation between copies of the same signal. One might be tempted to use a set of cyclic shifts consisting of [0 ns; 0 ns; -400 ns; +400 ns] which would result in an overall cyclic shift on the data of [0 ns; -400 ns; 0 ns; -400 ns] separating the same signals by 400 ns. However, in order for the HT-STF and the HT-LTF to match the data, the same spatial mapping matrix applied to the data must be applied to those fields. In the first case, the spatial mapping applied to the short training symbol would result in

$$\frac{1}{\sqrt{2}} \begin{bmatrix} S \cdot \exp(-j 2\pi k \Delta_F (0 \text{ ns})) \\ S \cdot \exp(-j 2\pi k \Delta_F (-400 \text{ ns})) \\ S \cdot \exp(-j 2\pi k \Delta_F (-200 \text{ ns})) \\ S \cdot \exp(-j 2\pi k \Delta_F (-600 \text{ ns})) \end{bmatrix}$$

which matches the cyclic shifts in Table 4.3. In the second case, the spatial mapping applied to the short training symbol would result in

$$\frac{1}{\sqrt{2}} \begin{bmatrix} S \cdot \exp(-j 2\pi k \Delta_F (0 \text{ ns})) \\ S \cdot \exp(-j 2\pi k \Delta_F (-400 \text{ ns})) \\ S \cdot \exp(-j 2\pi k \Delta_F (0 \text{ ns})) \\ S \cdot \exp(-j 2\pi k \Delta_F (-400 \text{ ns})) \end{bmatrix}$$

Since the same short training symbol is used over all streams, and subsequently all antennas, the resulting cyclic shift must be different for each transmit antenna chain. To summarize, the selection of the spatial mapping matrix must consider the properties of the HT-DATA as well as the HT-STF and HT-LTF.

The 802.11n standard does not specify spatial mapping matrices, though a few examples are given. Table 4.5 gives mapping matrices which result in cyclic shifts on the HT-STF and HT-LTF for each transmit antenna chain that match those in Table 4.3 for the equivalent number of spatial streams. In matrix notation, the resulting signal is $Q_k X_k$, where Q_k is the spatial expansion matrix of dimension $N_{\text{TX}} \times N_{\text{SS}}$ and X_k is the signal vector of dimension $N_{\text{SS}} \times 1$.

The waveform equations for HT-STF, HT-LTF, and HT-DATA defined by Eqs. (4.17), (4.19), and (4.30) are modified to incorporate the spatial mapping the matrix Q which

Table 4.5 Spatial expansion matrices

N_{TX}	N_{SS}	Q_k
2	1	$\frac{1}{\sqrt{2}} \begin{bmatrix} 1 \\ \exp(-j 2\pi k \Delta_F (-400 \text{ ns})) \end{bmatrix}$
3	1	$\frac{1}{\sqrt{3}} \begin{bmatrix} 1 \\ \exp(-j 2\pi k \Delta_F (-400 \text{ ns})) \\ \exp(-j 2\pi k \Delta_F (-200 \text{ ns})) \end{bmatrix}$
4	1	$\frac{1}{2} \begin{bmatrix} 1 \\ \exp(-j 2\pi k \Delta_F (-400 \text{ ns})) \\ \exp(-j 2\pi k \Delta_F (-200 \text{ ns})) \\ \exp(-j 2\pi k \Delta_F (-600 \text{ ns})) \end{bmatrix}$
3	2	$\sqrt{\frac{2}{3}} \begin{bmatrix} 1 & 0 \\ 0 & 1 \\ \exp(-j 2\pi k \Delta_F (-200 \text{ ns})) & 0 \end{bmatrix}$
4	2	$\frac{1}{\sqrt{2}} \begin{bmatrix} 1 & 0 & 0 \\ \exp(-j 2\pi k \Delta_F (-200 \text{ ns})) & 1 & 0 \\ 0 & 0 & \exp(-j 2\pi k \Delta_F (-200 \text{ ns})) \end{bmatrix}$
4	3	$\frac{\sqrt{3}}{2} \begin{bmatrix} 1 & 0 & 0 \\ 0 & 1 & 0 \\ 0 & 0 & 1 \\ \exp(-j 2\pi k \Delta_F (-600 \text{ ns})) & 0 & 0 \end{bmatrix}$

maps the N_{SS} spatial streams to the i_{TX} transmit antenna chain, as follows:

$$r_{\text{Field}}^{(i_{\text{TX}})}(t) = \frac{1}{\sqrt{N_{\text{SS}} \cdot N_{\text{Tone}}^{\text{Field}}}} \sum_{k=-N_{\text{SR}}}^{N_{\text{SR}}} \sum_{i_{\text{SS}}=1}^{N_{\text{SS}}} [Q_k]_{i_{\text{TX}}, i_{\text{SS}}} X_k^{(i_{\text{SS}})} \times \exp(j 2\pi k \Delta_F (t - T_{\text{GT}}^{\text{Field}} - T_{\text{CS}}^{i_{\text{SS}}})) \quad (4.31)$$

where $T_{\text{GT}}^{\text{Field}}$ is the 0.8 μs guard interval for HT-LTF and HT-DATA, 0 for HT-STF; $N_{\text{Tone}}^{\text{Field}} = 56$ for HT-LTF and HT-DATA, 12 for HT-STF; $N_{\text{SR}} = 28$ for HT-LTF and HT-DATA, 26 for HT-STF; N_{SS} is the number of spatial streams; $T_{\text{CS}}^{i_{\text{SS}}}$ represents the cyclic shift for spatial stream i_{SS} with values given by Table 4.3; and $X_k^{(i_{\text{SS}})} = S_k$ for HT-STF, $[P_{\text{HTLTF}}]_{i_{\text{SS}}, n}$ for HT-LTF, and $(D_{k, i_{\text{SS}}, n} + p_{n+3} P_{k, (i_{\text{SS}}, n)}^k)$ for HT-DATA.

As a final note, transmit beamforming is also used to map spatial streams to transmit antenna chains and is addressed in Chapter 12.

4.2.4 HT MF receive procedure

This section describes the receive procedure for the basic modes of operation for 20 MHz. Special conditions due to optional features are addressed in their corresponding sections.

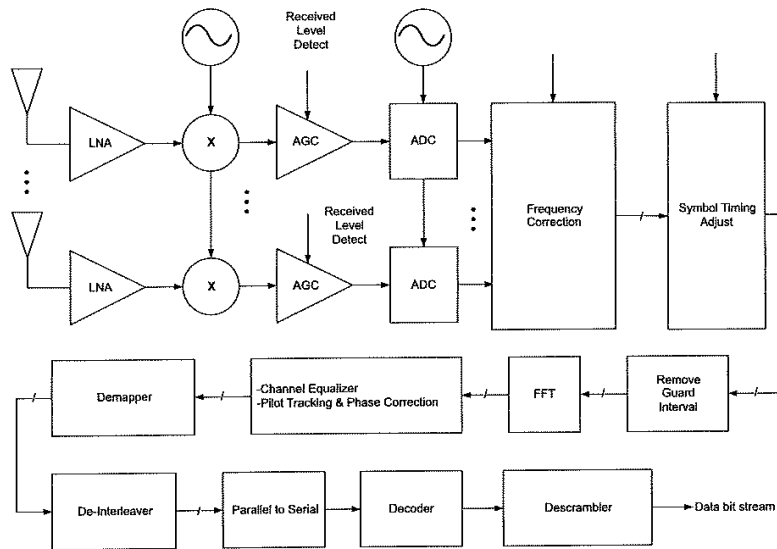


Figure 4.23 Receiver block diagram for MIMO.

Such features include GF preamble, sounding packet, short guard interval, LDPC, and STBC described in Sections 5.4, 12.6, 5.5, 6.4, and 6.3, respectively.

4.2.4.1 RF front end

A basic block diagram for a MIMO receiver is illustrated in Figure 4.23. A MIMO receiver design has multiple receive antenna chains each consisting of an antenna, LNA, down converter, AGC, and ADC. To maximize the dynamic range on each antenna branch, typically each AGC in each receive chain is individually adjusted.

Since the MF packet begins with the legacy preamble, the use of the L-STF for AGC, initial frequency correction, and initial time acquisition is similar to that of 802.11a, described in Section 4.1.6. However, 802.11n devices typically have multiple receive antennas, therefore correlation functions for start-of-packet detection and coarse timing should be summed over all the antennas to maximize detection performance. The coarse time adjustment is applied to all receive signals. Similarly, if in the transmitter all the transmit chains use the same oscillator, and in the receiver all the receive chains use the same oscillator, the frequency estimation described in Section 4.1.1 may also be summed over all receive antenna chains and applied to all received signals. Since the 802.11n standard mandates that the transmit center frequency and the symbol clock frequency for all transmit antennas is to be derived from the same reference oscillator, the requirement of common oscillator source is not onerous. Further description of time

synchronization and frequency estimation for MIMO systems is given in van Zelst and Schenk (2004).

4.2.4.2 Legacy part of the preamble

Processing of the L-LTF and L-SIG follows next as described in Section 4.1.6. However, with the multiple receive antennas of a MIMO receiver, more robust processing may be performed on the single stream transmission of the L-SIG than with a single antenna legacy receiver. Reception of a single stream transmission by multiple receive antennas is generally described in Section 3.2. Furthermore, the benefits of maximal ratio combining are described in Section 6.1.

With a legacy packet, a check of the SIG parity bit to determine the validity of the packet would occur. However, if the packet is an HT packet, it is not necessary to fail the packet upon an invalid L-SIG parity bit. Therefore, the check of the L-SIG parity bit only needs to occur if it has been determined based on auto-detection that the packet is a legacy packet. The auto-detection between legacy and HT MF packets occurs during the next symbol following the L-SIG, which is the HT-SIG for an HT MF packet as described in Section 4.2.2.1.

4.2.4.3 High Throughput Signal field (HT-SIG)

Except for the 90 degree phase rotation, the HT-SIG field is modulated and encoded similarly to the L-SIG and the receive processing steps are the same as that of the L-SIG. With regard to decoding, the HT-SIG data are encompassed by two OFDM symbols and are treated as a single block when decoded. Since the frequency correction and time synchronization were computed based on L-STF and L-LTF, it is advisable that phase rotation correction computed during the L-SIG and HT-SIG be used for further frequency offset and symbol timing correction before processing the HT-LTF. This minimizes the error in the channel estimates computed from the HT-LTF.

If signal loss occurs at any point during processing during the legacy part of the preamble and the HT-SIG, the receiver terminates reception. However, the device must not access the medium (defer) for as long as the received signal level remains above -62 dBm (for 20 MHz operation). This ensures that even if the receiver is not able to process the signal, the fundamental nature of carrier sense multiple access is still maintained. This basic operation is termed clear channel assessment (CCA).

In the next step of the receive procedure, the HT-SIG CRC is checked for validity. With a valid determination of the HT-SIG CRC, the information bits in the HT-SIG (described in Section 4.2.2.1) are used to configure the receiver for reception of the packet. At this point even following signal loss, the device refrains from accessing the medium until the transmit time (TXTIME) computed based on the HT-SIG values has expired:

$$\text{TXTIME} = T_{L\text{-STF}} + T_{L\text{-LTF}} + T_{L\text{-SIG}} + N_{LTF} \cdot T_{HT\text{-LTF}} + T_{HT\text{-SIG}} + T_{SYM} \cdot N_{SYM} \quad (4.32)$$

where $T_{L\text{-STF}}$, $T_{L\text{-LTF}}$, $T_{HT\text{-SIG}} = 8 \mu\text{s}$ duration; $T_{L\text{-SIG}}$, $T_{HT\text{-LTF}} = 4 \mu\text{s}$ duration; $T_{SYM} = 4 \mu\text{s}$ duration for 800 ns guard interval operation; N_{SYM} is the number of OFDM

symbols in the Data field defined by Eq. (4.22); and N_{LTF} is the number of long training symbols, determined from the MCS – the number of long training symbols is based on the number of spatial streams indicated by the MCS (excluding a sounding packet).

The computation of TXTIME changes for GF preamble, short guard interval, sounding packet, LDPC, and STBC. The modifications for TXTIME are described in Sections 5.4.4, 5.5, 12.6, 6.4, and 6.3, respectively.

TXTIME must be computed even if the indicated rate or mode is not supported. An exception to the TXTIME computation is if the reserved bit is decoded as 0. If the HT-SIG CRC is valid, then a reserved bit value of 0 indicates an invalid condition or a false positive of the CRC computation. Furthermore, if the MCS field is in the range 77–127 or the STBC field has a value of 3, TXTIME can not be computed. In these conditions, the device must not access the medium for as long as the received signal level remains above –62 dBm (for 20 MHz operation), rather than based on the TXTIME computation.

One further note on proper CCA, even though the GF preamble is optional, an HT device not capable of receiving such preamble type must still properly defer. Rules regarding CCA of a GF packet received by an HT non-GF capable device are discussed in Section 5.4.4.

4.2.4.4 High Throughput Training fields and MIMO channel estimation

Following the HT-SIG fields is the HT-STF that is used to readjust the AGC, as described in Section 4.2.2.2. The subsequent symbols comprise the HT-LTF used for MIMO channel estimation. To explain the steps to compute the channel estimates for the MIMO channel, we begin with a two stream, two transmit and two receive antenna example. The receiver processes each 3.2 μ s long training symbol by removing the cyclic prefix from the HT-LTF and performs an FFT to extract the training subcarriers. For the first symbol in time, indicated by t_1 , the frequency domain representation for each training subcarrier k at the receiver is represented by Eq. (4.33). Note that z is AWGN. Refer to Figure 3.4 in Section 3.3 for a general illustration of the MIMO system.

$$\begin{aligned} y_{1,t_1}^k &= h_{11}^k \cdot \text{HTLTF} \cdot e^{-j2\pi k \Delta_F T_{CS}^1} + h_{12}^k \cdot \text{HTLTF} \cdot e^{-j2\pi k \Delta_F T_{CS}^2} + z_{1,t_1}^k \\ y_{2,t_1}^k &= h_{21}^k \cdot \text{HTLTF} \cdot e^{-j2\pi k \Delta_F T_{CS}^1} + h_{22}^k \cdot \text{HTLTF} \cdot e^{-j2\pi k \Delta_F T_{CS}^2} + z_{2,t_1}^k \end{aligned} \quad (4.33)$$

Then at the second symbol in time, t_2 ,

$$\begin{aligned} y_{1,t_2}^k &= h_{11}^k \cdot \text{HTLTF} \cdot (-1) \cdot e^{-j2\pi k \Delta_F T_{CS}^1} + h_{12}^k \cdot \text{HTLTF} \cdot e^{-j2\pi k \Delta_F T_{CS}^2} + z_{1,t_2}^k \\ y_{2,t_2}^k &= h_{21}^k \cdot \text{HTLTF} \cdot (-1) \cdot e^{-j2\pi k \Delta_F T_{CS}^1} + h_{22}^k \cdot \text{HTLTF} \cdot e^{-j2\pi k \Delta_F T_{CS}^2} + z_{2,t_2}^k \end{aligned} \quad (4.34)$$

we see the effect of the orthogonal mapping matrix in Eq. (4.20). The “–1” term is from the first row and second column of the matrix. To simplify the expressions, we combine the channel tap h with the cyclic shift, $\tilde{h}_{i,m} = h_{i,m} e^{-j2\pi k \Delta_F T_{CS}^m}$. As a note, the cyclic shifts applied to the long training symbols are also applied to the data symbols. Redefining the channel taps to incorporate the cyclic shift has the added advantage of automatically removing the effect of the cyclic shift when equalizing the data with the channel estimate, since the channel estimate is the combination of the channel tap and the cyclic shift. The simplification for Eqs. (4.33) and (4.34) is given in the equation

below:

$$\begin{aligned} y_{1,t_1}^k &= \tilde{h}_{11}^k \cdot \text{HTLTF} + \tilde{h}_{12}^k \cdot \text{HTLTF} + z_{1,t_1}^k \\ y_{2,t_1}^k &= \tilde{h}_{21}^k \cdot \text{HTLTF} + \tilde{h}_{22}^k \cdot \text{HTLTF} + z_{2,t_1}^k \\ y_{1,t_2}^k &= \tilde{h}_{11}^k \cdot \text{HTLTF} \cdot (-1) + \tilde{h}_{12}^k \cdot \text{HTLTF} + z_{1,t_2}^k \\ y_{2,t_2}^k &= \tilde{h}_{21}^k \cdot \text{HTLTF} \cdot (-1) + \tilde{h}_{22}^k \cdot \text{HTLTF} + z_{2,t_2}^k \end{aligned} \quad (4.35)$$

Once the two symbols are processed and the subcarriers are extracted, the subcarriers from symbol one are combined with the subcarriers from symbol two to derive the channel estimates. First the symbols from antenna one are combined, as follows. The subcarriers from the first symbol are subtracted by the subcarriers from the second symbol to derive the channel estimate for $\tilde{h}_{11}^k$. This cancels out the $\tilde{h}_{12}^k$ term. Then the subcarriers from the first symbol are added with the subcarriers from the second symbol to derive the channel estimate for $\tilde{h}_{12}^k$. This cancels out the $\tilde{h}_{11}^k$ term. This is expressed in the equation below:

$$\begin{aligned} \tilde{h}_{11}^k &= \frac{y_{1,t_1}^k - y_{1,t_2}^k}{2 \cdot \text{HTLTF}} \\ \tilde{h}_{12}^k &= \frac{y_{1,t_1}^k + y_{1,t_2}^k}{2 \cdot \text{HTLTF}} \end{aligned} \quad (4.36)$$

Using the subcarriers from the second antenna, we similarly derive the channel estimate for $\tilde{h}_{21}^k$ and $\tilde{h}_{22}^k$, as follows:

$$\begin{aligned} \tilde{h}_{21}^k &= \frac{y_{2,t_1}^k - y_{2,t_2}^k}{2 \cdot \text{HTLTF}} \\ \tilde{h}_{22}^k &= \frac{y_{2,t_1}^k + y_{2,t_2}^k}{2 \cdot \text{HTLTF}} \end{aligned} \quad (4.37)$$

Combining the subcarriers from the two symbols to extract the channel estimate is possible due to the orthogonal mapping matrix.

The derivation of the channel estimates can be generalized to an arbitrary number of spatial streams and receive antennas. First we define the received signal vector Y_n with the number of elements equal to the number of receive antennas (N_{RX}), and the Long Training Field index n , $1 \leq n \leq N_{LTF}$. The received signal vector from each symbol is combined into a single matrix, $[Y_1 | Y_2 | \dots | Y_{N_{LTF}}]$ having the dimensions $N_{RX} \times N_{LTF}$, similarly with the noise vector Z , given in the equation below:

$$\begin{aligned} & \begin{bmatrix} Y_{t_1}^k | Y_{t_2}^k | \dots | Y_{t_{N_{LTF}}}^k \end{bmatrix} \\ &= \begin{bmatrix} \tilde{h}_{11}^k & \tilde{h}_{12}^k & \dots & \tilde{h}_{1N_{SS}}^k \\ \tilde{h}_{21}^k & \tilde{h}_{22}^k & \dots & \tilde{h}_{2N_{SS}}^k \\ \vdots & \vdots & \ddots & \vdots \\ \tilde{h}_{N_{RX}1}^k & \tilde{h}_{N_{RX}2}^k & \dots & \tilde{h}_{N_{RX}N_{SS}}^k \end{bmatrix} \cdot \tilde{P} \cdot \text{HTLTF}_k + [Z_{t_1}^k | Z_{t_2}^k | \dots | Z_{t_{N_{LTF}}}^k] \end{aligned} \quad (4.38)$$

where $\tilde{P}$ is an $N_{SS} \times N_{LTF}$ subset of the orthogonal mapping matrix P_{HTLTF} . For example, with three spatial streams and four long training symbols,

$$\tilde{P} = \begin{pmatrix} 1 & -1 & 1 & 1 \\ 1 & 1 & -1 & 1 \\ 1 & 1 & 1 & -1 \end{pmatrix}$$

To compute the channel estimate matrix we multiply the received signal vector matrix by the transpose of $\tilde{P}$ and normalize, as follows:

$$\begin{bmatrix} \hat{h}_{11}^k & \hat{h}_{12}^k & \cdots & \hat{h}_{1N_{SS}}^k \\ \hat{h}_{21}^k & \hat{h}_{22}^k & \cdots & \hat{h}_{2N_{SS}}^k \\ \vdots & \vdots & \ddots & \vdots \\ \hat{h}_{N_{RX}1}^k & \hat{h}_{N_{RX}2}^k & \cdots & \hat{h}_{N_{RX}N_{SS}}^k \end{bmatrix} = \begin{bmatrix} Y_{t1}^k |Y_{t2}^k| \cdots |Y_{tN_{LTF}}^k| \end{bmatrix} \cdot \tilde{P}^T \cdot \frac{1}{N_{LTF} \cdot HTLTF_k} \quad (4.39)$$

4.2.4.5 Data field

Following the HT-LTF is the Data field, which is processed as illustrated in Figure 4.23. For each 4 μ s symbol, the 0.8 μ s GI is removed and an FFT is performed. The data may be equalized by a linear equalizer as described in Section 3.6, using the channel estimates from the HT-LTF (Eq. (4.39)).

In processing the pilot subcarriers to track the phase rotation, we again highlight the advantage of the receiver using a common oscillator for all the receive chains. A particular pilot subcarrier in the transmitted OFDM symbol in the frequency domain is described by the equation below:

$$\begin{bmatrix} y_1 \\ y_2 \\ \vdots \\ y_N \end{bmatrix} = \begin{bmatrix} h_{11} & h_{12} & \cdots & h_{1M} \\ h_{21} & h_{22} & \cdots & h_{2M} \\ \vdots & \vdots & \ddots & \vdots \\ h_{N1} & h_{N2} & \cdots & h_{NM} \end{bmatrix} \begin{bmatrix} p_1 \\ p_2 \\ \vdots \\ p_M \end{bmatrix} + \begin{bmatrix} z_1 \\ z_2 \\ \vdots \\ z_N \end{bmatrix}$$

or, in vector notation,

$$Y = HP + Z$$

If a common oscillator is employed, the received signal is described by

$$R = e^{j\phi} Y = HP e^{j\phi} + Z e^{j\phi}$$

Extracting the pilot subcarriers and equalizing with the channel estimate results in

$$H^{-1}R = P e^{j\phi} + H^{-1}Z e^{j\phi}$$

for a ZF receiver. We may further enhance the estimation of the phase rotation by averaging over the spatial streams. As described in Akay and Ayanoglu (2004b), a weighted average using the channel estimates improves performance by de-weighting the subcarriers which are faded.

When a common oscillator is not employed, the received signal is described by the equation below:

$$\begin{bmatrix} r_1 \\ r_2 \\ \vdots \\ r_N \end{bmatrix} = \begin{bmatrix} e^{j\phi_1} & 0 & \cdots & 0 \\ 0 & e^{j\phi_2} & \cdots & 0 \\ \vdots & \vdots & \ddots & \vdots \\ 0 & 0 & \cdots & e^{j\phi_N} \end{bmatrix} \begin{bmatrix} y_1 \\ y_2 \\ \vdots \\ y_N \end{bmatrix}$$

or, in vector notation,

$$R = \Phi Y = \Phi HP + \Phi Z.$$

In this case, it is difficult to separate the phase rotation from the channel. Such implementations result in inferior pilot tracking performance.

4.2.4.6 Demapping, deinterleaving, decoding, and descrambling

The demapping operation on each spatial stream is the same for an 802.11a receiver. The deinterleaver uses the following operations to perform the inverse of the interleaving described in Section 4.2.3.4.

The index of the bit in the received block (per spatial stream i_{ss}) is denoted by k . The first permutation reverses the third (frequency rotation) permutation of the interleaver (defined in Eq. (4.27)) if there is more than one spatial stream:

$$y_r^{i_{ss}} = z_k^{i_{ss}} \quad r = \left(k + \left(((i_{ss} - 1) \cdot 2) \bmod 3 + 3 \cdot \left\lfloor \frac{i_{ss} - 1}{3} \right\rfloor \right) \cdot 11 \cdot N_{BPSS} \right) \bmod N_{CBPSS} \quad (4.40)$$

where $1 \leq i_{ss} \leq N_{SS}$, $k = 0, 1, \dots, N_{CBPSS} - 1$. N_{BPSS} is the number of coded bits per single subcarrier for each spatial stream, or equivalently the modulation order, and N_{CBPSS} is the coded bits per spatial stream.

The second permutation in the deinterleaver reverses the second permutation in the interleaver (defined in Eq. (4.26)) for each spatial stream:

$$w_j^{i_{ss}} = y_k^{i_{ss}} \quad j = s \cdot \text{floor}(k/s) + \left(k + \left\lfloor \frac{13 \cdot k}{N_{CBPSS}} \right\rfloor \right) \bmod s \quad (4.41)$$

where $1 \leq i_{ss} \leq N_{SS}$, $k = 0, 1, \dots, N_{CBPSS} - 1$, and s is defined in Eq. (4.23).

The third permutation in the deinterleaver reverses the first permutation of the interleaver (defined in Eq. (4.25)) for each spatial stream:

$$x_i^{i_{ss}} = w_k^{i_{ss}} \quad i = 13 \cdot k - (N_{CBPSS} - 1) \cdot \left\lfloor \frac{k}{4 \cdot N_{BPSS}} \right\rfloor \quad (4.42)$$

where $1 \leq i_{ss} \leq N_{SS}$, $k = 0, 1, \dots, N_{CBPSS} - 1$.

As the final step in the deinterleaving process, the stream parsing operation (described in Section 4.2.3.3) is reversed by taking blocks of s bits from each spatial stream in a round robin fashion to create a single bit stream for the decoder.

The resulting convolutional coded data stream is decoded typically with a Viterbi decoder. If the MCS indicates a rate higher than $\frac{1}{2}$, depuncturing is performed at the input of each decoder. Decoder performance can be substantially improved by using the MIMO channel estimates to weight the decoder metrics. Decoding is similar to 802.11a and described further in Section 4.1.6.

The first seven bits of the Service field give the initial state of the descrambler. The same scrambler used in 802.11a is used to descramble the data stream. Bits received after the data length, indicated by the HT-SIG, are discarded.

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Appendix 4.1: 20 MHz basic MCS tables

The following tables consist of the MCSs for 20 MHz “basic” rates. As defined herein, the basic rates include equal modulation and the mandatory 800 ns guard interval. 40 MHz

MCSs are described in Section 5.1.4, unequal modulation is described in Section 12.4, and short guard interval is described in Section 5.5.

For 20 MHz, there are 52 data subcarriers and four pilot subcarriers in an OFDM symbol in an individual spatial stream.

Table 4.6 describes the symbols used in the subsequent tables.

Table 4.6 Symbols used for MCS parameters (IEEE, 2007b)

Symbol	Explanation
R	Code rate
N_{BPSCS}	Number of coded bits per single carrier for each spatial stream (or modulation order)
N_{CBPS}	Number of coded bits per OFDM symbol
N_{DBPS}	Number of data bits per OFDM symbol

Table 4.7 defines the 20 MHz MCS parameters and data rates for single spatial stream transmission. The data rate is calculated by dividing N_{DBPS} by the symbol time of 4 μs (3.2 μs OFDM symbol plus 0.8 μs guard interval).

Table 4.7 20 MHz MCS parameters for single spatial stream (IEEE, 2007b)

MCS index	Modulation	R	N_{BPSCS}	N_{CBPS}	N_{DBPS}	Data rate (Mbps)
0	BPSK	$\frac{1}{2}$	1	52	26	6.5
1	QPSK	$\frac{1}{2}$	2	104	52	13.0
2	QPSK	$\frac{3}{4}$	2	104	78	19.5
3	16-QAM	$\frac{1}{2}$	4	208	104	26.0
4	16-QAM	$\frac{3}{4}$	4	208	156	39.0
5	64-QAM	$\frac{2}{3}$	6	312	208	52.0
6	64-QAM	$\frac{3}{4}$	6	312	234	58.5
7	64-QAM	$\frac{5}{6}$	6	312	260	65.0

Table 4.8 defines the 20 MHz MCS parameters and data rates for two spatial stream transmission.

Table 4.8 20 MHz MCS parameters for two spatial streams (IEEE, 2007b)

MCS index	Modulation	R	N_{BPSCS}	N_{CBPS}	N_{DBPS}	Data rate (Mbps)
8	BPSK	$\frac{1}{2}$	1	104	52	13.0
9	QPSK	$\frac{1}{2}$	2	208	104	26.0
10	QPSK	$\frac{3}{4}$	2	208	156	39.0
11	16-QAM	$\frac{1}{2}$	4	416	208	52.0
12	16-QAM	$\frac{3}{4}$	4	416	312	78.0
13	64-QAM	$\frac{2}{3}$	6	624	416	104.0
14	64-QAM	$\frac{3}{4}$	6	624	468	117.0
15	64-QAM	$\frac{5}{6}$	6	624	520	130.0

Table 4.9 defines the MCS parameters and data rates for three spatial stream transmission.

Table 4.9 20 MHz MCS parameters for three spatial streams (IEEE, 2007b)

MCS index	Modulation	R	N_{BPSCS}	N_{CBPS}	N_{DBPS}	Data rate (Mbps)
16	BPSK	$1/2$	1	156	78	19.5
17	QPSK	$1/2$	2	312	156	39.0
18	QPSK	$3/4$	2	312	234	58.5
19	16-QAM	$1/2$	4	624	312	78.0
20	16-QAM	$3/4$	4	624	468	117.0
21	64-QAM	$2/3$	6	936	624	156.0
22	64-QAM	$3/4$	6	936	702	175.5
23	64-QAM	$5/6$	6	936	780	195.0

Table 4.10 defines the MCS parameters and data rates for four spatial stream transmission.

Table 4.10 20 MHz MCS parameters for four spatial streams (IEEE, 2007b)

MCS index	Modulation	R	N_{BPSCS}	N_{CBPS}	N_{DBPS}	Data rate (Mbps)
24	BPSK	$1/2$	1	208	104	26.0
25	QPSK	$1/2$	2	416	208	52.0
26	QPSK	$3/4$	2	416	312	78.0
27	16-QAM	$1/2$	4	832	416	104.0
28	16-QAM	$3/4$	4	832	624	156.0
29	64-QAM	$2/3$	6	1248	832	208.0
30	64-QAM	$3/4$	6	1248	936	234.0
31	64-QAM	$5/6$	6	1248	1040	260.0

5 High throughput

802.11n is the high throughput amendment to the 802.11 standard. This section describes all the aspects of the physical layer which increase the data rate. MIMO/SDM is a key feature of 802.11n, which is discussed in Chapters 3 and 4. The other significant increase in data rate is derived from the new 40 MHz channel width. This section also discusses short guard interval, Greenfield preamble, and other modifications to the 20 MHz waveform.

5.1 40 MHz channel

In the last several years, regulatory domains have made much more spectrum available for unlicensed operation in the 5.47–5.725 GHz band for wireless local area networking. The addition of the new spectrum has more than doubled the number of available 20 MHz channels in the USA and Europe. Table 5.9 and Table 5.10 in Appendix 5.1 describe the current allocation in the USA and Europe, and the corresponding 802.11 channel number. Even with doubling the channel width to create 40 MHz channels, there are still more channels available for frequency re-use than in the early days of 802.11a. Furthermore, products currently in the market with proprietary 40 MHz modes have demonstrated that the cost for 40 MHz products is roughly the same as for 20 MHz products. Therefore, with free spectrum and relatively no increase in hardware cost, doubling the channel bandwidth is the simplest and most cost effective way to increase data rate.

There are some issues that arise with the adoption of the new 40 MHz channel. Coexistence and interoperability between 20 MHz and 40 MHz devices and modes of operation need to be addressed. In addition, though a minor point, frequency re-use in 40 MHz is higher than with 20 MHz. Larger high density network deployments may need to consider this in their network plan. Finally, although the initial intention of the 40 MHz channel was for 5 GHz because of the additional new spectrum, 40 MHz channels and operation are permitted in IEEE (2007) in the 2.4 GHz band. Due to the limited spectrum and overlapping channels, 40 MHz operation in 2.4 GHz requires special attention, and is addressed in Section 10.3.3.2.

In this section simulation results of packet error rate versus SNR are provided for 40 MHz to demonstrate performance. The simulations are performed in channel model D, NLOS, as described in Section 3.5. Physical layer impairments are included in

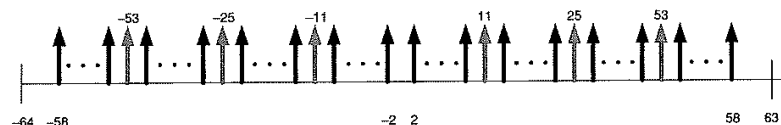


Figure 5.1 40 MHz subcarrier design.

the simulation, as described in Section 3.5.4. The equalizer is based on MMSE. The simulations include synchronization, channel estimation, and phase tracking.

5.1.1 40 MHz subcarrier design and spectral mask

The 40 MHz subcarrier design is illustrated in Figure 5.1. The frequency domain design for 40 MHz encompasses 128 subcarrier locations, with a subcarrier spacing the same as for 20 MHz, $40 \text{ MHz}/128 = 312.5 \text{ kHz}$. For the guard band on the edges of the channel, there are six null subcarriers on the left side of the band, subcarriers $-64, \dots, -59$, and there are five null subcarriers on the right side of the band, $59, \dots, 63$. In addition, there are three nulls around DC, $-1, 0, 1$ to ease the implementation of direct down conversion receivers. In comparison, the 20 MHz HT subcarrier design has four null subcarriers on the left side of the band, three null subcarriers on the right side, and only has one null subcarrier at DC. For 40 MHz, there is a total of 14 null subcarriers. There are 114 populated subcarriers in locations $-58, \dots, -2, \dots, 58$. Of these, six are for pilot subcarriers in locations $-53, -25, -11, 11, 25, 53$. The wider bandwidth called for an additional two pilot subcarriers with respect to 20 MHz to accurately track the phase across the entire band. The remaining 108 subcarriers are dedicated to data. This is more than double the number of data subcarriers as 20 MHz which has 52 for HT and 48 for legacy, and is reflected in the 40 MHz data rates which are more than double those of 20 MHz. The MCS tables for 40 MHz, and the associated 40 MHz data rates are given in Tables 5.14–5.18 in Appendix 5.2. For example, MCS 15 for 20 MHz is 130 Mbps, and for 40 MHz is 270 Mbps.

With the new channel width and subcarrier design, a new 40 MHz spectral mask is necessary. The basic premise of the design of the 40 MHz spectral mask is that adjacent channel interference between neighboring 40 MHz devices is the same as the adjacent channel interference between neighboring 20 MHz HT devices. The new spectral mask is given in Figure 5.2. For reference, the new 20 MHz HT spectral mask is given in Figure 5.1.1.

When a 40 MHz device and a 20 MHz device are in neighboring channels, the 20 MHz device experiences more adjacent channel interference than when neighboring another 20 MHz device. Figure 5.3 illustrates a 20 MHz device with a neighboring 20 MHz device and a neighboring 40 MHz device, all at equal received power level. We see that the roll-off from the 40 MHz spectral mask is at levels of -20 dB to -28 dB within the 20 MHz pass band. This is interference which the 20 MHz device can not filter out. On the other hand, we see that the roll-off from the neighboring 20 MHz device

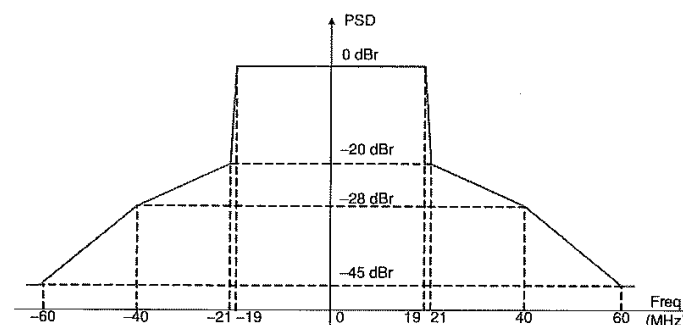


Figure 5.2 40 MHz spectral mask. Reproduced with permission from IEEE (2007) © IEEE.

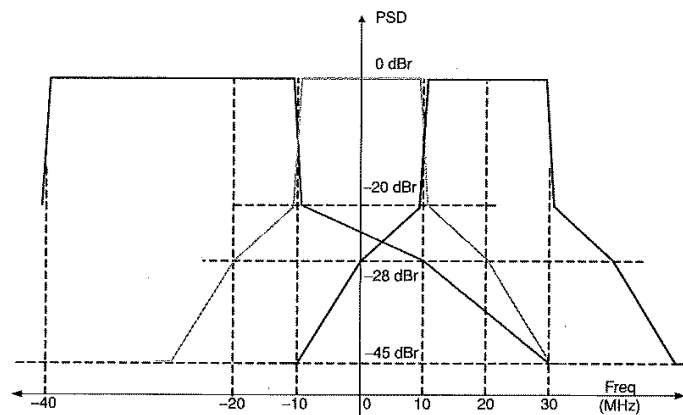


Figure 5.3 Adjacent channel interference between 20 MHz and 40 MHz.

is at levels of -20 dB to -45 dB within the 20 MHz pass band. This demonstrates the possibility of higher adjacent channel interference from the 40 MHz device than the 20 MHz device. Note, however, that regulatory restrictions remain, so near the edge of a regulatory band, the situation can be better.

The other component of adjacent channel interference is the relative power between the signal-of-interest and the interference. With the additional spectrum allocation in 5 GHz, most 5 GHz deployments are capable of placing spatially neighboring APs on non-neighboring channels. Therefore, APs on neighboring channels have a large spatial separation and the interference power is lower. This is a reason why the less stringent 40 MHz spectral mask may not have an effect on system performance.

The main reason why the 40 MHz spectral mask was not made tighter, was to minimize the increase in the cost and complexity and minimize the decrease in power efficiency of the transmit power amplifier and transmit filtering. The main source of out-of-band interference is due to spectral re-growth from the power amplifier. To meet a tighter mask would have meant further power backoff to run the power amplifier at a more linear operating point. Additional backoff reduces the output power, thereby reducing coverage. Conversely, a larger power amplifier could be used to maintain the same output power. However, this is more costly and consumes more DC power hence reducing battery life in power sensitive devices.

5.1.2 40 MHz channel design

Fundamental to the design of the 40 MHz channel is that only adjacent 20 MHz channels are combined to form a 40 MHz channel. Frequency band and associated channel numbering for 20 MHz is discussed Appendix 5.1. The 20 MHz channel numbers are given in Tables 5.9 and 5.10. A 40 MHz channel is specified by two fields: (Nprimary_ch, Secondary). The first field represents the channel number of the primary 20 MHz channel. This is the channel by which AP transmits all control and management frames in a mixed 20/40 MHz environment. In the same type of environment, all 20 MHz clients only associate to the primary channel since the beacon is only transmitted in the primary channel. The second field indicates whether the secondary 20 MHz channel is above or below the primary channel (1 → above, -1 → below). The secondary channel number is equal to (Nprimary_ch + Secondary*4). For example, a 40 MHz channel consisting of a Nprimary_ch = 36 and Secondary = 1 specifies the use of channel 36 for the primary channel and channel 40 for the secondary channel.

As can be seen from Tables 5.9 and 5.10 in Appendix 5.1, the 20 MHz channels in the 5 GHz band are non-overlapping. This makes coexistence and interoperability between 20 MHz and 40 MHz devices much simpler. Keeping with this same philosophy, the new 40 MHz channels are also non-overlapping, as given in Tables 5.11 and 5.12.

Unfortunately in the 2.4 GHz band, the channels are overlapping. To better adjust the selection of a 40 MHz channel to the environment, the new 40 MHz channel numbers are also overlapping. The channel numbers for 2.4 GHz are also given in Tables 5.11 and 5.12.

5.1.3 40 MHz mixed format preamble

In order to interoperate and coexist with legacy 802.11a/g OFDM devices the legacy portion of the 20 MHz MF preamble is replicated over the other 20 MHz portion of the 40 MHz band. This is illustrated in Figure 5.4. This allows a legacy OFDM device in either 20 MHz portion of the 40 MHz channel to process the L-STF, L-LTF, and decode the L-SIG to properly defer transmission. Therefore legacy OFDM devices in a 40 MHz BSS which operate on the primary channel coexist with the 40 MHz HT devices. Furthermore legacy OFDM devices in a neighboring BSS which overlaps the secondary channel of the 40 MHz BSS also coexist with the 40 MHz HT devices.

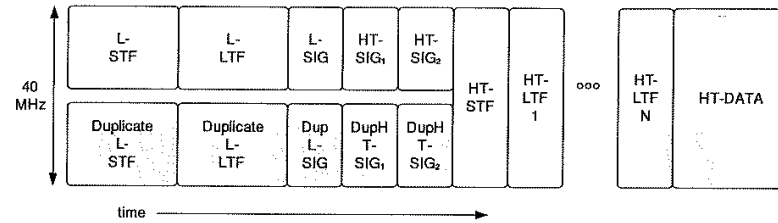


Figure 5.4 40 MHz mixed mode preamble.

Similarly, the HT-SIG fields are also replicated over the other 20 MHz portion of the 40 MHz channel. This allows 20 MHz HT devices in either portion of the 40 MHz channel to decode the HT-SIG fields and properly defer transmission. Therefore 20 MHz and 40 MHz HT devices in a 40 MHz BSS will coexist, and similarly with 20 MHz HT devices in a neighboring BSS on the secondary channel.

As noted in Section 4.2.2.1, when the CBW 20/40 bit in the HT-SIG is set to 0, a 20 MHz transmission is indicated. When the bit is set to 1, a 40 MHz transmission is indicated. This allows a 20 MHz device to terminate reception in the presence of a 40 MHz transmission. A 40 MHz device may try to auto-detect between 20 MHz and 40 MHz prior to decoding the HT-SIG. For example, at initial packet detection, a receive signal strength measure on either half of the band may be used to indicate that both portions contain energy, indicating 40 MHz. Since the subcarriers in the 40 MHz preamble are replicated, the upper and lower portions may be combined for extra frequency diversity and processing gain. This improves detection of L-SIG and HT-SIG. Care needs to be taken to not confuse a 20 MHz transmission in one portion of the 40 MHz channel and interference on the other 20 MHz portion of the 40 MHz channel with a 40 MHz transmission.

The 20 MHz OFDM symbol waveform equations for L-STF, L-LTF, L-SIG, and HT-SIG defined by Eqs. (4.13)–(4.16) are modified for 40 MHz transmission in two ways, see Eq. (5.1). The 20 MHz waveform is shifted lower by 32 subcarriers. A copy of the waveform is shifted higher by 32 subcarriers and phase rotated by 90 degrees (the j term preceding the second summation in Eq. (5.1)). The phase rotation on the upper subcarriers is added to provide a reduction in peak-to-average ratio of the 40 MHz waveform in some conditions. The general waveform equation for MF L-STF, L-LTF, L-SIG, and HT-SIG fields for 40 MHz transmission is given in the equation below.

$$r_{\text{Field}}^{(i_{\text{TX}})}(t) = \frac{1}{\sqrt{N_{\text{TX}} N_{\text{Field}}^{\text{Tone}}}} \left[\sum_{k=-26}^{26} X_k \exp \left(j 2\pi (k - 32) \Delta_F (t - T_{\text{GI}}^{\text{Field}} - T_{\text{CS}}^{i_{\text{TX}}}) \right) + j \sum_{k=-26}^{26} X_k \exp \left(j 2\pi (k + 32) \Delta_F (t - T_{\text{GI}}^{\text{Field}} - T_{\text{CS}}^{i_{\text{TX}}}) \right) \right] \quad (5.1)$$

where N_{TX} is the number of transmit antenna chains; $N_{\text{Field}}^{\text{Tone}} = 24$ for L-STF, and 104 for L-LTF, L-SIG, and HT-SIG; $X_k = S_k$ in Eq. (4.13) for L-STF, L_k in Equation (4.14) for L-LTF, $(D_k + p_0 P_k)$ in Eq. (4.15) for L-SIG, and $(j D_{k,n} + p_{n+1} P_k)$ in Eq. (4.16) for

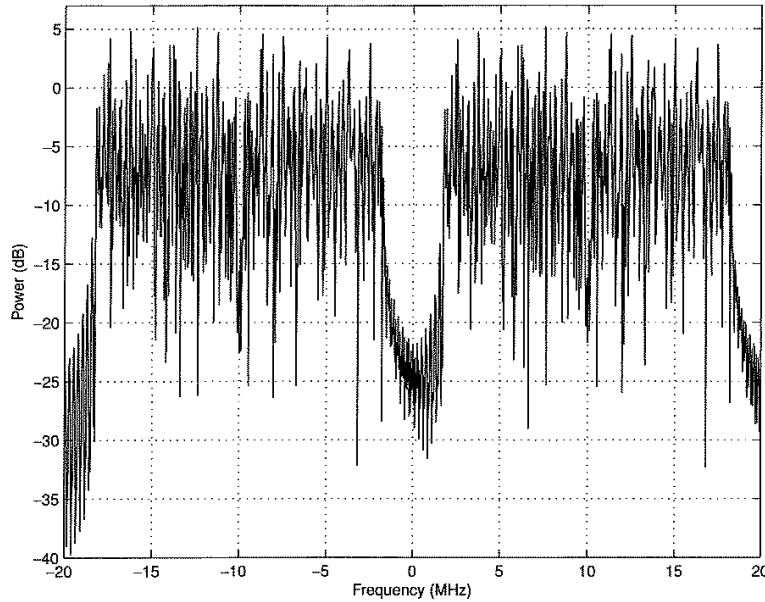


Figure 5.5 Spectrum of the legacy portion of the 40 MHz MF preamble.

HT-SIG; T_{GI}^{Field} is the 1.6 μ s guard interval for L-LTF, 0.8 μ s for L-SIG and HT-SIG, 0 for L-STF; and T_{CS}^{TX} is the cyclic shift for transmit antenna chain i_{TX} with values given by Table 4.2.

An illustration of the spectrum of the legacy portion of the 40 MHz MF preamble is given in Figure 5.5. As can be seen, the legacy portion of the preamble contains many null subcarriers in the area around DC.

Once the legacy portion of the preamble is complete, legacy compatibility no longer constrains the preamble design. The HT-STF, HT-LTF, and HT-DATA incorporate all the additional subcarriers available in 40 MHz based on the subcarrier design illustrated in Figure 5.1. Again, for peak-to-average ratio considerations, the upper portion of the subcarriers are phase rotated by 90 degrees. The 20 MHz waveform equations for HT-STF, HT-LTF, and HT-DATA defined by Eqs. (4.17), (4.19), and (4.30) are modified for 40 MHz, as follows:

$$r_{Field}^{(iss)}(t) = \frac{1}{\sqrt{N_{SS} \cdot N_{Tone}^{Field}}} \left[\sum_{k=-58}^0 X_k \exp(j 2\pi k \Delta_F (t - T_{GI}^{Field} - T_{CS}^{iss})) \right. \\ \left. + j \sum_{k=1}^{58} X_k \exp(j 2\pi k \Delta_F (t - T_{GI}^{Field} - T_{CS}^{iss})) \right] \quad (5.2)$$

where N_{SS} is the number of spatial streams; $N_{Tone}^{Field} = 24$ for HT-STF, and 114 for HT-LTF and HT-DATA; $X_k^{(iss)} = S_k$ for HT-STF, $[P_{HTLTF}]_{iss,n}$ HTLTF_k for HT-LTF, and $(D_{k,iss,n} + p_{n+3} P_{k,iss,n}^k)$ for HT-DATA; T_{GI}^{Field} : 0.8 μ s guard interval for HT-LTF and HT-DATA, 0 for HT-STF; and T_{CS}^{iss} is the cyclic shift for spatial stream i_{ss} with values given by Table 4.3.

The HT-STF subcarrier sequence for 40 MHz is derived from the 20 MHz L-STF sequence defined in Eq. (4.13), since its low peak-to-average ratio, high autocorrelation properties, and short repetition periods are still desirable. The 20 MHz sequence is shifted lower by 32 and a copy is shifted up by 32, as given in Eq. (5.3) (IEEE, 2007):

$$S_{-58,58} = \sqrt{1/2} \{0, 0, 1 + j, 0, 0, 0, -1 - j, 0, 0, 0, 1 + j, 0, 0, 0, \\ -1 - j, 0, 0, 0, -1 - j, 0, 0, 0, 1 + j, 0, 0, 0, 0, 0, 0, 0, 0, -1 - j, 0, 0, 0, \\ -1 - j, 0, 0, 0, 1 + j, 0, 0, 0, 1 + j, 0, 0, 0, 1 + j, 0, 0, 0, 1 \\ + j, 0, 0, 0, 0, 0, 0, 0, 0, 0, 0, 0, 0, 0, 0, 1 + j, 0, 0, 0, -1 - j, 0, 0, 0, 1 \\ + j, 0, 0, 0, 0, -1 - j, 0, 0, 0, -1 - j, 0, 0, 0, 1 + j, 0, 0, 0, 0, 0, 0, 0, 0, \\ -1 - j, 0, 0, 0, -1 - j, 0, 0, 0, 1 + j, 0, 0, 0, 1 + j, 0, 0, 0, 1 \\ + j, 0, 0, 0, 1 + j, 0, 0\} \quad (5.3)$$

The 40 MHz subcarrier sequence for HT-LTF is derived from the L-LTF sequence defined in Eq. (4.2), to similarly retain its peak-to-average ratio and autocorrelation properties. The sequence is constructed by first shifting the 20 MHz sequence lower by 32 and a copy is shifted up by 32. Next the equivalent 20 MHz DC subcarriers, -32 and 32, are both filled in with the value 1. The subcarriers near 40 MHz DC [-5, -4, 3, -2, 2, 3, 4, 5] are filled with the values [-1, -1, -1, 1, -1, 1, 1, -1], respectively. The subcarrier sequence is given in Eq. (5.4):

$$HTLTF_{-58,58} = \{1, 1, -1, -1, 1, 1, -1, 1, -1, 1, 1, 1, 1, 1, -1, -1, 1, 1, -1, 1, \\ -1, 1, 1, 1, 1, 1, -1, -1, 1, 1, -1, 1, -1, 1, -1, -1, -1, -1, -1, -1, \\ -1, 1, 1, -1, -1, 1, -1, 1, -1, 1, 1, 1, 1, 1, -1, -1, -1, 1, 0, 0, 0, -1, \\ 1, 1, -1, 1, 1, -1, -1, 1, 1, -1, 1, 1, 1, 1, 1, 1, -1, -1, 1, 1, \\ -1, 1, -1, 1, 1, 1, 1, 1, -1, -1, 1, 1, -1, 1, -1, -1, -1, -1, \\ -1, -1, 1, 1, -1, -1, 1, -1, 1, -1, 1, 1, 1\} \quad (5.4)$$

As previously mentioned, there are six pilot subcarriers in 40 MHz. The pilot subcarriers are inserted at subcarrier locations -53, -25, -11, 11, 25, 53, as given in Eq. (5.5) (IEEE, 2007):

$$P_{i_{STS}}^{(-58,58,n)} = \{0, 0, 0, 0, 0, \psi_{i_{STS},n \oplus 6}^{(N_{STS})}, 0, 0, 0, 0, 0, 0, 0, 0, 0, 0, 0, 0, 0, 0, 0, 0, \\ 0, 0, 0, 0, 0, 0, 0, \psi_{i_{STS},(n+1) \oplus 6}^{(N_{STS})}, 0, 0, 0, 0, 0, 0, 0, 0, 0, 0, 0, 0, 0, \\ \psi_{i_{STS},(n+2) \oplus 6}^{(N_{STS})}, 0, 0, 0, 0, 0, 0, 0, 0, 0, 0, 0, 0, 0, 0, 0, 0, 0, 0, \\ \psi_{i_{STS},(n+3) \oplus 6}^{(N_{STS})}, 0, 0, 0, 0, 0, 0, 0, 0, 0, 0, 0, 0, \psi_{i_{STS},(n+4) \oplus 6}^{(N_{STS})}, 0, 0, 0, 0, 0, \\ 0, 0, 0, 0, 0, 0, 0, 0, 0, 0, 0, 0, 0, 0, 0, 0, 0, \psi_{i_{STS},(n+5) \oplus 6}^{(N_{STS})}, \\ 0, 0, 0, 0, 0\} \quad (5.5)$$

Table 5.1 Pilot values for 40 MHz transmission (IEEE, 2007)

N_{ss}	i_{ss}	$\psi_{i_{ss},0}^{(N_{ss})}$	$\psi_{i_{ss},1}^{(N_{ss})}$	$\psi_{i_{ss},2}^{(N_{ss})}$	$\psi_{i_{ss},3}^{(N_{ss})}$	$\psi_{i_{ss},4}^{(N_{ss})}$	$\psi_{i_{ss},5}^{(N_{ss})}$
1	0	1	1	1	-1	-1	1
2	0	1	1	-1	-1	-1	-1
2	1	1	1	1	-1	1	1
3	0	1	1	-1	-1	-1	-1
3	1	1	1	1	-1	1	1
3	2	1	-1	1	-1	-1	1
4	0	1	1	-1	-1	-1	-1
4	1	1	1	1	-1	1	1
4	2	1	-1	1	-1	-1	1
4	3	-1	1	1	1	-1	1

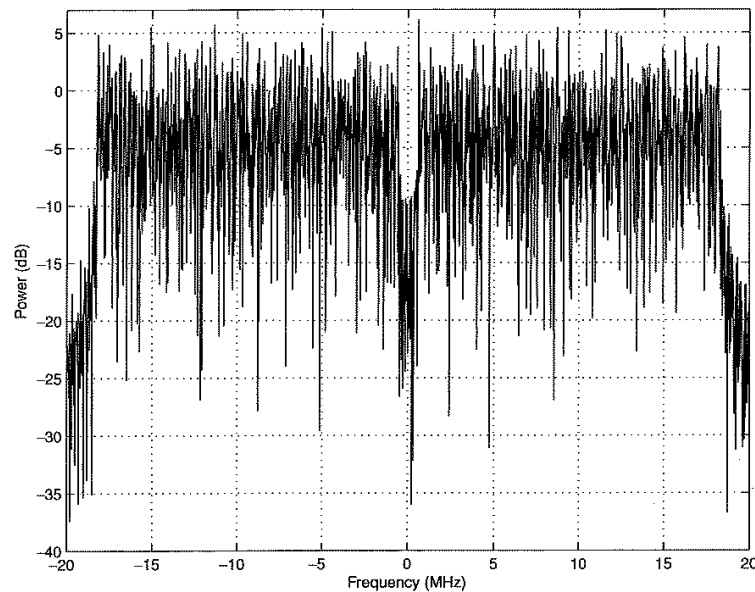


Figure 5.6 Spectrum of the HT portion of the 40 MHz MF preamble.

where $n \oplus a$ indicates symbol number modulo integer a and the patterns $\psi_{i_{ss},n}^{(N_{ss})}$ are defined in Table 5.1.

An illustration of the spectrum of the HT portion of the 40 MHz MF preamble is given in Figure 5.6. As can be seen, in the HT portion of the preamble and data most of the null subcarriers in the area around DC are filled with data subcarriers for higher spectral efficiency.

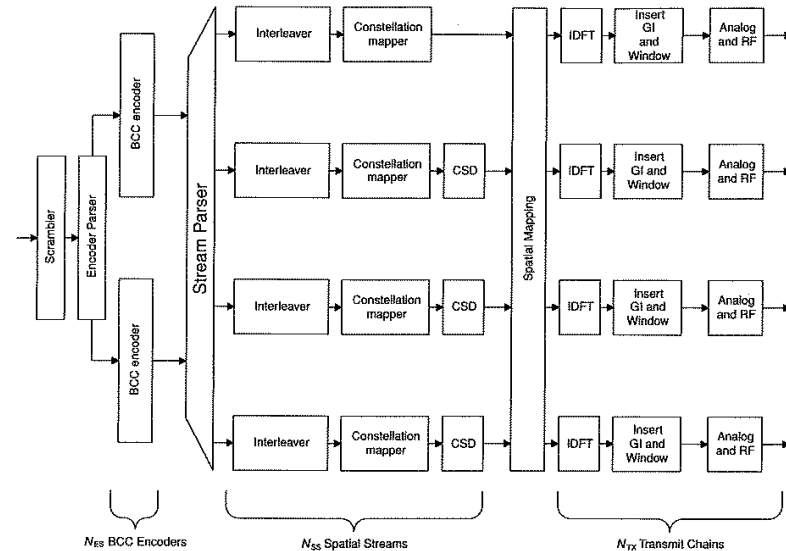


Figure 5.7 Transmitter block diagram for HT Data field with two encoders. Reproduced with permission from IEEE (2007) © IEEE.

To accommodate transmitting with more antennas than spatial streams, Eq. (5.2) may be modified to incorporate a spatial mapping matrix, which maps spatial streams to antennas, as described in Section 4.2.3.8.

5.1.4 40 MHz data encoding

The primary difference in data encoding between 20 MHz and 40 MHz is that all 20 MHz MCSs use a single encoder but some of the 40 MHz MCSs use two encoders. It was decided during 802.11n standard development that convolutional encoding and decoding of data rates above 300 Mbps with a single encoder and decoder carries an excessive implementation burden. Therefore, when the selected MCS has a data rate greater than 300 Mbps, computed based on the 800 ns guard interval, two encoders are used. With two encoders, a number of functions need to be modified. The transmitter block diagram for the HT Data field with two encoders is given in Figure 5.7.

The MCS parameters and data rates for 40 MHz, equal modulation, and 800 ns guard interval are defined in Tables 5.14–5.17 in Appendix 5.2, for one to four spatial streams, respectively. The tables indicate which require two encoders. For MCSs that only require one encoder, the data encoding is identical to 20 MHz and described in Section 4.2.3. The following sections describe the modifications necessary with two encoders.

5.1.4.1 Bit string with two encoders

With two BCC encoders, twelve tail bits, all zeros, are appended to the PSDU. The number of OFDM symbols in the Data field is as follows:

$$N_{\text{SYM}} = \left\lceil \frac{8 \cdot \text{length} + 16 + 12}{N_{\text{DBPS}}} \right\rceil \quad (5.6)$$

where length is the value of the Length field in octets in the HT-SIG field illustrated in Figure 4.16; and N_{DBPS} is the number of data bits per OFDM symbol determined by the selected MCS.

The symbol $\lceil x \rceil$ denotes the smallest integer greater than or equal to x , corresponding to the ceiling function. The number of pad bits is therefore $N_{\text{SYM}} \cdot N_{\text{DBPS}} - 8 \cdot \text{length} - 16 - 12$.

5.1.4.2 Scrambling, encoder parsing, and encoding with two encoders

The resulting data bit string is scrambled with the same scrambler as 802.11a, described in Section 4.1.4. When two BCC encoders are used, the scrambled bits are divided between two encoders by sending alternate bits to each encoder. The i th bit is sent to the j th encoder, where b is the data bit string, as follows in Eq. (5.7) (IEEE, 2007):

$$x_i^{(j)} = b_{2i+j} \quad ; \quad j \in \{0, 1\}, \quad i \geq 0 \quad (5.7)$$

Equation (5.8) gives the tail bits which are replaced with zeros (IEEE, 2007):

$$x_i^{(j)} : j \in \{0, 1\} \quad ; \quad \frac{\text{length} \cdot 8 + 16}{2} \leq i \leq \frac{\text{length} \cdot 8 + 16}{2} + 5 \quad (5.8)$$

Each scrambled, parsed bit string is encoded with the rate $1/2$ BCC. After encoding, each encoded bit string is punctured to the selected code rate. Both encoding and puncturing are the same as 802.11a and described in Section 4.1.5.

In the receiver, the encoder parsing operation is reversed. In this, bits at the output of decoding are alternately selected to create the bit stream for the descrambler.

5.1.4.3 Stream parsing with two encoders

After encoding and puncturing, the two bit strings are re-arranged into a new set of bit strings equal to the number of spatial streams (N_{SS}). To simplify the description, this section only describes the stream parsing function for equal MCSs. For unequal MCSs, refer to Section 12.4. The output of the stream parser is N_{SS} bit strings, each of length N_{CBPSS} (coded bits per spatial stream).

With two encoders, a block of $s \cdot N_{\text{SS}}$ bits (s defined in Eq. (4.23) as $\max\{1, N_{\text{BPSS}}/2\}$) from the output of each encoder is alternately used. The output of the two encoders is given by $y_i^{(j)}$, with bit i from encoder j . The new equation for the output of the stream parser with two encoders in Eq. (5.9) is modified from Eq. (4.24) as follows. In Eq. (4.24), y is a function of the bit stream index i only; however, in Eq. (5.9), y is also a function of the encoder index j . Furthermore, the sequence i is modified to account for the two encoders by modifying the component $\lfloor k/s \rfloor$ in

Eq. (4.24) to $\lfloor k/2 \cdot s \rfloor$ in Eq. (5.9):

$$\begin{aligned} x_k^{i_{\text{SS}}} &= y_i^{(j)} \\ j &= \left\lfloor \frac{k}{s} \right\rfloor \bmod 2 \\ i &= (i_{\text{SS}} - 1) \cdot s + s \cdot N_{\text{SS}} \cdot \left\lfloor \frac{k}{2s} \right\rfloor + k \bmod s \end{aligned} \quad (5.9)$$

where $1 \leq i_{\text{SS}} \leq N_{\text{SS}}$; $k = 0, 1, \dots, N_{\text{CBPSS}} - 1$; $\lfloor v \rfloor$ is the largest integer less than or equal to v , corresponding to the floor; and $v \bmod w$ is the remainder from the division of the integer v by the integer w .

Once the bit strings are parsed into a new set of bit streams equal to the number of spatial streams, encoding of the data continues equivalently to the steps in Section 4.2.3.

In the receiver, stream parsing operation is reversed. Blocks of s bits from each spatial stream are taken in a round robin fashion and assigned to one decoder then blocks of s bits from each spatial stream are alternately assigned to the input of the other decoder.

5.1.5 MCS 32: High throughput duplicate format

Consider a 20 MHz device with 17 dBm output power. Typically, a 40 MHz device is designed with the same output power to maintain the same cost as a 20 MHz device. However, with the same output power, the 40 MHz device has 3 dB lower power spectral density. Therefore, the 40 MHz range is "3dB worse" than 20 MHz due to wider noise bandwidth.

To compensate for the reduced range in 40 MHz relative to 20 MHz, a special HT duplicate format, MCS 32, was created in 40 MHz. With MCS 32, the signal is duplicated on the upper and lower channel during the Data field. Combining the subcarriers in the upper and lower channel at the receiver improves the required SNR of the HT duplicate format resulting in a comparable maximum range for the lowest data rates of 20 MHz and 40 MHz. Figure 5.8 illustrates the packet error rate for 20 MHz MCS 0, 40 MHz MCS 0, and MCS 32.

The simulation uses a 2×2 antenna configuration for all three cases. 40 MHz, MCS 0 performs 1 dB better at a PER of 1% than 20 MHz, MCS 0 due to frequency diversity gain. With MCS 32, the required SNR for the lowest rate in 40 MHz is improved by 2 dB over MCS 0.

Table 5.2 demonstrates a link budget which compares the range for 20 MHz MCS 0, 40 MHz MCS 0, and 40 MHz MCS 32. For this link budget example, the transmitter power is 15 dBm. The transmit and receive antenna gains are 2 dBi, typical for omnidirectional antennas used on 802.11 devices. Both the transmitter and the receiver have two antennas. A typical receiver noise figure of 6 dB is used. The path loss is derived for channel model D based on Section 3.5.5 for the operating frequency of 5.2 GHz. The received signal strength indication (RSSI) is calculated as follows, with all parameters in dB:

$$\text{RSSI} = P_T + A_T - \text{PL} - \text{SF} + A_R$$

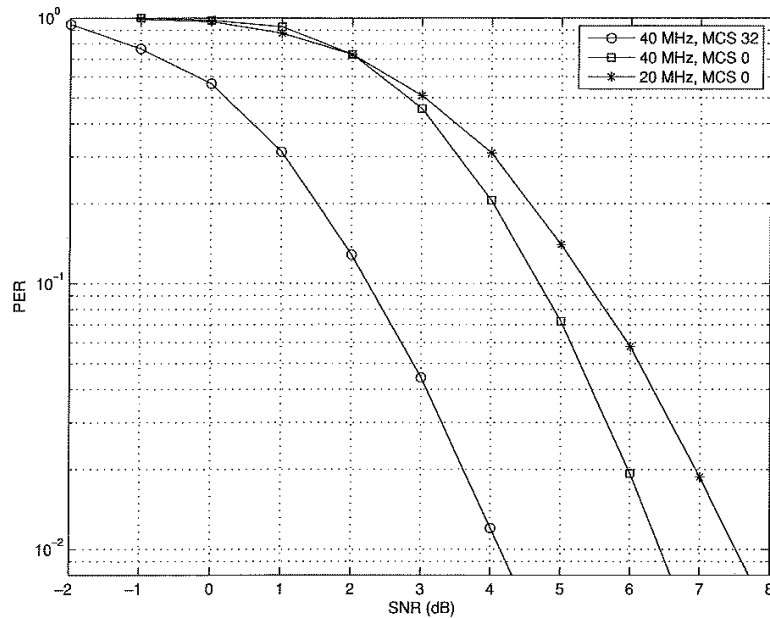


Figure 5.8 Comparison of 40 MHz MCS 32 to MCS 0.

where P_T is the transmit power; A_T , A_R are the transmit or receive antenna gain; PL is the path loss; and SF is shadow fading. With a noise figure of 6 dB, the thermal noise power is -95 dBm for 20 MHz and -92 dBm for 40 MHz. The distance between the transmitter and receiver is calculated to meet the required SNR for channel model D for the specific MCS at a packet error rate of 1%.

As can be seen from the link budget, the range for 40 MHz, MCS 32 and 20 MHz, MCS 0 is almost the same. With MCS 32, the range of 40 MHz is equivalent to the range of 20 MHz.

The data is transmitted on only one spatial stream with BPSK and rate $1/2$ coding. The subcarrier design for the Data field matches that of legacy 802.11a 6 Mbps rate described in Section 4.1.4, and replicates the subcarriers in the lower and upper half of the channel thus maintaining the 6 Mbps data rate for MCS 32. The MCS parameters are given in Table 5.18. During the 802.11n standard development there was discussion of whether to use the 20 MHz HT data format for a higher HT duplicate format data rate. The rationale for using 52 subcarriers of legacy rather than 56 subcarriers of HT 20 MHz is due to the filtering requirements of 40 MHz devices.

Table 5.2 802.11n link budget

Parameter	Units	20 MHz MCS 0	40 MHz MCS 0	40 MHz MCS 32
Tx power	dBm	15	15	15
Tx antenna gain	dBi	2	2	2
Pathloss				
frequency	GHz	5.2	5.2	5.2
distance	m	98.2	86.6	100.8
breakpoint	m	10	10	10
shadow fading	dB	5	5	5
free space	dB	66.8	66.8	66.8
total loss	dB	106.5	104.6	106.9
Rx antenna gain	dBi	2	2	2
RSSI	dBm	-87.5	-85.6	-87.9
Noise power				
NF	dB	6	6	6
BW	MHz	20	40	40
total	dBm	-95.0	-92.0	-92.0
Required SNR	dB	7.5	6.4	4.1

The span of normal 40 MHz format is from subcarrier -58 to 58. The span of the HT duplicate format with replicating the 802.11a subcarrier format is also -58 to 58, requiring the same filtering. If the 20 MHz HT data format was used, however, the span would have been -60 to 60, exceeding that of the standard 40 MHz mode. The 802.11n standard development group decided to not have the HT duplicate format drive the filtering requirements of a 40 MHz device, and the 802.11a subcarrier format was selected as the basis for HT duplicate format.

There are some interesting points about the HT duplicate format waveform that should be highlighted. The legacy portion of the HT MF preamble, the HT-SIG, and the HT-LTF remain unchanged from the standard 40 MHz waveform. This means there are extra subcarriers in the HT-LTF that are not necessarily required for channel estimation. Furthermore, to maximize the SNR on the data subcarriers a scaling factor of 104 was selected. This has the implication of a mismatch in per-subcarrier scaling on the HT-LTF (which has a scaling factor of 114) relative to the Data field. This mismatch needs to be accounted for in the channel estimates or the equalizer. The waveform for the HT duplicate format Data field is given below:

$$r_{\text{HT-DUP}}^{(i_{\text{TX}})}(t) = \frac{1}{\sqrt{104}} \sum_{n=0}^{N_{\text{SYM}}-1} \left[\sum_{k=-26}^{26} (D_{k,n} + p_{n+1}P_k) \exp(j2\pi(k-32)\Delta_F) \right. \\ \times (t - n \cdot T_{\text{SYM}} - T_{\text{GI}}) + j \sum_{k=-26}^{26} (D_{k,n} + p_{n+1}P_k) \exp(j2\pi(k+32)\Delta_F) \\ \left. \times (t - n \cdot T_{\text{SYM}} - T_{\text{GI}}) \right] \quad (5.10)$$

where N_{SYM} is the number of OFDM symbols defined in Eq. (4.22); $(D_{k,n} + p_{n+1}P_k)$ are the data and pilots, defined in Eq. (4.15); $\Delta_F = 312.5$ kHz (40 MHz/128); $T_{\text{SYM}} = 4$ μ s

symbol time, composed of $3.2 \mu\text{s}$ OFDM symbol time and $0.8 \mu\text{s}$ guard interval; and $T_{\text{GI}} = 0.8 \mu\text{s}$ guard interval.

To accommodate transmitting with more antennas than spatial streams, Eq. (5.10) may be modified to incorporate a spatial mapping matrix, which maps spatial streams to antennas, as described in Section 4.2.3.8.

5.1.6 20/40 MHz coexistence with legacy in the PHY

Neighboring legacy APs and clients may be present in the primary or secondary channels of the 40 MHz channel. A common way to reserve the medium for transmission is to send an RTS/CTS. The neighboring APs and clients will respect management frames. To improve efficiency for a 40 MHz device, it is desirable to make the reservation on both the primary and secondary channels simultaneously.

The non-HT duplicate format is designed to be receivable by legacy (and HT) devices in both the primary and secondary channels simultaneously. In essence a packet is transmitted in legacy OFDM format in the upper and lower halves of the 40 MHz channel simultaneously.

The L-STF, L-LTF, and L-SIG are transmitted identically as given by Eq. (5.1). This is followed immediately by the Data field which is transmitted in legacy format similar to the L-SIG. The Data field may be transmitted at any legacy data rate. The equation for the waveform of the Data field is given below:

$$r_{\text{LEG-DUP}}^{(i_{\text{TX}})}(t) = \frac{1}{\sqrt{104}} \sum_{n=0}^{N_{\text{SYM}}-1} \left[\sum_{k=-26}^{26} (D_{k,n} + p_{n+1}P_k) \exp(j2\pi(k-32)\Delta_F) \times (t - n \cdot T_{\text{SYM}} - T_{\text{GI}} - T_{\text{CS}}^{i_{\text{TX}}}) + j \sum_{k=-26}^{26} (D_{k,n} + p_{n+1}P_k) \times \exp(j2\pi(k+32)\Delta_F) (t - n \cdot T_{\text{SYM}} - T_{\text{GI}} - T_{\text{CS}}^{i_{\text{TX}}}) \right] \quad (5.11)$$

where N_{SYM} is the number of OFDM symbols defined in Eq. (4.16); $(D_{k,n} + p_{n+1}P_k)$ are the data and pilots, defined in Eq. (4.15); $\Delta_F = 312.5 \text{ kHz}$ (40 MHz/128); $T_{\text{SYM}} = 4 \mu\text{s}$ symbol time, composed of $3.2 \mu\text{s}$ OFDM symbol time and $0.8 \mu\text{s}$ guard interval; $T_{\text{GI}} = 0.8 \mu\text{s}$ guard interval; and $T_{\text{CS}}^{i_{\text{TX}}}$ is the cyclic shift for transmit antenna chain i_{TX} with values given by Table 4.2.

5.1.7 Performance improvement with 40 MHz

Just based on maximum data rate from the MCS tables, 40 MHz provides more than double the data rate as 20 MHz. In the absence of the required SNR of the individual data rates, it is difficult to determine the actual benefit in a real system. Figure 5.9 illustrates simulation results for over-the-air throughput as a function of range. Over-the-air throughput is defined as the data rate multiplied by one minus the packet error rate. MAC overhead is not accounted for in this analysis. The link budget used to convert from SNR to range follows Table 5.2.

Note that the maximum throughput plateaus at the short range. This artifact is due to a maximum SNR of 35 dB. SNRs above 35 dB become more difficult to achieve with

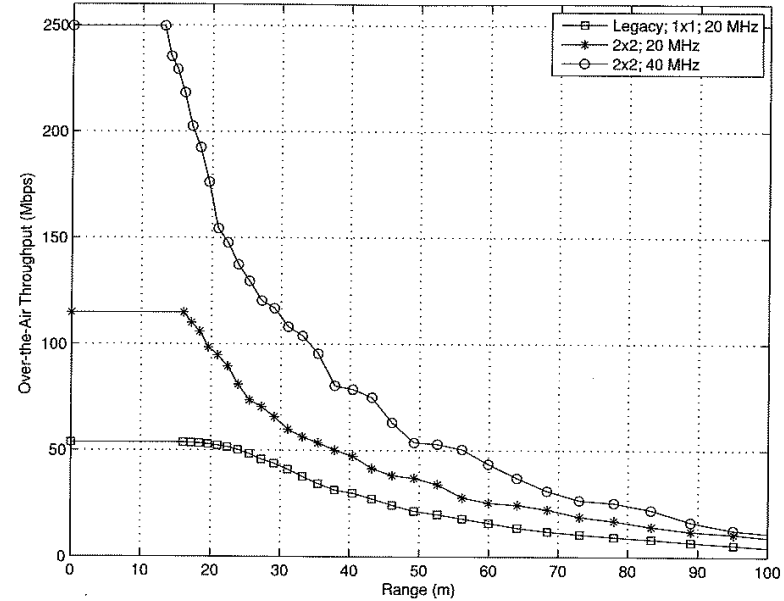


Figure 5.9 Throughput versus range comparing 40 MHz and 20 MHz.

low cost hardware. Therefore throughput values requiring greater than 35 dB SNR were not considered.

We see that in channel model D, the maximum throughput is achieved at less than 15 m at which point 40 MHz provides more than double the data rate as 20 MHz and more than quadruple the data rate of legacy 802.11a/g. Note that this is not true throughput provided by the wireless device. Actual throughput includes overhead due to PHY preambles and the MAC.

We also observe that 40 MHz has an unexpected benefit of range improvement at moderate data rates. For example, if we examine the ranges at 50 Mbps in Figure 5.9, we see that the range is approximately 25 m for legacy, 40 m for 20 MHz HT, and 55 m for 40 MHz HT. Similarly, at 100 Mbps, the approximate range for 20 MHz HT is 20 m and for 40 MHz HT is 35 m. At 100 Mbps, the 20 MHz HT operates with MCS 14. On the other hand, 40 MHz HT could operate with MCS 5 composed of a single spatial stream. This enables receive diversity gain for 40 MHz HT and increases range. Receive diversity was briefly introduced in Section 3.2, and is discussed in more detail in Section 6.1.

For legacy, the simulation uses one transmit antenna and one receive antenna, typical of 802.11a/g devices. At 50 Mbps, legacy operates with 64 QAM, $r = 3/4$, requiring a

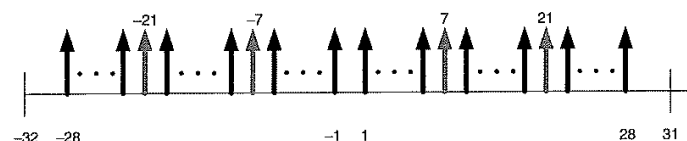


Figure 5.10 20 MHz subcarrier design.

high SNR with the 1×1 system, which then results in a low range. The HT devices operating at 50 Mbps benefit from the diversity gain equivalent to a 1×2 system.

5.2 20 MHz enhancements: Additional data subcarriers

As alluded to in Chapter 4, the number of data subcarriers in 20 MHz was increased to 52, which improves the data rate of 802.11n over legacy 802.11a/g for the same MCS even with single stream. As discussed in Section 4.1, legacy 802.11a/g has 48 data subcarriers. This results in an 8% increase in data rate.

Two data subcarriers are added to each end of the spectrum. The total number of subcarriers is 56. The data subcarriers occupy subcarriers -28 to -22 , -20 to -8 , -7 to -1 , 1 to 6 , 8 to 20 , and 22 to 28 . The pilot subcarriers are located in the same subcarriers as legacy, -21 , -7 , 7 , 21 . Again, as with legacy, there is still a null at DC. The new subcarrier design for 20 MHz HT operation is given in Figure 5.10.

When it was proposed to add two subcarriers at either edge of the band, measurements were performed which demonstrated that the signal still fits within the 802.11a spectral mask with very little extra PA backoff or filtering. The measurements are given in Coffey *et al.* (2004, Slides 26–27). Furthermore, analysis was given in Hedberg (2005), which demonstrated that the adjacent channel interference and Tx and Rx filter requirements are comparable between 56 subcarriers and 52 subcarriers. However, to improve the adjacent channel interference in alternate channels, the floor of the 802.11a mask was tightened to -45 dBr. The new 20 MHz HT spectral mask is given in Figure 5.11.

5.3 MCS enhancements: Spatial streams and code rate

For completeness of the section on high throughput, we must mention the obvious that with MIMO we achieve large data rate increases over 802.11a/g with two, three, and four spatial streams. MIMO provides a data rate increase over 802.11a/g proportional to the number of spatial streams.

In addition, a higher code rate (R) of $5/6$ was added to further increase the data rate by 11%. The SNR requirement with 64-QAM, rate $5/6$ is higher than the highest data rate in 802.11a/g, which only uses 64-QAM, $R = 3/4$. However, with the added robustness

Table 5.3 MCS definition (Peraiah *et al.*, 2006; IEEE, 2007)

MCS indices for 1/2/3/4 spatial streams	Modulation	FEC code rate	Minimum sensitivity (dBm) (20 MHz channel spacing)	Minimum sensitivity (dBm) (40 MHz channel spacing)
0/8/16/24	BPSK	$\frac{1}{2}$	-82	-79
1/9/17/25	QPSK	$\frac{1}{2}$	-79	-76
2/10/18/26	QPSK	$\frac{3}{4}$	-77	-74
3/11/19/27	16 QAM	$\frac{1}{2}$	-74	-71
4/12/20/28	16 QAM	$\frac{3}{4}$	-70	-67
5/13/21/29	64 QAM	$\frac{2}{3}$	-66	-63
6/14/22/30	64 QAM	$\frac{3}{4}$	-65	-62
7/15/23/31	64 QAM	$\frac{5}{6}$	-64	-61

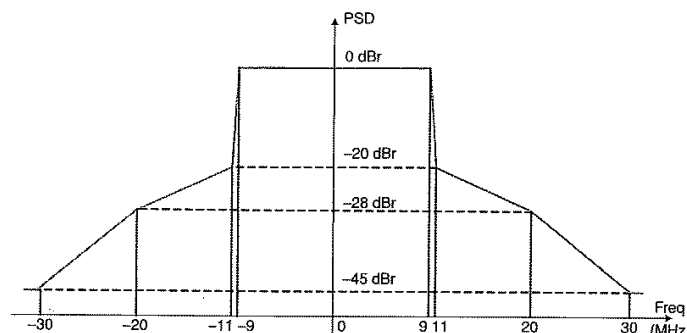


Figure 5.11 20 MHz HT spectral mask. Reproduced with permission from IEEE (2007) © IEEE.

features in 802.11n, the MCSs with 64-QAM, $R = 5/6$ are functional at a reasonable range.

The MCSs for 20 MHz are given in Tables 4.7–4.10, and the MCSs for 40 MHz are given in Tables 5.14–5.18. A summary of the MCSs as they map to modulation and code rate and their corresponding minimum receiver sensitivity is given in Table 5.3. Minimum receiver sensitivity is measured in a conducted test environment at a PER of 10% with packets of length 4096 bytes.

To determine the implementation margin, we simulate the 802.11n MIMO system in an AWGN environment without impairments. This type of system model is designated as Comparison Criteria 59 in Stephens (2004). To generate AWGN results for an $N \times N$ MIMO system, a Fourier matrix is used for the channel. The expression for the Fourier

matrix is given in Eq. (5.12) (Berke, 2004):

$$\mathbf{F}_N = \begin{pmatrix} 1 & 1 & 1 & \dots & 1 \\ 1 & W^1 & W^2 & \dots & W^{(N-1)} \\ 1 & W^2 & W^4 & \dots & W^{2(N-1)} \\ \vdots & \vdots & \vdots & \ddots & \vdots \\ 1 & W^{(N-1)} & W^{(N-1)^2} & \dots & W^{(N-1)^2} \end{pmatrix} \quad (5.12)$$

where $W = \exp(-j2\pi/N)$, and can be expressed explicitly as

$$\begin{aligned} \mathbf{F}_1 &= 1 \\ \mathbf{F}_2 &= \begin{pmatrix} 1 & 1 \\ 1 & -1 \end{pmatrix} \\ \mathbf{F}_3 &= \begin{pmatrix} 1 & 1 & 1 \\ 1 & e^{-j2\pi/3} & e^{-j4\pi/3} \\ 1 & e^{-j4\pi/3} & e^{-j8\pi/3} \end{pmatrix} \\ \mathbf{F}_4 &= \begin{pmatrix} 1 & 1 & 1 & 1 \\ 1 & -j & -1 & j \\ 1 & -1 & 1 & -1 \\ 1 & j & -1 & -j \end{pmatrix} \end{aligned} \quad (5.13)$$

In this manner all receive antennas are coupled with the transmit antennas, yet the matrix is orthogonal allowing for complete separation of the spatial streams without noise enhancement. The capacity of such a system can be derived from Eq. (3.10) as

$$C(\text{bps/Hz}) = N_{\text{SS}} \cdot \log_2 [1 + \text{SNR}] \quad (5.14)$$

and the capacity per spatial stream (SS) is given as

$$C(\text{bps/Hz/SS}) = \log_2 [1 + \text{SNR}] \quad (5.15)$$

Since the capacity per spatial stream is solely a function of SNR, the PER curves are only dependent on modulation, code rate, and SNR. The PER performance with no impairments with packets of length 4096 bytes for each modulation and code rate, as a function of SNR, is given in Figure 5.12.

Extracting the required SNR at 10% PER from Figure 5.12 and assuming a receiver noise figure of 6 dB, we can calculate the margin relative to the minimum sensitivity as given below:

$$\begin{aligned} \text{Margin (dB)} &= \text{RecSen(dBm)} - 30 \\ &\quad - [\text{SNR(dB)} + (10 \cdot \log(1.38 \times 10^{-23} \cdot 290) \\ &\quad + 10 \cdot \log(\text{BW(Hz)}) + \text{NF(dB)})] \end{aligned} \quad (5.16)$$

where RecSen is the minimum receive sensitivity from Table 5.3. The margin ranges from 11 to 12 dB, as given in Table 5.4. The margin is the same for 20 MHz and 40 MHz since the 3 dB difference in bandwidth is equivalent to the difference in minimum sensitivity. Such a large margin may seem excessive; however, algorithmic degradations and hardware impairments degrade performance and require sufficient margin for cost effective implementations.

Table 5.4 Margin relative to minimum sensitivity

Modulation	FEC code rate	Required SNR (dB) @ 10% PER	Margin relative to minimum sensitivity (dB)
BPSK	1/2	1	12.0
QPSK	1/2	4	12.0
QPSK	3/4	6.5	11.5
16-QAM	1/2	9.75	11.2
16-QAM	3/4	13	12.0
64-QAM	2/3	17.25	11.7
64-QAM	3/4	18.75	11.2
64-QAM	5/6	19.75	11.2

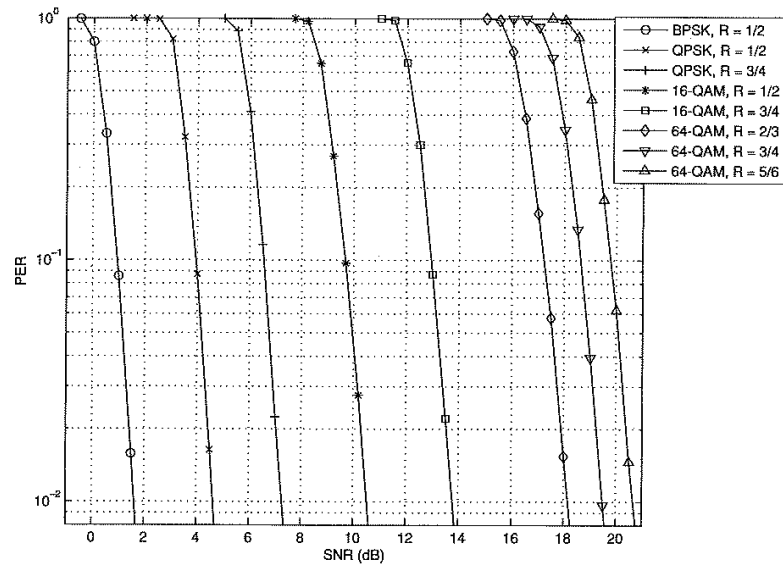


Figure 5.12 MIMO PER performance in AWGN.

The above analysis only provides reference with respect to measurements in a conductive test environment. To get a true sense of the required SNR in a frequency selective fading MIMO environment, we simulate the MCSs in the channel models described in Section 3.5 and with physical layer impairments described in Section 3.5.4. The equalizer is based on MMSE. Synchronization, channel estimation, and phase tracking are included in the simulation to give an accurate model for algorithm degradations.

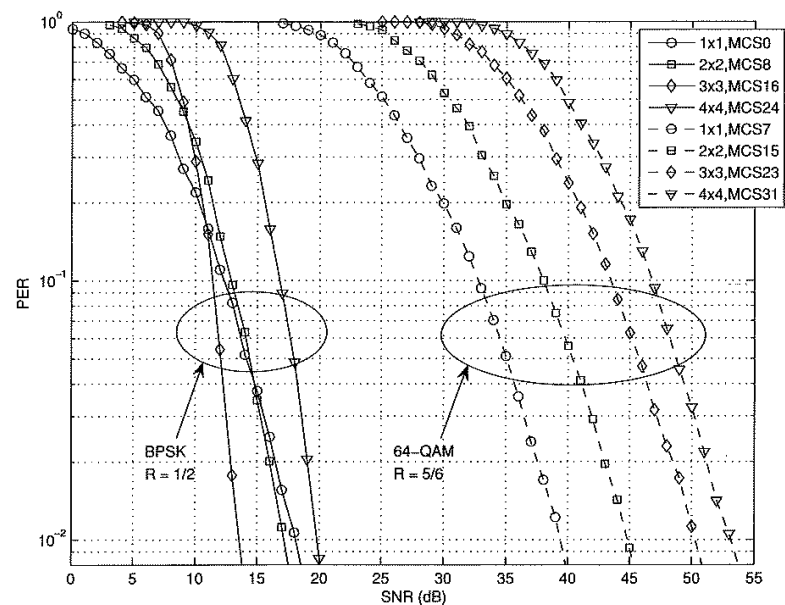


Figure 5.13 Simulated results with channel model B, NLOS, and impairments for 20 MHz.

Figures 5.13–5.15 illustrate the simulated results for 20 MHz in channel models B, D, and E, respectively, in an NLOS environment. The curves with circles are one spatial stream. The curves with squares are two spatial streams. The curves with diamonds are three spatial streams. Finally, the curves with triangles are four spatial streams. All curves have transmit and receive antennas equal to the number of spatial streams. The solid lines correspond to MCSs with BPSK and $R = 1/2$. The dotted lines correspond to MCSs with 64-QAM and $R = 5/6$.

We should note that obtaining an SNR above 35 dB is difficult to achieve with reasonable cost hardware since it exceeds the noise figure, impairments, and implementation losses of typical transmitters and receivers, particularly in RF circuits and ADC/DAC. Required SNRs beyond 35 dB are provided in these figures for comparison purposes only and are not intended to indicate realistic hardware capabilities. In general, links with MCSs of 64-QAM, $R = 5/6$ are difficult to establish without additional means to reduce the required SNR.

We see that for BPSK, $R = 1/2$ the curves are tightly clustered, whereas for 64-QAM, $R = 5/6$ the required SNR substantially increases with the number of spatial streams. With higher order modulation, the system is more sensitive to impairments as the number of spatial streams increases.

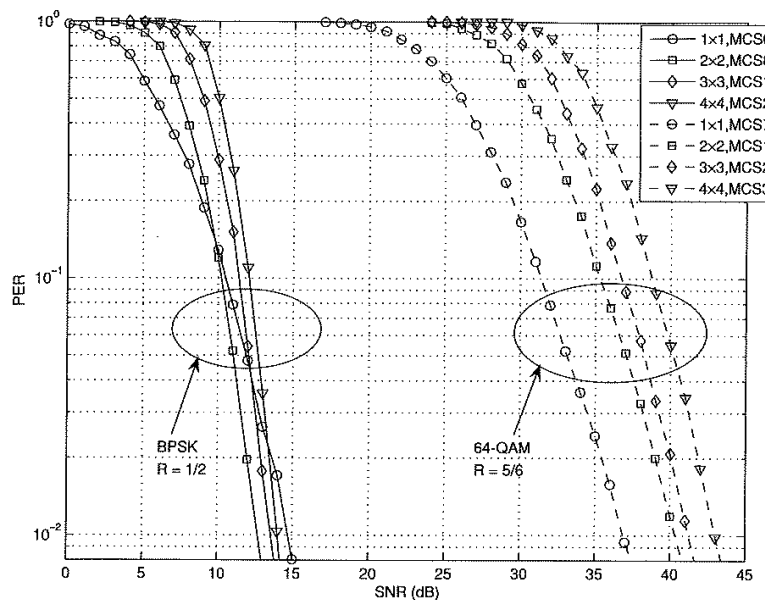


Figure 5.14 Simulated results with channel model D, NLOS, and impairments for 20 MHz.

The required SNR with channel model B is higher than with channel model D. Channel model D has a longer delay spread than channel model B, and therefore more frequency selective fading and less flat fading. The benefit of coded OFDM is its ability to mitigate frequency selective fading, resulting in better performance in channel model D. As seen in Figure 5.15, the performance degrades a little in channel model E as compared to channel model D. The longer delay spread of channel model E begins to effect synchronization and tracking.

5.4

Greenfield (GF) preamble

The MF preamble trades throughput efficiency for interoperability between HT and legacy devices by including the L-STF, L-LTF, and L-SIG in the preamble. During the development of the 802.11n standard amendment, many were of the opinion that network conditions would exist that would benefit from a more efficient preamble. To improve efficiency, a preamble without legacy fields which is not compatible with legacy OFDM devices was adopted.

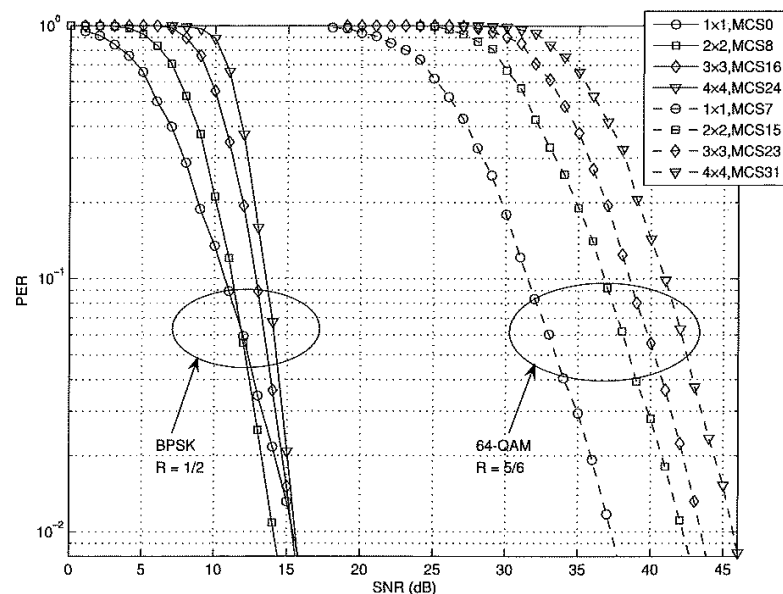


Figure 5.15 Simulated results with channel model E, NLOS, and impairments for 20 MHz.

Environments which do not require legacy compatibility are termed “Greenfield” deployments (from which the name of the preamble was derived). At initial development of 802.11n, adoption and deployment of 802.11a in 5 GHz was slow in much of the world and at that time much of the band was unused. Furthermore, new installations of WLANs in office buildings where no legacy devices were present would not require legacy compatibility. As another example, single family dwellings with no neighbors utilizing legacy devices would also not require legacy compatibility (van Nee and Jones, 2006).

Proponents of GF preamble format also proposed the use of the GF preamble even in the presence of legacy devices. If access to the medium is controlled through MAC protection mechanisms (legacy RTS/CTS, for example), then within the reserved time legacy compatibility is not required. During the reserved time, GF may be used to reduce overhead. MAC protection mechanisms and using the NAV to reserve time are discussed in Section 10.6.

5.4.1 Format of the GF preamble

The fields of the GF preamble are illustrated in Figure 5.16. The MF preamble is also included in Figure 5.16 for direct comparison.

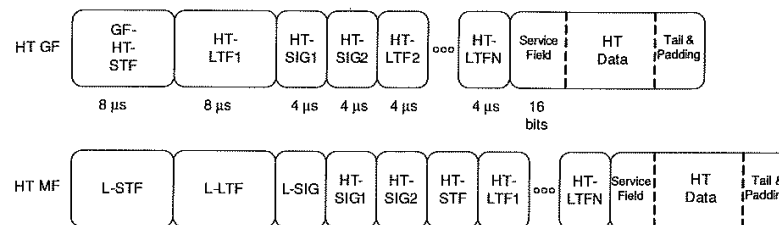


Figure 5.16 Greenfield format preamble.

A new HT-STF for GF (GF-HT-STF) replaces the L-STF in the MF preamble. The waveform for GF-HT-STF is derived in the same manner as the MF HT-STF given in Eq. (4.17), except that it is 8 μ s in length. The increase in length matches the L-STF to provide enough time for start-of-packet detection, AGC setting, frequency estimation, and time synchronization. The HT cyclic shifts given in Table 4.3 are applied to the GF-HT-STF and all subsequent fields in a GF packet. Furthermore, if spatial expansion or beamforming is applied to the spatial streams, the spreading matrix, Q , is applied to the GF-HT-STF and all subsequent fields in a GF packet as defined in Eq. (4.31) for 20 MHz and Eq. (5.2) for 40 MHz.

Since we are not concerned about legacy compatibility with the GF preamble, long HT cyclic shifts are applied to the spatial streams of the GF-HT-STF. With the long cyclic shifts, accurate acquisition and AGC setting required by MIMO receivers is attained. There is no need for a subsequent Short Training field as is required in the MF preamble. This saves 4 μ s compared to the MF preamble.

Without the requirement of legacy compatibility, there is also no need for the L-SIG. All information necessary for 11n is conveyed in the HT-SIG₁ and HT-SIG₂. This saves an additional 4 μ s compared to the MF preamble. With the absence of the L-SIG, the L-LTF is not required. Therefore, HT-LTF1 replaces the L-LTF as compared to MF.

In MF, the two long training symbols of the L-LTF were used for accurate frequency and timing as described in Section 4.1.2. For comparable frequency and time estimation capability, the GF preamble also requires repetition of the long training symbols. Therefore for GF the same Long Training field is used as in MF (given in Eq. (4.19) for 20 MHz and Eq. (5.2) for 40 MHz), except that HT-LTF1 is twice as long as the other HT-LTFs. HT-LTF1 consists of two periods of the long training symbols, preceded by a 1.6 μ s cyclic prefix (IEEE, 2007). This results in a length of 8 μ s.

Following HT-LTF1 in GF are the two HT-SIG fields. The content is identical to that of MF, given in Section 4.2.2.1. The waveform is similar to that of MF (defined by Eq. (4.16) for 20 MHz and Eq. (5.1) for 40 MHz); however, there are some differences which are outlined as follows.

To match the HT-SIGs in GF to HT-LTF1, the HT-SIGs must include the orthogonal mapping matrix. The first column of the orthogonal mapping matrix is applied to the HT-SIGs. Furthermore, the scaling of HT-LTF is based on 56 subcarriers for 20 MHz

and 114 subcarriers for 40 MHz. However, the HT-SIG fields only utilize 52 subcarriers in 20 MHz and 104 subcarriers in 40 MHz to match that of the HT-SIG fields in the MF preamble. The scaling for the HT-SIGs is modified to match the per-subcarrier scaling of the HT-SIG in the GF preamble to HT-LTF1. In addition, in the GF preamble the scaling is based on the number of spatial streams rather than the number of transmit antennas. Therefore in the GF preamble the scaling is $1/\sqrt{N_{SS} \cdot 56}$ for 20 MHz and $1/\sqrt{N_{SS} \cdot 114}$ for 40 MHz for the HT-SIG.

In MF, the first value of the pilot polarity is applied to the L-SIG. In GF, the first two values of the pilot polarity are applied to the HT-SIGs. The same pilot polarity sequence is used as in MF.

The 20 MHz waveform for the GF HT-SIG (including spatial expansion) is given in the equation below:

$$r_{HT-SIG}^{(i_{TX})}(t) = \frac{1}{\sqrt{N_{SS} \cdot 56}} \sum_{n=0}^1 \sum_{k=-26}^{26} \sum_{i_{SS}=1}^{N_{SS}} \left[[Q_k]_{i_{TX}, i_{SS}} \cdot [P_{HTLTF}]_{i_{SS}, 1} \cdot (jD_{k,n} + p_n P_k) \right. \\ \left. \times \exp \left(j2\pi k \Delta_F \left(t - n \cdot T_{SYM} - T_{GI} - T_{CS}^{i_{SS}} \right) \right) \right] \quad (5.17)$$

where N_{SS} is the number of spatial streams; Q is the spatial mapping matrix, discussed further in Section 4.2.3.8; P_{HTLTF} is the orthogonal mapping matrix, as defined by Eq. (4.20); $D_{k,n}$ is the mapping of complex data values to subcarrier locations, which is the same as L-SIG defined by Eq. (4.15); p_n is the pilot polarity defined to be the same as legacy OFDM and given by Eq. (4.5); P_k are the pilot subcarriers defined to be the same as legacy OFDM and given by Eq. (4.4); $\Delta_F = 312.5$ kHz; $T_{SYM} = 4$ μ s symbol time, composed of 3.2 μ s OFDM symbol time and 0.8 μ s guard interval; $T_{GI} = 0.8$ μ s guard interval; and $T_{CS}^{i_{SS}}$ represents the cyclic shift for spatial stream i_{SS} with values given by Table 4.3.

The 40 MHz waveform for the GF HT-SIG (including spatial expansion) is given in the equation below, with the same parameter definitions as in Eq. (5.17):

$$r_{HT-SIG}^{(i_{TX})}(t) = \frac{1}{\sqrt{N_{SS} \cdot 114}} \left\{ \sum_{n=0}^1 \sum_{k=-26}^{26} \sum_{i_{SS}=1}^{N_{SS}} \left[[Q_{k-32}]_{i_{TX}, i_{SS}} \cdot [P_{HTLTF}]_{i_{SS}, 1} \cdot (jD_{k,n} + p_n P_k) \right. \right. \\ \left. \times \exp \left(j2\pi (k-32) \Delta_F \left(t - n \cdot T_{SYM} - T_{GI} - T_{CS}^{i_{SS}} \right) \right) \right] \\ + \sum_{n=0}^1 \sum_{k=-26}^{26} \sum_{i_{SS}=1}^{N_{SS}} \left[[Q_{k+32}]_{i_{TX}, i_{SS}} \cdot [P_{HTLTF}]_{i_{SS}, 1} \cdot (jD_{k,n} + p_n P_k) \right. \\ \left. \times \exp \left(j2\pi (k+32) \Delta_F \left(t - n \cdot T_{SYM} - T_{GI} - T_{CS}^{i_{SS}} \right) \right) \right] \right\} \quad (5.18)$$

Since HT-LTF1 in GF replaced the L-LTF in MF, only the remaining HT-LTFs for two or more spatial streams are needed following the HT-SIG. These are identical to the HT-LTFs in the MF, the expressions of which are given in Eq. (4.19) for 20 MHz and Eq. (5.2) for 40 MHz, respectively. The additional HT-LTFs are 4 μ s in length. The total number of HT-LTFs as a function of the number of spatial streams is the same as MF, and given in Eq. (4.21).

Table 5.5 Preamble length

Number of spatial streams	MF (μ s)	GF (μ s)
1	36	24
2	40	28
3	44	32
4	48	36

For two or more spatial streams, the Data field follows the additional HT-LTFs. For a single spatial stream, there is only the HT-LTF1, therefore the Data field follows the HT-SIGs. The Data field in GF is the same as in MF, as described in Section 4.2.3.

5.4.2 PHY efficiency

The GF preamble reduces the overhead of the PHY preamble by 12 μ s as compared to the MF preamble. The preamble length, as a function of number of spatial streams, is given in Table 5.5.

During the 802.11n standard development process, the proponents of GF claimed substantial improvements in efficiency over MF, especially with higher data rates and shorter packet lengths. The efficiency improvement in the PHY is illustrated in Figure 5.17. Several data rates ranging from 6.5 Mbps with 20 MHz and one spatial stream (SS) to 540 Mbps with 40 MHz and four spatial streams are given. For applications like voice over IP (VoIP), which uses short data packets (of the order of 100 bytes), the benefit of GF preamble would be considerable. Additionally, the time on air is shorter, which means the system can support more callers. This is a benefit since the handheld devices that are the typical VoIP clients are very sensitive to battery life. Shorter time on air reduces the time that the transmitter and receiver are active, which decreases power consumption and increases battery life.

5.4.3 Issues with GF

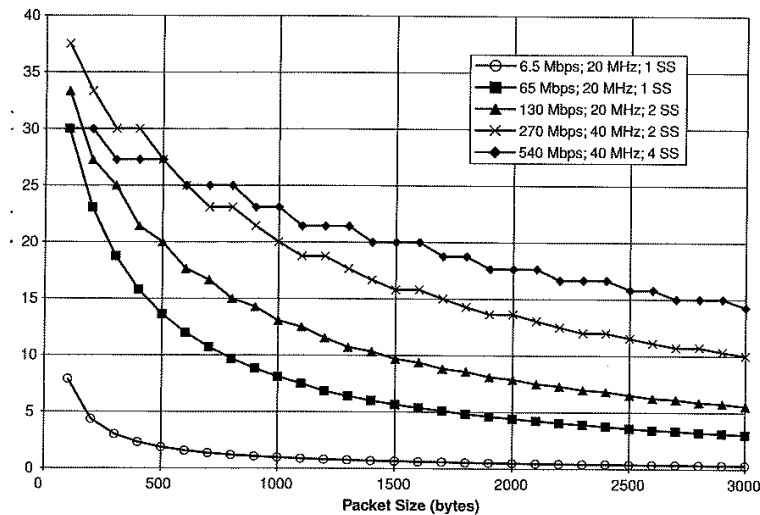
The detractors of the GF preamble argued during the 802.11n standard development process that the GF preamble has many issues and is not necessary (McFarland *et al.*, 2006). These issues include small network efficiency improvement, lack of interoperability with legacy devices, implementation issues, and the added burden of auto-detection between multiple 11n preamble types. In the end, a compromise was reached resulting in the inclusion of the GF preamble in 802.11n, but as an optional preamble format.

5.4.3.1 Network efficiency

During the proposal phase of the 802.11n standard development, a MAC simulation was performed to demonstrate the sensitivity of network throughput to the PHY preamble length (Stephens *et al.*, 2005). The MAC simulations were run over system scenarios for residential, large enterprise, and hot spot as described in Chapter 1 and defined in

Table 5.6 Network throughput versus PHY overhead duration (Stephens *et al.*, 2005)

System scenario	2 TX × 2 RX × 20 MHz			2 TX × 2 RX × 40 MHz		
	25 μ s preamble	50 μ s preamble	Efficiency improvement	25 μ s preamble	50 μ s preamble	Efficiency improvement
Residential	109 Mbps	106 Mbps	3%	169 Mbps	160 Mbps	5%
Large enterprise	103 Mbps	98 Mbps	5%	228 Mbps	226 Mbps	1%
Hot spot	86 Mbps	80 Mbps	7%	194 Mbps	184 Mbps	5%

**Figure 5.17** PHY efficiency improvement of GF over MF.

Stephens *et al.* (2004a). Separate simulations were performed for 20 MHz and 40 MHz, although both assumed devices with two transmit and two receive antennas. The simulations we repeated for different PHY preamble lengths. Table 5.6 provides the results for 25 and 50 μ s preambles, which encompass the range of the length for the MF and GF preambles. We see that the network efficiency only improves by a very small amount.

Further analysis given in McFarland *et al.* (2006) demonstrated small throughput improvement of GF compared to MF preamble with UDP and TCP traffic. The assumptions of the analysis were one AP and one client, ideal traffic patterns in the sense that there was packet bursting with unlimited TXOP lengths, and there was no PHY or MAC packet loss. Results were given for a range of packet lengths, data rates, and both aggregated and un-aggregated. For a detailed discussion on aggregation, see Section 8.2.

Table 5.7 UDP throughput improvement of GF compared to MF without aggregation (McFarland *et al.*, 2006)

Data rate	Single packet exchange		Bursting	
	64 byte packet	1500 byte packet	64 byte packet	1500 byte packet
6.5 Mbps; 20 MHz; 1 SS	3%	0%	4%	0%
65 Mbps; 20 MHz; 1 SS	6%	3%	13%	5%
130 Mbps; 20 MHz; 2 SS	7%	5%	14%	7%
300 Mbps; 40 MHz; 2 SS; Short GI	7%	4%	15%	10%

Table 5.8 Throughput improvement of GF compared to MF with aggregation (McFarland *et al.*, 2006)

Data rate	UDP		TCP	
	20 subframes per packet; 64 bytes per subframe	20 subframes per packet; 1500 bytes per subframe	20 subframes per packet; 64 bytes per subframe	20 subframes per packet; 1500 bytes per subframe
6.5 Mbps; 20 MHz; 1 SS	0%	2%	5%	2%
65 Mbps; 20 MHz; 1 SS	2%	1%	4%	1%
130 Mbps; 20 MHz; 2 SS	3%	0%	2%	1%
300 Mbps; 40 MHz; 2 SS; Short GI	4%	0%	0%	0%

Table 5.7 illustrates the small improvement to UDP traffic without aggregation. Four data rates are given ranging from 6.5 Mbps with 20 MHz and a single spatial stream to 300 Mbps with 40 MHz, two spatial streams, and short guard interval. Two types of simulations are given. The first is a single packet exchange. The second is bursting of data/ACK sequences separated by SIFS with an infinite TXOP. Even without aggregation, the maximum benefit of GF is only 15% throughput improvement.

One of the key MAC features in 802.11n is aggregation to improve efficiency. Table 5.8 illustrates the further reduced benefit of GF compared to MF for UDP and TCP with A-MPDU aggregation. Results are given for two packet lengths. Each packet consists of 20 subframes. Each subframe contains either 64 bytes or 1500 bytes. With aggregation, the maximum benefit of GF compared to MF reduces to 5%.

5.4.3.2 Interoperability issues with legacy

Obviously a legacy device cannot process a GF packet. Therefore, in the presence of legacy devices in the BSS, it is required to protect GF packets. However, this protection does not extend to neighboring BSSs. A GF capable AP with only HT devices associated to it could allow unprotected GF transmissions. Legacy devices in neighboring BSSs would not be protected from GF transmissions.

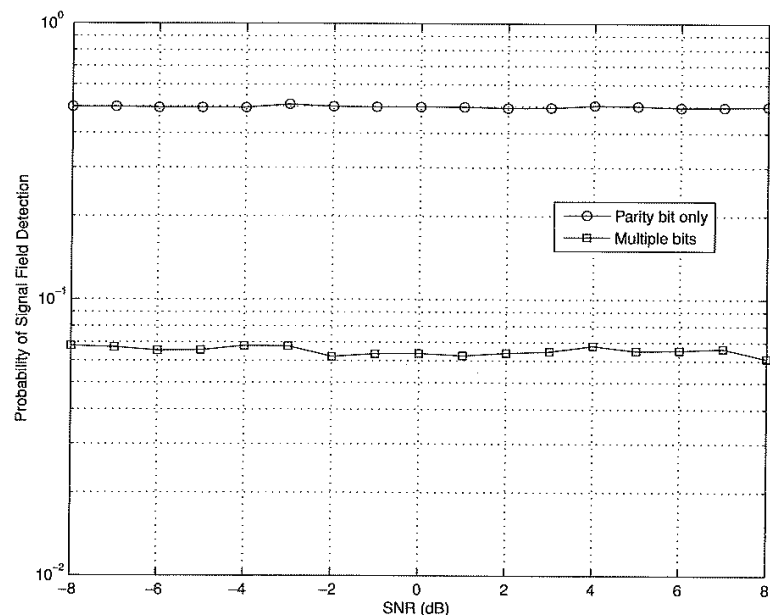


Figure 5.18 False positive detection of Signal field by legacy device with GF transmission in channel model D, NLOS.

Several situations can occur when a legacy device receives a GF packet. Both the GF and legacy preambles have an 8 μ s STF and an 8 μ s LTF followed by the SIG. Since the legacy and GF preambles are similar when the legacy device receives a GF preamble, there is a high likelihood of false detects. This is especially true since there is only a single parity bit in the legacy Signal field, which leads to false positives. Fortunately, many legacy devices use extra fields in the Signal field to reduce the false detect rate. Some possible techniques are described in Section 4.1.3. Figure 5.18 illustrates simulation results that demonstrate the probability that a legacy device falsely detects a valid Signal field when receiving a GF transmission. When only using the single parity bit, the false detect probability is 50%. However, when checking additional bits in the Signal field for valid conditions the probability is reduced to 7%.

False detects lead to legacy devices staying off the air for a random amount of time. If the amount of time is too short, the legacy device interferes with the HT GF device. If the amount of time is too long, the legacy device is unfairly disadvantaged.

Even when the legacy device properly determines that the received signal is not a legacy packet, it only relies on the energy level of the received signal to determine how

long to stay off the air. Based on the standard, the legacy device is only required to remain off the air for a received level of -62 dBm or greater. Therefore, HT GF devices will most likely be interfered with by legacy devices in neighboring BSSs.

5.4.3.3 Implementation issues

There are some minor implementation issues with the GF preamble. Since with one spatial stream the Data field follows the HT-SIG field and with two or more spatial streams the HT-LTFs follows the HT-SIG, it is necessary to rapidly decode the HT-SIG to determine the number of HT-LTFs which dictates what field comes next (McFarland *et al.*, 2006). In addition, the GF preamble uses long cyclic shifts on the STF, which requires extra care in start-of-packet and time synchronization algorithms (McFarland *et al.*, 2006).

5.4.4 Preamble auto-detection

Having two HT preamble formats requires auto-detection between MF and GF formats. Since MF is mandatory and GF is optional, obviously a GF-capable device must auto-detect between MF and GF. However an MF-only device is also required to partially process a GF packet.

The 802.11n standard amendment states, "An HT STA that does not support the reception of an HT Greenfield format packet shall be able to detect that an HT Greenfield format packet is an HT transmission (as opposed to a non-HT transmission). In this case the receiver shall decode the HT-SIG and determine if the HT-SIG cyclic redundancy check (CRC) passes" (IEEE, 2007). Therefore, even a device that does not support GF must process a GF packet up to the HT-SIG and check the CRC. All HT devices need to auto-detect for legacy, HT MF, and HT GF.

The timing of the fields of the legacy, MF, and GF preambles is given in Figure 5.19. Upon receiving a packet, a device must parallel process all the different types of preambles until the correct format is determined. During the time from 0 to 16 μ s, a device must parallel process the legacy STF/LTF for legacy and MF and the HT STF/LTF for GF. From 16 to 20 μ s, the device must process the legacy SIG for legacy and MF preambles. From 16 to 24 μ s, the device must process the HT-SIGs for a GF preamble.

As was described in Section 4.2.2.1, a 90 degree rotation between the L-SIG and HT-SIG₁ may be used to auto-detect between legacy and HT MF. For legacy and GF there is no rotation between 16 and 20 μ s and 20 and 24 μ s. Between 16 and 24 μ s, the HT device must process the HT-SIG₁ and HT-SIG₂ in the case of a GF preamble. Similarly, between 20 and 28 μ s, the HT device must process the HT-SIG₁ and HT-SIG₂ in the case of MF. The combinations of rotations, parity check on L-SIG, and CRC checks on HT-SIG at different times allows differentiation between legacy, HT MF, and HT GF.

If it has been determined that the packet is neither MF nor GF (failed CRC checks and improper rotations), but the parity check of the L-SIG passes, the packet is processed as

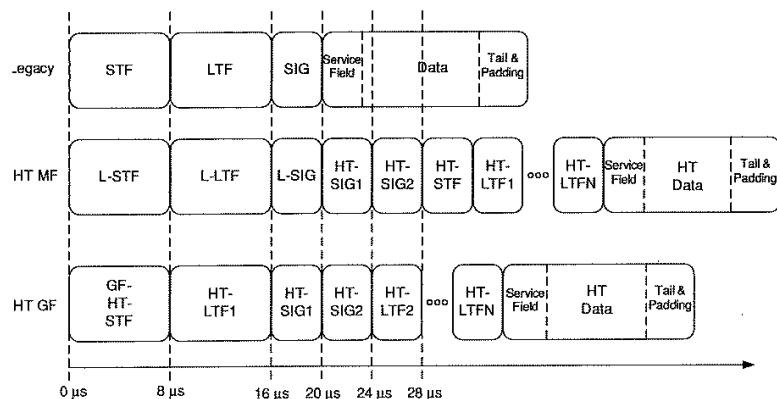


Figure 5.19 Timing of the preamble fields in legacy, MF, and GF.

legacy OFDM. On the other hand, if it has been determined that the packet is GF, but the device is non-GF capable, the device must still properly defer transmission.

The proponents of GF wanted to make it mandatory, thereby ensuring interoperability between all HT devices. The opponents to GF did not feel that the marginal benefit of GF warranted the added complexity of an additional preamble type. As a compromise, a non-GF capable device is given two options when receiving a GF packet. In the first option, the device may extract the Signal field parameters and calculate the transmit time (TXTIME) of the packet. The PHY of the device indicates to the MAC that the channel is busy for this duration. This is the clear channel assessment (CCA). The second option is that the PHY of the non-GF capable device maintains CCA busy until the received level drops below -72 dBm for 20 MHz and -69 dBm for 40 MHz. These thresholds are derived from the receiver minimum input level sensitivity (-82 dBm for 20 MHz and -79 dBm for 40 MHz) plus 10 dB. This allows the non-GF capable device to not have to extract the contents of the HT-SIG when it occurs during a GF packet. A non-GF capable device is only required to check the CRC of a GF packet, which does not require extracting the contents.

A GF capable device must receive all packet formats, and maintain CCA busy based on TXTIME calculated from the contents of the Signal field. The correct duration must be calculated even for rates and modes that are not supported by the device. The TXTIME equation for the GF preamble is given below:

$$\text{TXTIME} = T_{\text{GF-HT-STF}} + T_{\text{HT-LTF1}} + T_{\text{HT-SIG}} + (N_{\text{LTF}} - 1) \cdot T_{\text{HT-LTFs}} + T_{\text{SYM}} \cdot N_{\text{SYM}} \quad (5.19)$$

where $T_{\text{HT-GF-STF}}, T_{\text{HT-LTF1}}, T_{\text{HT-SIG}} = 8 \mu\text{s}$ duration; $T_{\text{HT-LTFs}} = 4 \mu\text{s}$ duration; $T_{\text{SYM}} = 4 \mu\text{s}$ duration for 800 ns guard interval operation; N_{SYM} is the number of OFDM symbols in the Data field defined by Eq. (4.22) for one BCC encoder and Eq. (5.6)

for two BCC encoders; and N_{LTF} is the number of long training symbols, determined from the MCS – the number of long training symbols is based on the number of spatial streams indicated by the MCS (excluding a sounding packet).

The computation of TXTIME changes for short guard interval, sounding packet, LDPC, and STBC, which are described in Sections 5.5, 12.6, 6.4, and 6.3, respectively.

There are some exceptions, which also apply to a non-GF capable device receiving a MF packet. An exception to the TXTIME computation is if the reserved bit is decoded as 0. If the HT-SIG CRC is valid, then a reserved bit value of 0 indicates an invalid condition or a false positive of the CRC computation. Furthermore, if the MCS field is in the range 77–127 or the STBC field has a value of 3, TXTIME can not be computed. In these conditions, the device must not access the medium for as long as the received signal level remains above -62 dBm for 20 MHz and -59 dBm for 40 MHz, rather than based on TXTIME computation. These thresholds are derived from the receiver minimum input level sensitivity (-82 dBm for 20 MHz and -79 dBm for 40 MHz) plus 20 dB.

For further information on the preamble detection, refer to Section 4.2.4.3.

5.5 Short guard interval

The standard OFDM data symbol for 802.11n has a $4 \mu\text{s}$ duration, comprising a $0.8 \mu\text{s}$ guard interval (GI) and $3.2 \mu\text{s}$ of data. This is the same as it is for legacy OFDM symbols. To further increase the data rate in 802.11n, an optional short GI was adopted. With the short GI, the guard interval of the data symbols is reduced to $0.4 \mu\text{s}$ and the overall symbol length is $3.6 \mu\text{s}$. The guard interval of the preamble was not modified. This corresponds to an increase in data rate of 11%. With the short GI, the maximum data rate for 802.11n with 20 MHz is 289 Mbps and with 40 MHz is 600 Mbps.

The waveform equations for the Data field must be changed such that the data symbol duration, T_{SYM} , is reduced to $3.6 \mu\text{s}$ and the guard interval time, T_{GI} , is reduced to $0.4 \mu\text{s}$. This occurs in Eq. (4.30) for MF and 20 MHz operation. Then with spatial expansion in Eq. (4.31), the same changes are made, but only for the Data field. Similarly for 40 MHz, the same changes (limited to the Data field) are made to Eq. (5.2). And finally, Eq. (5.10) for the HT duplicate is also modified with reduced values when short GI is utilized. Correspondingly, the TXTIME computation for the MF preamble in Eq. (4.32) and the GF preamble in Eq. (5.19) is modified by also reducing T_{SYM} to $3.6 \mu\text{s}$.

There is one packet format in which short GI may not be used. With the GF preamble and one spatial stream, the data symbols immediately follow the HT-SIG. However, the receiver does not know that this situation has occurred until it parses the HT-SIG. In a typical design, the receiver continues processing the subsequent symbol in parallel to extracting the information from the HT-SIG. The first step is to remove the guard interval before performing the Fourier transform. In order to receive a GF with one spatial stream and short GI, the receiver would have to buffer the signal until it determined the number of spatial streams and the length of the GI from the HT-SIG. In order to avoid this

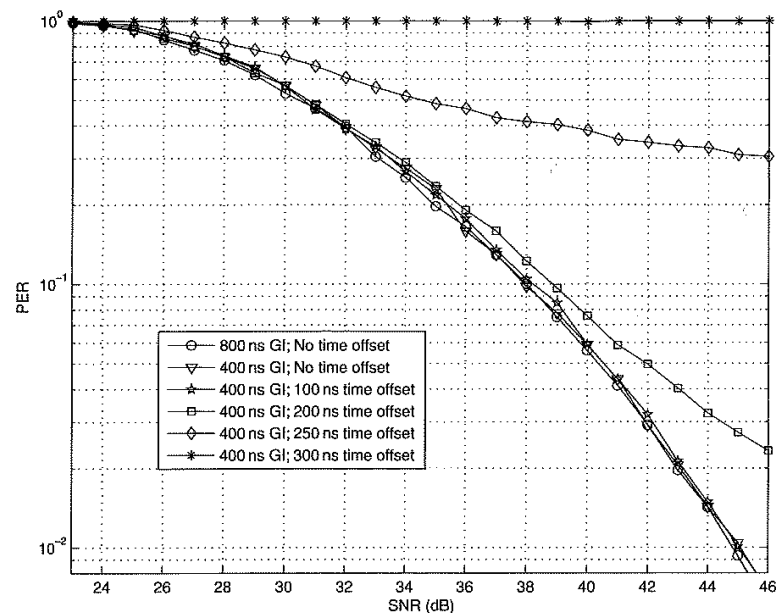


Figure 5.20 Sensitivity to time offset with MCS 15, 2×2 , in channel model B, NLOS.

complication, the use of short GI with GF with one spatial stream is not permitted by the 802.11n standard.

As discussed in Section 2.1, the guard interval is used to mitigate inter-symbol interference due to the delay spread of the channel environment. By shortening the GI, the signal is more susceptible to longer delay spread channels. Delay spread is also caused by filtering in the transmitter and receiver. Transmit filters are used to reduce out-of-band emissions. The larger the reduction of interference to adjacent bands, the longer the delay spread of the filter. Similarly on receive, the more there is rejection of adjacent bands, the greater the delay spread that is incurred from the receive filter.

By shortening the guard interval, the receiver requires much more precise time synchronization, especially in channels with longer delay spread. Therefore, one way to quantify the receiver sensitivity to shortening of the guard interval is to model the receiver performance with increasing fixed time offsets in the time synchronization. We compare the performance with 800 ns GI to 400 ns GI. Figure 5.20 illustrates the performance for MCS 15 with two transmit antennas and two receiver antennas in channel model B, with NLOS conditions. Figures 5.21 and 5.22 illustrate results with the same configuration,

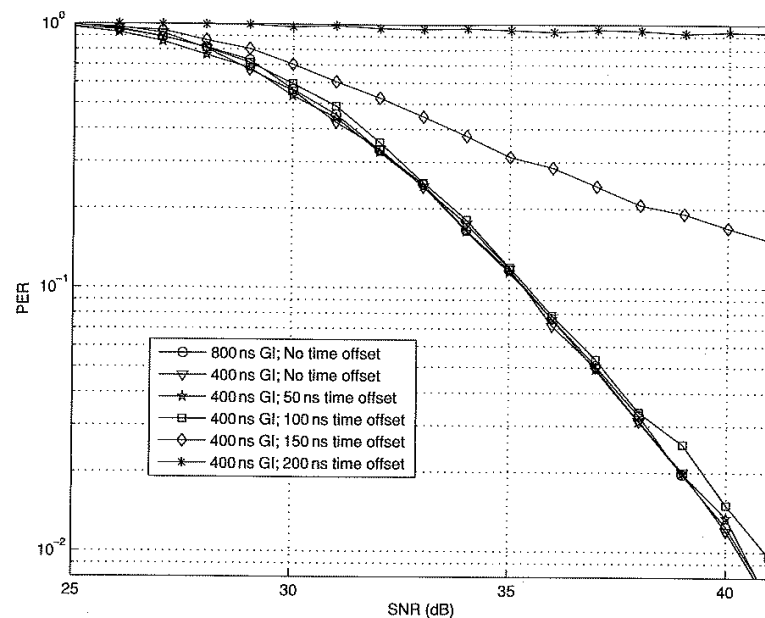


Figure 5.21 Sensitivity to time offset with MCS 15, 2×2 , in channel model D, NLOS.

except that the simulation is performed in channel models D and E, respectively, with NLOS conditions.

Though not illustrated in the figures, packets transmitted with 800 ns GI experience no degradation with time offsets up to 300 ns with channel models B, D, and E. On the other hand, with 400 ns GI the performance degradation is directly coupled to the delay spread of the channel. In channel model B, which has an RMS delay spread of 15 ns, there is no degradation up to a time offset of 100 ns. As the offset is increased to 200 ns, slight degradation is experienced. The receiver fails when the time offset is increased to 250 ns or more. In channel model D which has an RMS delay spread of 50 ns, there is no degradation with a time offset of 50 ns. There is a slight degradation with a time offset of 100 ns. There is a large degradation with a time offset of 150 ns, and the receiver fails with an offset greater than 150 ns. In channel model E which has an RMS delay spread of 100 ns, there is also no degradation with a time offset of 50 ns. However with a time offset of 100 ns there is more degradation than in channel model D. There is a large degradation with a time offset of 150 ns, and the receiver fails with offset greater than 200 ns.

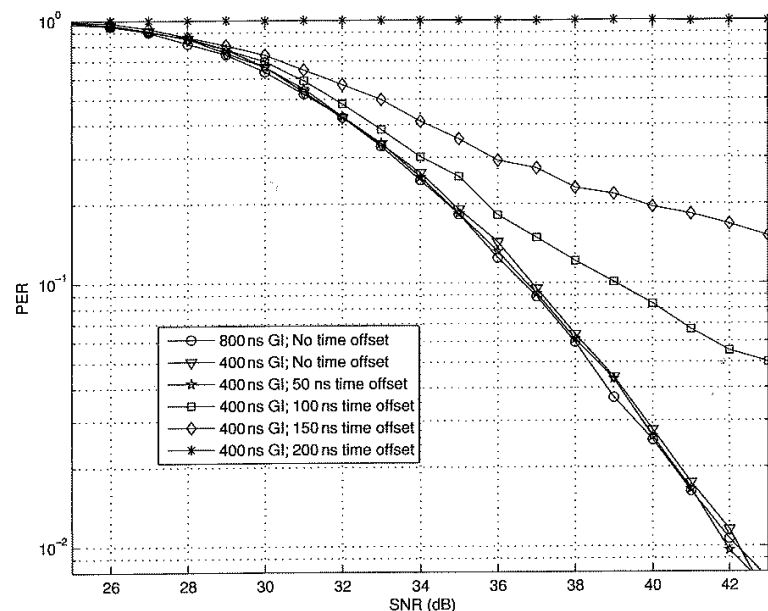


Figure 5.22 Sensitivity to time offset with MCS 15, 2×2 , in channel model E, NLOS.

To summarize, noticeable degradation occurs at time offsets which decrease from 200 ns for channel model B to 150 ns for channel model D to 100 ns for channel model E. This decrement roughly corresponds to the increase in RMS delay spread between the channels. When critically sampled, a time offset of 100 ns for channel model E is only two samples in time, which requires much more accuracy in the time synchronization and tracking algorithm in the receiver.

Short GI adds another dimension to rate adaptation beyond modulation, code rate, and number of spatial streams. With each added dimension, the complexity of rate adaptation increases significantly. In addition, short GI only provides an incremental improvement in data rate in shorter delay spread channels. Similar incremental improvements may be attained by changing the MCS. Therefore, a typical approach to the use of short GI is to only try to switch to short GI when the highest data rate of the device with long GI has been achieved. The highest MCS is indicative of a good channel environment and no further increases in data rate can be achieved by modulation, code rate, and number of spatial streams. As a final rate adaptation step, the device can switch to short GI for the maximum data rate of the device. If acknowledgements are received to the transmitted packets, the device knows that the channel supports short GI and the maximum data

rate. With added complexity, one may determine the delay spread of the channel from the channel estimate to try to use short GI in more situations.

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Appendix 5.1: Channel allocation

Channel center frequencies are defined at every integral multiple of 5 MHz beginning from the channel starting frequency for the specific band, as follows:

$$\text{Channel center frequency} = \text{channel starting frequency} + 5 \cdot n_{\text{ch}} \text{ (MHz)} \quad (5.20)$$

where n_{ch} is the channel number ($= 0, 1, \dots, 200$ for 5 GHz, and $= 1, 2, \dots, 13$ for 2.4 GHz). For the 5 GHz band, the channel starting frequency is equal to 5000. For the 2.4 GHz band, the channel starting frequency is equal to 2407.

Tables 5.9 and 5.10 give the 20 MHz channel allocations for USA and Europe, respectively. Tables 5.11 and 5.12 give the 40 MHz channel allocations for USA and Europe, respectively. The 40 MHz channel allocations also include both conditions where the secondary 20 MHz channel is below or above the primary 20 MHz channel. This is indicated by the secondary field.

Table 5.9 20 MHz channel allocation for USA

Frequency band (GHz)	Regulatory class	Channel number	Center frequency (MHz)	Transmit power limit (mW)
ISM 2.4–2.483	12	1, 2, ..., 11	2412, 2417, ..., 2462	1000
U-NII lower 5.15–5.25	1	36, 40, 44, 48	5180, 5200, 5220, 5240	40
U-NII middle 5.25–5.35	2	52, 56, 60, 64	5260, 5280, 5300, 5320	200
Additional U-NII 5.47–5.725	4	100, 104, ..., 140	5500, 5520, ..., 5700	200
Upper U-NII 5.725–5.85	3	149, 153, 157, 161	5745, 5765, 5785, 5805	800
	5	149, 153, 157, 161, 165	5745, 5765, 5785, 5805, 5825	1000

Table 5.10 20 MHz channel allocation for Europe

Frequency band (GHz)	Regulatory class	Channel number	Center frequency (MHz)	Transmit power limit (mW)
ISM 2.4–2.483	4	1, 2, ..., 13	2412, 2417, ..., 2472	100
5.15–5.25	1	36, 40, 44, 48	5180, 5200, 5220, 5240	200
5.25–5.35	2	52, 56, 60, 64	5260, 5280, 5300, 5320	200
5.47–5.725	3	100, 104, ..., 140	5500, 5520, ..., 5700	1000

Table 5.11 40 MHz channel allocation for USA

Frequency band (GHz)	Regulatory class	Primary channel number	Secondary field	Secondary channel number
ISM 2.4–483	32	1	1	5
		2	1	6
		3	1	7
		4	1	8
		5	1	9
		6	1	10
		7	1	11
		5	–1	1
		6	–1	2
		7	–1	3
		8	–1	4
U-NII lower 5.15–5.25	22	9	–1	5
		10	–1	6
		11	–1	7
		36	1	40
		44	1	48
	27	40	–1	36
		48	–1	44
	23	52	1	56
		60	1	64
	28	56	–1	52
		64	–1	60
Additional U-NII 5.47–5.725	24	100	1	104
		108	1	112
		116	1	120
		124	1	128
		132	1	136
	29	104	–1	100
		112	–1	108
		120	–1	116
		128	–1	124
		136	–1	132
Upper U-NII 5.725–5.85	25 (for 800 MW) or 26 (for 1000 MW)	149	1	153
		157	1	161
	30 (for 800 MW) or 31 (for 1000 MW)	153	–1	149
		161	–1	157

Table 5.12 40 MHz channel allocation for Europe

Frequency band (GHz)	Regulatory class	Primary channel number	Secondary field number	Secondary channel number
ISM 2.4–2.483	11	1	1	5
		2	1	6
		3	1	7
		4	1	8
		5	1	9
		6	1	10
		7	1	11
		8	1	12
		9	1	13
	12	5	–1	1
		6	–1	2
		7	–1	3
		8	–1	4
		9	–1	5
		10	–1	6
		11	–1	7
5.15–5.25	5	12	–1	8
		13	–1	9
	8	36	1	40
		44	1	48
5.25–5.35	6	40	–1	36
		48	–1	44
	9	52	1	56
		60	1	64
5.47–5.725	7	56	–1	52
		64	–1	60
		100	1	104
		108	1	112
		116	1	120
		124	1	128
	10	132	1	136
		104	–1	100
		112	–1	108
		120	–1	116
		128	–1	124
		136	–1	132

Table 5.13 Symbols used for MCS parameters (IEEE, 2007)

Symbol	Explanation
R	Code rate
$N_{\text{BPS}}^{\text{CS}}$	Number of coded bits per single carrier for each spatial stream (or modulation order)
N_{CBPS}	Number of coded bits per OFDM symbol
N_{DBPS}	Number of data bits per OFDM symbol
N_{ES}	Number of BCC encoders

Table 5.14 40 MHz MCS parameters for single spatial stream (IEEE, 2007)

MCS index	Modulation	R	N_{ES}	$N_{\text{BPS}}^{\text{CS}}$	N_{CBPS}	N_{DBPS}	Data rate (Mbps)
0	BPSK	$1/2$	1	1	108	54	13.5
1	QPSK	$1/2$	1	2	216	108	27.0
2	QPSK	$3/4$	1	2	216	162	40.5
3	16-QAM	$1/2$	1	4	432	216	54.0
4	16-QAM	$3/4$	1	4	432	324	81.0
5	64-QAM	$2/3$	1	6	648	432	108.0
6	64-QAM	$3/4$	1	6	648	486	121.5
7	64-QAM	$5/6$	1	6	648	540	135.0

Appendix 5.2: 40 MHz basic MCS tables

The following tables consist of the MCSs for 40 MHz “basic” rates. As defined herein, the basic rates include equal modulation and the mandatory 800 ns guard interval. Unequal modulation is described in Section 12.4. Short guard interval is described in Section 5.5.

For 40 MHz, the number of data subcarriers is 108 and the number of pilot subcarriers is 6. The only exception is for HT duplicate format (MCS 32), where the number of data subcarriers is 96 and the number of pilot subcarriers is 8.

Table 5.13 describes the symbols used in the subsequent tables.

Table 5.14 defines the 40 MHz MCS parameters and data rates for single spatial stream transmission. The data rate is calculated by dividing N_{DBPS} by the symbol time of 4 μs (3.2 μs OFDM symbol plus 0.8 μs guard interval).

Table 5.15 defines the 40 MHz MCS parameters and data rates for two spatial stream transmission.

Table 5.16 defines the 40 MHz MCS parameters and data rates for three spatial stream transmission.

Table 5.17 defines the 40 MHz MCS parameters and data rates for four spatial stream transmission.

Table 5.18 defines the 40 MHz MCS parameters and data rates for HT duplicate format.

Table 5.15 40 MHz MCS parameters for two spatial streams (IEEE, 2007)

MCS index	Modulation	R	N_{ES}	N_{BPSCS}	N_{CBPS}	N_{DBPS}	Data rate (Mbps)
8	BPSK	$1/2$	1	1	216	108	27.0
9	QPSK	$1/2$	1	2	432	216	54.0
10	QPSK	$3/4$	1	2	432	324	81.0
11	16-QAM	$1/2$	1	4	864	432	108.0
12	16-QAM	$3/4$	1	4	864	648	162.0
13	64-QAM	$2/3$	1	6	1296	864	216.0
14	64-QAM	$3/4$	1	6	1296	972	243.0
15	64-QAM	$5/6$	1	6	1296	1080	270.0

Table 5.16 40 MHz MCS parameters for three spatial streams (IEEE, 2007)

MCS index	Modulation	R	N_{ES}	N_{BPSCS}	N_{CBPS}	N_{DBPS}	Data rate (Mbps)
16	BPSK	$1/2$	1	1	324	162	40.5
17	QPSK	$1/2$	1	2	648	324	81.0
18	QPSK	$3/4$	1	2	648	486	121.5
19	16-QAM	$1/2$	1	4	1296	648	162.0
20	16-QAM	$3/4$	1	4	1296	972	243.0
21	64-QAM	$2/3$	2	6	1944	1296	324.0
22	64-QAM	$3/4$	2	6	1944	1458	364.5
23	64-QAM	$5/6$	2	6	1944	1620	405.0

Table 5.17 40 MHz MCS parameters for four spatial streams (IEEE, 2007)

MCS index	Modulation	R	N_{ES}	N_{BPSCS}	N_{CBPS}	N_{DBPS}	Data rate (Mbps)
24	BPSK	$1/2$	1	1	432	216	54.0
25	QPSK	$1/2$	1	2	864	432	108.0
26	QPSK	$3/4$	1	2	864	648	162.0
27	16-QAM	$1/2$	1	4	1728	864	216.0
28	16-QAM	$3/4$	2	4	1728	1296	324.0
29	64-QAM	$2/3$	2	6	2592	1728	432.0
30	64-QAM	$3/4$	2	6	2592	1944	486.0
31	64-QAM	$5/6$	2	6	2592	2160	540.0

Table 5.18 40 MHz MCS parameters for HT duplicate format (IEEE, 2007)

MCS index	Modulation	R	N_{ES}	N_{BPSCS}	N_{CBPS}	N_{DBPS}	Data rate (Mbps)
32	BPSK	$1/2$	1	1	48	24	6.0

Appendix 5.3: Physical layer waveform parameters

Table 5.19 provides a summary and comparison of waveform parameters for 802.11a, 802.11n 20 MHz and 40 MHz.

Table 5.19 Physical layer waveform parameters

	802.11a	802.11n 20 MHz	802.11n 40 MHz	802.11n 40 MHz MCS 32	802.11n non-HT duplicate
Data subcarriers	48	52	108	96	96
Unique modulated data values	48	52	108	48	48
Pilot subcarriers	4	4	6	8	8
Subcarrier frequency spacing	312.5 kHz (20 MHz/64)	312.5 kHz (20 MHz/64)	312.5 kHz (40 MHz/128)	312.5 kHz (40 MHz/128)	312.5 kHz (40 MHz/128)
IDFT/DFT period	3.2 μ s	3.2 μ s	3.2 μ s	3.2 μ s	3.2 μ s
Guard interval of data symbols	0.8 μ s	Normal: 0.8 μ s Short: 0.4 μ s	Normal: 0.8 μ s Short: 0.4 μ s	Normal: 0.8 μ s Short: 0.4 μ s	Normal: 0.8 μ s

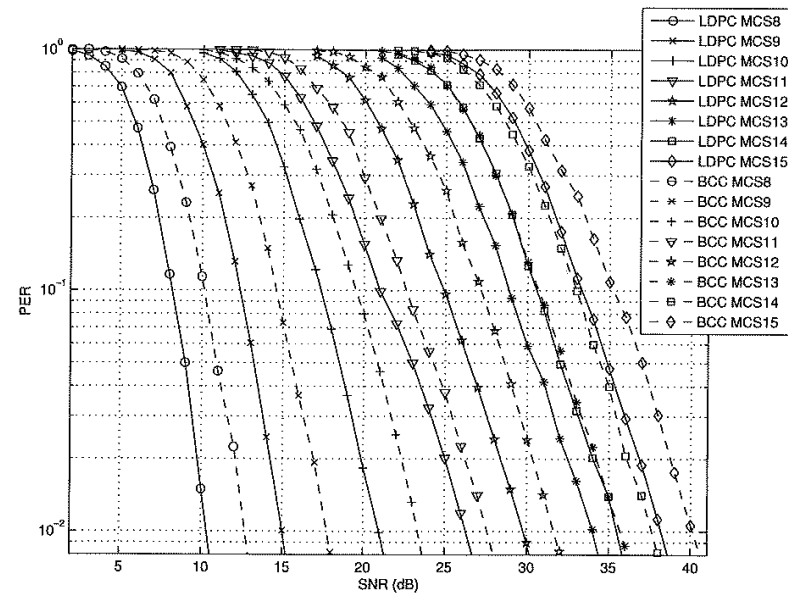


Figure 6.24 Comparison between LDPC and BCC for 2×20 MHz, channel model D, NLOS, and payload length of 970 bytes.

of the actual transmitted information than the number of information code word bits. In many cases the parity bits are punctured, which increases the effective code rate, and must be accounted for in the metric. Repetition in LDPC is very similar to padding in BCC, so we do not account for repetition in the new metric when comparing BCC and LDPC performance. The new metric for effective code rate is given as follows:

$$\tilde{R} = \frac{N_{\text{pld}}}{N_{\text{pld}} + N_{\text{CW}} \cdot N_{\text{LDPC}} \cdot (1 - R) - N_{\text{punc}}} \quad (6.34)$$

Figure 6.23 illustrates the effective code rate for LDPC for MCS 11, 13, 14, and 15 for 20 MHz mode for a range of data packet lengths. The code rates for these MCSs are $1/2$, $2/3$, $3/4$, and $5/6$, respectively. In all four cases, the effective code rate converges to the MCS code rate by a data packet length of 500 bytes. Shorter packets differ the most between the effective code rate and the MCS code rate for all four cases. As we see in Figure 6.20, shorter packets have a much larger number of shortening bits. Therefore, the ratio between payload bits and information bits in a code word is much smaller than the native rate of the code.

6.4.3 LDPC coding gain

Since BCC is the mandatory coding scheme of 802.11n, a device is required to have a BCC encoder and decoder. Using the optional LDPC code means implementing both the BCC and LDPC encoders and decoders. Therefore the basic questions regarding LDPC coding are: (a) how much additional gain does it provide to the system and (b) is it worth having a second encoder/decoder type in the device?

To give some insight into answering these questions, we provide simulation results for LDPC and BCC with two transmit antennas, two receive antennas and 20 MHz mode. The simulations are performed in the conditions detailed at the beginning of the chapter.

For a fair comparison between BCC and LDPC, a payload length of 970 bytes was selected such that the effective code rate of LDPC is very close to the rate of the MCS. The PER versus SNR waterfall curves are illustrated in Figure 6.24. The solid lines are the LDPC curves, and the dashed lines are the BCC curves. The gain of LDPC over BCC ranges from 1.5 dB to 3 dB, at a PER of 1%, depending on the MCS.

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Appendix 6.1: Parity check matrices

Since the parity check matrices are extremely large and difficult to display in print, certain notational short cuts are taken in IEEE (2007) to represent the matrices. This notation does not have anything to do with the design of the parity check matrices.

duration of the TXOP. The Data/ACK exchange provides more limited protection near the transmitter (due to the higher order modulation used for data frames), but is more efficient than the RTS/CTS exchange with which no information transfer takes place.

It should be noted that the block transfer is independent of the TXOP. The block transfer may occur over multiple TXOPs or it may be contained within a single TXOP.

References

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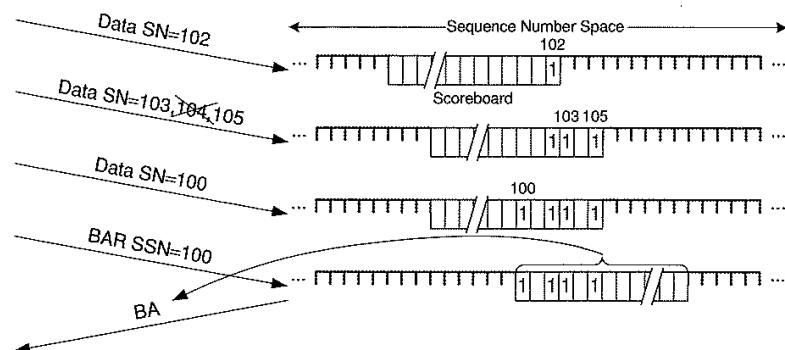


Figure 8.16 Scoreboard operation.

by the same station will occur before data belonging to another block ack session is received by the recipient, erasing the session's block ack scoreboard. The originator can thus simply use a BAR frame in a subsequent channel access to retrieve the ack state. In the unlikely event that the block ack scoreboard is no longer available, the BA will show all zeros and the originator will need to retransmit the MSDUs. Alternatively, if the originator sent a single aggregate consisting of QoS Data frame with Normal Ack policy and no BA was received, then the originator may assume that none of the QoS Data frames made it through and simply retransmit the QoS Data frames.

While more complex behavior is permitted, most implementations will conform to partial state rules by sending a single aggregate in each TXOP, soliciting a BA by setting the ack policy of the QoS Data frames making up the aggregate to Normal Ack. On the rare occasion when an MSDU lifetime expires before it is acknowledged, the originator will send a BAR frame to flush the holes in the reorder buffer. An originator implementing this behavior would work with both a full state and partial state implementation in the recipient.

8.4.4 HT-immediate block ack TXOP sequences

Typical TXOP sequences under HT-immediate block ack are shown in Figure 8.17. The TXOP begins with a short frame exchange as a collision detect mechanism. In these sequences an RTS/CTS exchange, which also provides enhanced protection, is used, but the more efficient Data/ACK exchange could also be used. In sequence (a) a single aggregate data transmission is made with the ack policy set to Normal Ack in the QoS Data MPDUs making up the aggregate. This solicits an immediate BA response.

Sequence (b) is similar except that two aggregate transmissions are sent back-to-back as a burst. This might be done to improve robustness, for example if a single transmission was too long and possibly subject to rapidly changing channel conditions for which a

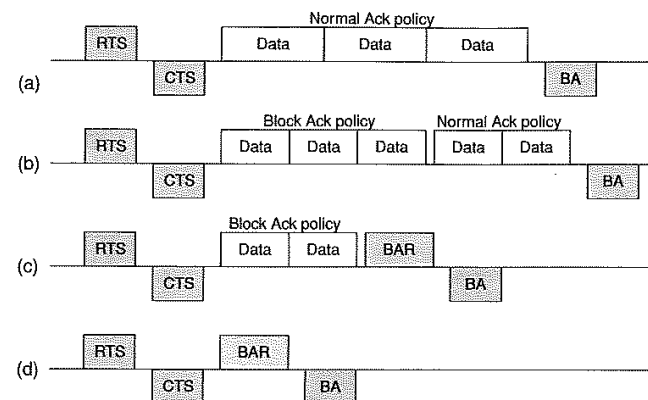


Figure 8.17 Typical HT-immediate TXOP sequences.

new channel estimate would be beneficial. The first aggregate would be sent with Ack Policy set to Block Ack and the second aggregate with Ack Policy set to Normal Ack in order to solicit a BA.

Sequences (c) and (d) may be used when an MSDU is discarded due to lifetime expiry, when the BAR is required to release potentially stalled complete MSDUs in the recipient's reorder buffer. In sequence (c) data frames are available and would typically be sent ahead of the BAR. The aggregate is sent with Ack Policy set to Block Ack since the BAR will solicit the BA response. Note that it is permitted to aggregate the BAR with the data frames; however, the BAR would typically be sent as a separate more robustly modulated frame. Since the BAR is sent infrequently, sequences involving BAR do not need to be optimized as the overall gain is small.

8.5 HT-delayed block ack

HT-delayed block ack is an extension to the delayed block ack protocol and differs from it in the manner in which BAR and BA frames are acknowledged. Support for HT-delayed block ack is optional and a station advertises support by setting the HT-delayed Block Ack capability bit in the HT Capabilities element (see Section 11.3.2.2). A peer station may establish a HT-delayed block ack session with a station that advertises itself as HT-delayed block ack capable.

Under HT-delayed block ack the BAR and BA frames carry a BAR Ack Policy and BA Ack Policy field, respectively. If set to 1 this indicates that the receiver of the frame should not return an ACK response.

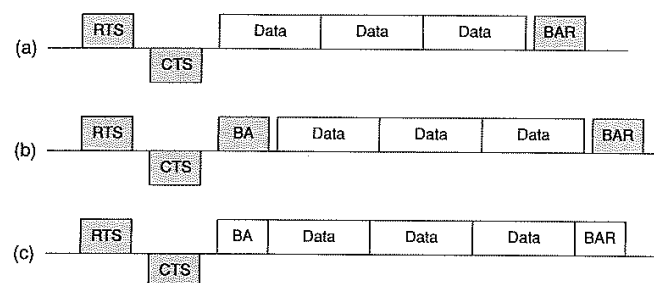


Figure 8.18 Typical HT-delayed TXOP sequences.

8.5.1 HT-delayed block ack TXOP sequences

Typical TXOP sequences under HT-delayed block ack are shown in Figure 8.18. The TXOP begins with a short frame exchange which could be RTS/CTS (as shown) or Data/ACK. HT-delayed block allows the TXOP to be fully utilized by the originator since no immediate response is required from the responder. Sequence (a) shows the originator sending an aggregate with ack policy set to Block Ack followed by a BAR. When data flows both ways, the BA may be combined with data in the reverse direction. This is shown in sequence (b).

It is possible, with reduced robustness, to aggregate the BA and BAR frames together with the data. This is shown in sequence (c).

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9 Advanced channel access techniques

This chapter introduces some of the advanced channel access techniques in the 802.11 standard. In addition to, and built upon, the distributed channel access techniques described in Chapter 7, the 802.11 standard includes two centrally coordinated channel access techniques. The PCF was introduced in the original 802.11 specification and HCCA was introduced in the 802.11e amendment to support parameterized QoS and to fix some of the deficiencies in the PCF. The chapter then introduces new channel access techniques in the 802.11n amendment.

A very simple technique called the reverse direction protocol was introduced with one bit of additional signaling and some simple changes to the rules for operating a TXOP. This technique is particularly effective under EDCA for improving throughput for certain traffic patterns.

During the development of the 802.11n amendment, a strong interest emerged among many participants for improving the power efficiency of the MAC protocol. While outside the scope of the 802.11n PAR this resulted in the power-save multi-poll (PSMP).

9.1 PCF

Infrastructure network configurations may optionally include the point coordination function (PCF). With PCF, the point coordinator (PC), which resides in the AP, establishes a periodic contention free period (CFP) during which contention free access to the wireless medium is coordinated by the PC. During the CFP the NAV of all nearby stations is set to the maximum expected duration of the CFP. In addition, all frame transfers during the CFP use an inter-frame spacing that is less than that used to access the medium under DCF, preventing stations from gaining access to the medium using contention-based mechanisms. At the end of the CFP, the PC resets the NAV of all stations and regular contention-based access proceeds.

9.1.1 Establishing the CFP

The PC periodically establishes a CFP using a Beacon frame as illustrated in Figure 9.1. The CFP repetition interval is a fixed multiple of the DTIM (delivery traffic indication message) period where the DTIM or delivery traffic indication message is an element that

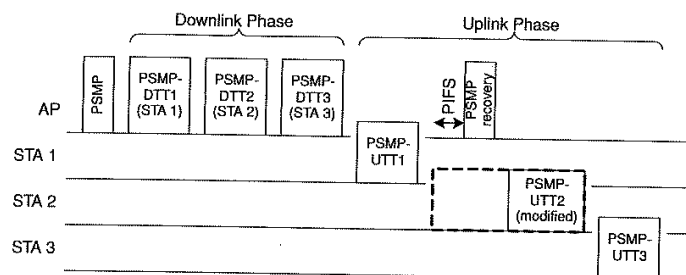


Figure 9.7 PSMP recovery.

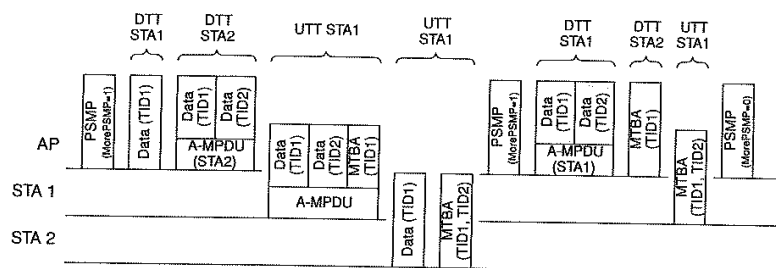


Figure 9.8 PSMP burst. Reproduced with permission from IEEE (2007) © IEEE.

This is illustrated in Figure 9.7 where STA 2 has not transmitted a frame at the start of PSMP-UTT2. The AP detects this condition and transmits a PSMP recovery frame PIFS after the start of the UTT. The PSMP recovery frame contains an uplink schedule for STA 2 using the remaining time of the original PSMP-UTT2 and referenced from the end of the PSMP recovery frame.

9.4.2 PSMP burst

The AP may perform a burst or series of PSMP exchanges as illustrated in Figure 9.8. After an initial data exchange in the first PSMP sequence, additional resource allocation can be made and retransmissions scheduled in subsequent PSMP exchanges. The PSMP frame in the first exchange has its MorePSMP bit set to 1 to indicate that another PSMP exchange will follow. The MorePSMP bit is set in the PSMP frame of each subsequent exchange as long as the burst continues. The burst is terminated with a PSMP frame that has the MorePSMP bit set to 0.

9.4.3 Resource allocation

A station uses the TSPEC mechanism defined for HCCA to reserve resources. The AP establishes a PSMP schedule based on the TSPECs received from one or more stations such that the QoS requirements of all stations are met.

To accommodate variability in traffic loads and retransmission, a station may adjust the resource allocation long term by updating its TSPEC or, more dynamically, using the Queue Size or TXOP Duration Request fields in the QoS Data frame (see Section 11.1.5). If a station cannot transmit all its queued data within the allocated PSMP-UTT, the station will transmit as much data as possible and piggyback a resource request for an additional PSMP-UTT. If the AP receives a resource request it will attempt to satisfy that request in the next PSMP sequence.

9.4.4 Block ack usage under PSMP

PSMP does not permit the use of ACK for data exchanges, thus data transfer that requires acknowledgment must occur under an established block ack session. Data may be transferred without a block ack session, however, it can only do so using No Ack ack policy.

PSMP requires that the block ack session be HT-immediate or HT-delayed. Under HT-immediate, use of the Multi-TID BAR and Multi-TID BA is required, while under HT-delayed the BAR and BA frames should have their BAR/BA ack policy set to No Ack.

References

- IEEE (2007). *IEEE P802.11n™/D3.00, Draft Amendment to STANDARD for Information Technology – Telecommunications and Information Exchange Between Systems – Local and Metropolitan Networks – Specific Requirements. Part 11: Wireless LAN Medium Access Control (MAC) and Physical Layer (PHY) Specifications: Amendment 4: Enhancements for Higher Throughput.*
- Mujtaba, S. A., Stephens, A. P., Myles, A., et al. (2005). *TGnSync Proposal MAC Results*, IEEE 802.11-04/0892r7, July.

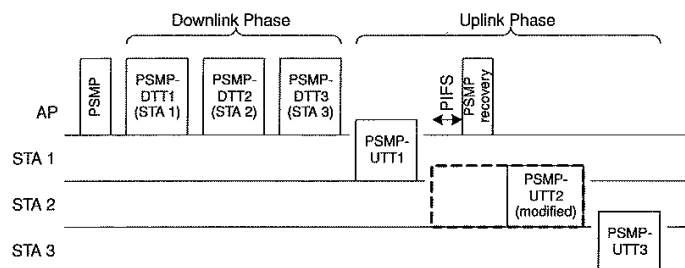


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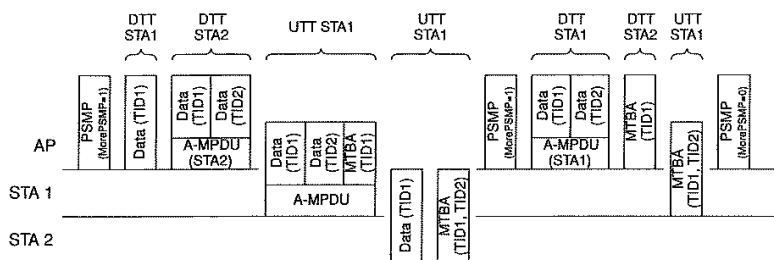


Figure 9.8 PSMP burst. Reproduced with permission from IEEE (2007) © IEEE.

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10 Interoperability and coexistence

The 802.11n amendment, more so than any previous 802.11 amendment, introduces many optional features geared toward specific market segments that will likely only be deployed in specific classes of devices. Many of the optional features are complex and, given time to market concerns and cost constraints, many implementations will only adopt a subset of the available features, perhaps phasing in features with time or in higher end products. Some features appear in multiple flavors. This is due to the many unknowns present at the time the features were being discussed during the standard development process. Very often there was no single clear direction to take that would clearly suit all situations and so variations were introduced.

The popularity of 802.11 also means that there are a large number of legacy 802.11 devices deployed making interoperability and coexistence with those devices essential.

This chapter discusses various features that help ensure interoperability and coexistence between 802.11n compliant devices as well as legacy 802.11 devices.

0.1 Station and BSS capabilities

With the large number of optional features in the 802.11n amendment a fair bit of signaling is required to establish device capabilities and ensure interoperability. Also, care must be taken to ensure that a feature used by one station does not adversely affect the operation of a neighboring station that is not directly involved in the data frame exchange.

HT station capabilities are signaled using the HT Capabilities element. This element is present in Beacon, Association Request, Association Response, Reassociation Request, Reassociation Response, Probe Request, and Probe Response management frames sent by an HT AP or HT station. The optional capabilities signaled by the HT Capabilities element are summarized in the following sections, with the details on the HT Capabilities element itself provided in Section 11.3.2.2.

0.1.1 HT station PHY capabilities

The HT station PHY capabilities include the following:

- **Supported channel width set.** Indicates whether the station is capable of 20 MHz operation only or both 20 MHz and 40 MHz operation.

- **Greenfield format PPDU.** Indicates whether or not the station is capable of receiving HT Greenfield format PPDU.
- **LDPC coding.** Indicates whether or not the station is capable of receiving LDPC coded PPDU.
- **Short GI in 20 MHz PPDU formats.** Indicates whether or not the station is capable of receiving 20 MHz frames using the short guard interval.
- **Short GI in 40 MHz PPDU formats.** Indicates whether or not the station is capable of receiving 40 MHz frames using the short guard interval.
- **Transmitting STBC formats.** Indicates whether or not the station is capable of transmitting space-time block coded frames.
- **Receiving STBC formats.** Indicates whether or not the station is capable of receiving space-time block coded frames and if so, whether that capability is limited to one spatial stream, one or two spatial streams or one, two, and three spatial streams.
- **DSSS/CCK in 40 MHz.** Indicates whether or not the station is capable of transmitting and receiving DSSS/CCK format frames (802.11b) while operating with 40 MHz channel width.
- **40 MHz intolerant.** Indicates whether or not the station requires the AP to prevent 40 MHz transmissions.

10.1.2 HT station MAC capabilities

The HT station MAC capabilities include the following:

- **Spatial multiplexing power save mode.** Indicates the level of support for Spatial Multiplexing (SM) Power Save: static, dynamic or disabled.
- **HT delayed block ack.** Indicates whether or not the station supports HT-delayed block ack.
- **PSMP support.** Indicates whether or not the station supports PSMP operation.
- **L-SIG TXOP protection.** Indicates whether or not the station supports the L-SIG TXOP protection mechanism.
- **+HTC field support.** Indicates whether or not the station can parse received frames containing the HT Control (HTC) field.
- **Reverse direction protocol.** Indicates whether or not the station can act as a reverse direction responder.
- **Maximum A-MSDU length.** Indicates the maximum A-MSDU length that the station can receive: 3839 or 7935 octets.
- **A-MPDU parameters.** Indicates the processing limitations with respect to A-MPDU delineation. The station may be limited in the length of A-MPDU it can receive as well as the rate at which MPDUs within the A-MPDU can be de-aggregated.

10.1.3 BSS capabilities

Some of the HT Capability fields have additional meaning that applies to the BSS when transmitted by the AP in Beacon, Probe Response, Association Response, and Reassociation Response frames. These fields are:

- **PCO Support.** This indicates the BSS supports phased coexistence operation (PCO).
- **PSMP Support.** This indicates that the BSS supports PSMP operation.
- **DSSS/CCK Mode in 40 MHz.** This indicates whether or not member stations of the BSS are permitted to transmit DSSS/CCK format PPDUs during 20/40 MHz BSS operation.

10.1.4 Advanced capabilities

Some advanced capabilities for which support is indicated in the HT Capabilities element are:

- MCS feedback
- transmit beamforming support
- antenna selection.

These capabilities are discussed in more detail in Chapter 12.

10.2 Controlling station behavior

To control the operation of stations on a BSS, the AP uses the HT Information element in Beacon, Probe Response, Association Response, and Reassociation Response frames. Details on the HT Information element can be found in Section 11.3.2.3.

The AP controls 20 MHz and 20/40 MHz operation of the BSS through the Supported Channel Width Set field in the HT Capabilities element and the following fields in the Information element:

- **Primary Channel.** This field gives the channel number of the primary channel for the BSS.
- **Secondary Channel Offset.** This field gives the offset of the secondary channel relative to the primary channel.
- **STA Channel Width.** Defines the channel widths with which stations may transmit to the AP.

Twenty MHz and 20/40 MHz BSS operation is described in more detail in later sections.

The AP also provides information on the capabilities of stations that are members of the BSS or, optionally, detected operating in the same channel(s). This enables protection for certain HT sequences a station may use but for which nearby station may not correctly defer. The fields in the HT Information element that provide information on nearby stations for the purpose of enabling protection are the following:

- **HT Protection.** This indicates the types of stations that are present as members or non-members of the BSS and thus the nature of the protection required. Four categories are distinguished as listed in Table 10.1.

Table 10.1 HT Protection field encoding

Encoding	Condition
0 (HT members with same channel width, HT non-members)	One of the following is true: • This is a 20/40 MHz BSS with only 20/40 MHz HT station members. Any non-member stations detected in the primary or secondary channel are also HT, i.e. no non-HT stations are detected • This is a 20 MHz BSS with 20 MHz HT station members. Any non-member stations detected in the channel are also HT, i.e. no non-HT stations are detected
1 (HT members with different channel widths, non-HT non-members)	One of the following is true: • This is a 20/40 MHz BSS with either 20/40 MHz HT station or 20 MHz HT station members and at least one non-HT station non-member has been detected in the primary or secondary channel • This is a 20 MHz BSS with 20 MHz HT station members and at least one non-HT station non-member has been detected in the primary channel
2 (HT members with different channel widths, HT non-members)	This is a 20/40 MHz BSS possibly with 20/40 MHz HT station members, but with at least one 20 MHz HT station member and only HT station non-members, if any, have been detected in the primary and secondary channels
3 (HT and non-HT members)	Otherwise, i.e. 20/40 MHz or 20 MHz BSS with both HT and non-HT members

- **Non-Greenfield HT STAs Present.** This indicates that there are HT stations associated with the BSS or in neighboring BSSs that are unable to receive HT Greenfield format PPDUs. When set, stations must protect HT Greenfield format transmissions.
- **L-SIG TXOP Protection Full Support.** This indicates whether or not all stations on the BSS support L-SIG TXOP Protection. A station may use this to determine whether or not to use L-SIG TXOP Protection.
- **OBSS Non-HT STAs Present.** This indicates that there may be non-HT stations present that are not members of the BSS. A station may optionally use this to determine if protection is necessary for HT sequences.
- **RIFS Mode.** The AP can directly control whether or not RIFS bursting is permitted through the RIFS Mode bit. Stations may only use RIFS bursting when this bit is set to 1.

The HT Information element also provides information for controlling PSMP operation (see Section 9.4), PCO operation (see Section 10.4) and the use of Dual Beacons to extend range.

Finally, the Basic MCS Set field of the HT Information element provides the set of MCS values that are supported by all HT stations in the BSS.

10.3 20 MHz and 20/40 MHz operation

802.11n introduces optional 40 MHz channel width operation but must maintain interoperability with legacy and HT 20 MHz stations. There are thus three basic classes of

device: 20 MHz legacy or non-HT stations, 20 MHz HT stations, and 20/40 MHz HT stations. Twenty MHz HT stations are capable of sending and receiving 20 MHz transmissions, while 20/40 MHz stations are capable of sending and receiving both 20 MHz and 40 MHz transmissions. A 20/40 MHz HT station is also capable of operating as a 20 MHz HT station, i.e. a 20/40 MHz station may choose to advertise itself as a 20 MHz HT station and send and receive only 20 MHz transmissions. A legacy or non-HT station will not include the HT Capabilities element in any management frames.

A 20 MHz HT station sets the Supported Channel Width Set subfield in the HT Capabilities Info field of the HT Capabilities element to 0, indicating that it only supports 20 MHz operation. A 20/40 MHz HT station sets the Supported Channel Width Set subfield in the HT Capabilities Info field of the HT Capabilities element to 1, indicating that it supports both 20 MHz and 40 MHz operation.

In addition to, and because of, the various device classes there are two basic modes of BSS operation:

- **20 MHz BSS operation.** If 40 MHz operation is not feasible, either because the AP is not capable of supporting a 20/40 MHz BSS or because the environment does not permit 40 MHz operation then the AP operates a 20 MHz BSS.
- **20/40 MHz BSS operation.** If 40 MHz operation is feasible then the AP may operate a 20/40 MHz BSS. For 20/40 MHz BSS operation two adjacent 20 MHz channels are used, one of which is designated the primary channel and the other the secondary channel. A 20/40 MHz capable HT station may associate with the AP and operate with 40 MHz channel width. 20 MHz only HT stations and legacy stations may also associate with the AP and operate only on the primary 20 MHz channel.

13.1 Beacon transmission

The Beacon is generally transmitted using a format that is most broadly supported for a particular frequency band. The Beacon is always transmitted in a 20 MHz non-HT format on the primary channel (which is the only channel in the case of 20 MHz operation). In the case of the 5 GHz bands, a non-HT OFDM format is used since all 802.11 devices using these bands are expected to be backward compatible with 802.11a. In the case of the 2.4 GHz band, a DSSS/CCK format is typically used since most devices are backward compatible with 802.11b. A non-HT OFDM format may be used in the 2.4 GHz band if the AP is only capable of, or only wishes to support, OFDM transmissions.

An AP may also transmit a secondary Beacon using an STBC PPDU to extend the range over which the Beacon may be detected. This Beacon would be transmitted at half beacon period shift relative to the primary Beacon and has the STBC Beacon bit set in the HT Information element.

13.2 20 MHz BSS operation

Any HT AP (20 MHz or 20/40 MHz capable) may operate a 20 MHz BSS. A 20/40 MHz capable AP may choose to operate a 20 MHz BSS as a configuration option or if it is

unable to find a suitable pair of 20 MHz channels on which to operate a 20/40 MHz BSS.

The HT AP indicates 20 MHz BSS operation by setting the Secondary Channel Offset subfield in the HT Information element to 0 to indicate that there is no secondary channel. The Primary Channel subfield in the HT Information element is set to indicate the 20 MHz channel number of the BSS (the channel on which the Beacon and all other frames are transmitted).

An HT station, whether it is 20 MHz only or 20/40 MHz capable, that associates with a 20 MHz BSS will send and receive 20 MHz transmissions.

In other respects, the 20 MHz BSS operation is relatively straightforward and similar to legacy 20 MHz BSS operation.

10.3.3 20/40 MHz BSS operation

Any HT AP (20 MHz or 20/40 MHz capable) may operate a 20/40 MHz BSS. The HT AP indicates 20/40 MHz operation by setting the Secondary Channel Offset subfield in the HT Information element to a non-zero value indicating the offset of the secondary channel relative to the primary channel, i.e. +1 or -1. The Primary Channel subfield indicates the channel number of the primary channel.

An HT station may associate with the AP as either a 20 MHz capable HT station or as a 20/40 MHz capable HT station. A 20 MHz HT station would send and receive only 20 MHz PPDU's and would only use the primary 20 MHz channel. A 20/40 MHz station would be able to send and receive either 20 MHz or 40 MHz PPDU's. All 20 MHz transmissions would only use the primary channel. The secondary channel should remain clear except for 40 MHz use and possible neighboring BSS use, although this should be avoided through careful channel selection.

If the HT AP is only 20 MHz capable the AP will not be able to send or receive 40 MHz transmissions. In this case, 40 MHz transmissions will only be exchanged between 20/40 MHz capable member stations of the BSS using direct links.

Coexistence with neighboring BSSs and among the different device classes associated with the BSS during 20/40 MHz operation is managed through a number of mechanisms, including the following:

- Overlapping BSS scanning and careful channel selection on initial BSS setup to avoid channels already in use by other BSSs.
- Changing channels or operating width after BSS setup if a new BSS is detected operating on the secondary channel.
- Independent CCA on both the primary and secondary channels.
- Protection mechanisms such as CTS-to-self sent in legacy compatible format on both primary and secondary channels.
- BSS Operating modes such as phased coexistence operation.

Also, because of the very different nature of the 2.4 GHz and 5 GHz bands, different rules apply in each case.

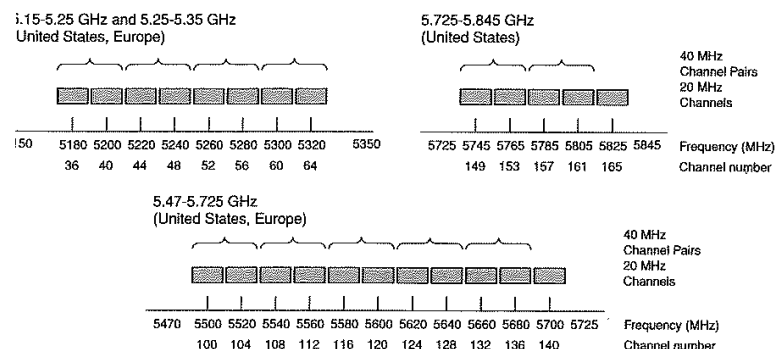


Figure 10.1 20 MHz channel pairings for 40 MHz operation in the 5 GHz bands.

0.3.3.1 20/40 MHz operation in the 5 GHz bands

Forty MHz operation in the 5 GHz bands is accomplished by utilizing two adjacent 20 MHz channels. One of the 20 MHz channels is designated the primary channel and the other the secondary channel. The permissible channel options for various regulatory domains are listed in Appendix 5.1 and summarized in Figure 10.1.

Compared to the 2.4 GHz band, the 5 GHz bands have far more bandwidth and easily accommodate multiple 40 MHz channels. Also, unlike the 2.4 GHz band, there are no partial channel overlaps. The 5 GHz band is thus preferred for 40 MHz operation, particularly in environments where multiple APs are deployed, such as a large enterprise.

0.3.3.2 20/40 MHz operation in the 2.4 GHz band

Forty MHz operation in the 2.4 GHz band is complicated by the limited spectrum available, overlapping channels and the ubiquitous deployment of legacy 802.11b and 802.11g devices operating in this band. During the development of the standard, arguments were made to disallow 40 MHz operation in the 2.4 GHz band because of the difficult coexistence issues. However, enough members in the task group believed that a solution was possible, however imperfect, and the argument was made that since 40 MHz operation provided a significant performance gain over 20 MHz operation at little additional product cost, vendors would develop 40 MHz devices whether or not the standard permitted 40 MHz operation. It was felt, although perhaps by only a slim majority of the members, that standardized operation would provide better coexistence in the end than attempting to disallow it, which would effectively lead to non-standardized 40 MHz operation.

The 2.4 GHz band is composed of channels spaced 5 MHz apart. The vast majority of legacy deployments, however, make use of channels 1, 6, and 11, which is clearly shown in an informal survey of access points taken in the San Francisco Bay area, USA (1088 APs) and in the Netherlands, Belgium, and Italy (1722 APs) shown in

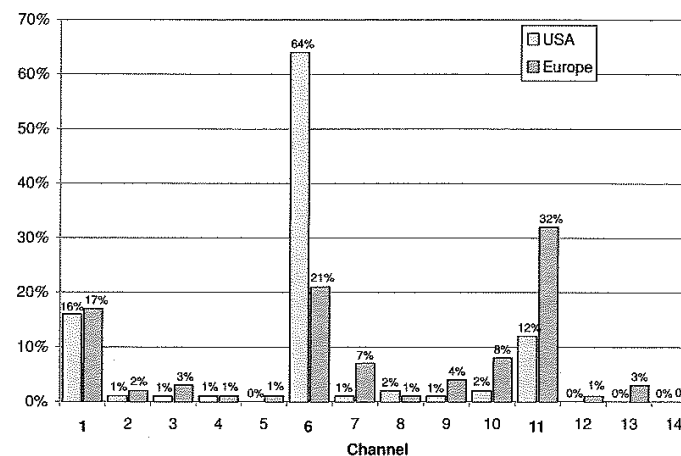


Figure 10.2 Percentage of legacy APs using a given channel from an informal survey of 1088 APs in San Francisco Bay area (representing the USA) and 1722 APs in the Netherlands, Belgium, and Italy (representing Europe). Reproduced from van Nee *et al.* (2006).

Figure 10.2. These are the so called non-overlapping channels and are spaced 25 MHz apart and referred to here as the widely used channels.

A major issue with supporting 40 MHz operation in the 2.4 GHz band is coexistence with neighboring BSSs on one or more of these widely used channels. To highlight the issue consider Figure 10.3. This shows two 40 MHz channel options coexisting with possible neighboring 20 MHz BSSs on channels 1, 6, and 11. If neighboring BSSs exist on only two of the widely used channels then it is possible to select adjacent primary and secondary channels such that the primary channel coincides with one of the used channels and the secondary channel occupies spectrum overlapping only unutilized channels.

If all three of the widely used channels are occupied then channel selection is more problematic. If one of the widely used channels is selected for the primary channel then the secondary channel will partially overlap another of the widely used channels.

Another issue with 40 MHz operation in the 2.4 GHz band is the limited bandwidth available. Two 40 MHz BSSs can be established in regulatory domains, such as Europe, where 13 (or more) channels are available. However, in the USA where only 11 are available, only one non-overlapping 40 MHz BSS can be established.

10.3.3.3 A brief history of 40 MHz in the 2.4 GHz band

The first draft of the 802.11n amendment (IEEE, 2007b) paid scant attention to the coexistence issues in the 2.4 GHz band. CCA was not required on the secondary channel and this meant that a 20/40 MHz station could disregard any traffic on channels overlapping

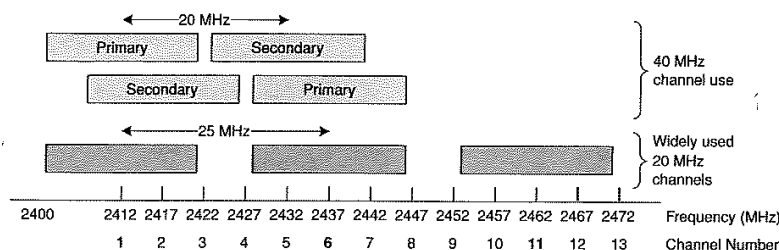


Figure 10.3 40 MHz and 20 MHz channel use in the 2.4 GHz ISM band. Reproduced with permission from van Nee *et al.* (2006) © IEEE.

the secondary channel when transmitting. When selecting a channel for a 20/40 MHz BSS, it was recommended that the AP scan for neighboring BSSs prior to the establishment of the 40 MHz BSS, but no recommendations were made on how to behave if a legacy BSS started operating nearby after a channel had been selected.

Leading up to the second draft, simulation results were presented (Kandala *et al.*, 2006) that showed that a heavily loaded 20/40 MHz BSS would significantly impair the operation of a neighboring 20 MHz BSS on or overlapping with the secondary channel. The results also showed that simply performing an energy detect CCA on the secondary channel and deferring for any detected transmissions significantly improved the situation. Combining CCA on the secondary channel with CCA on the primary channel and performing the full random backoff procedure provided a moderate further improvement for traffic on the neighboring BSS. Based on these results the group adopted the requirement that CCA be performed on the secondary channel and that the 20/40 MHz station defer for traffic on the secondary channel. It was decided, however, that the full random backoff based on CCA on the secondary channel was not required.

Scanning for neighboring BSSs prior to establishing a 20/40 MHz BSS was made mandatory. Rules were adopted that would prevent the operation of a 20/40 MHz BSS when existing BSSs are detected on or overlapping the secondary channel.

Concerns were also expressed that non-802.11 use of the 2.4 GHz band might be interfered with by 40 MHz operation. The example given was an IEEE 802.15.1 (Bluetooth) wireless personal area network (WPAN) that can operate on a reduced channel set in the presence of 20 MHz 802.11 transmissions but would fail if the number of available channels fell below a certain threshold, as may happen in the presence of 40 MHz 802.11 transmissions. As a result, the 40 MHz Intolerant information bit was adopted which may be set by any station to prevent 40 MHz operation. A station that has knowledge of incompatibilities with 40 MHz operation (perhaps a system with both 802.11 and 802.15.1 radios) could set this bit and thus prevent an AP from operating a 40 MHz BSS. Somewhat controversially, there are no restrictions on the use of this bit and any station can set this bit for any reason.

10.3.4 Clear channel assessment in 20 MHz

An HT station with an operating channel width of 20 MHz must provide clear channel assessment (CCA) as follows:

- CCA must indicate busy with a sensitivity of at least -82 dBm to a valid 20 MHz PPDU transmission. This is referred to as signal detect CCA.
- CCA must indicate busy with a sensitivity of at least -62 dBm to any signal in the 20 MHz channel. This is referred to as energy detect CCA.
- A station that does not support the reception of Greenfield format PPDU must still indicate CCA busy with a sensitivity of at least -72 dBm for any valid Greenfield format PPDU.

The first two requirements are identical to the 802.11a and 802.11g requirements. The last requirement ensures that even a station that is unable to receive the optional Greenfield format PPDU will defer for Greenfield format PPDU at sensitivity levels better than the normal energy detect levels.

10.3.5 Clear channel assessment in 40 MHz

An HT station with an operating channel width of 40 MHz must provide CCA on both the primary and secondary channels. CCA goes busy under the following conditions:

- When the secondary channel is idle, CCA on the primary channel must go busy with a sensitivity of at least -82 dBm for a valid 20 MHz PPDU transmission.
- When the secondary channel is idle, CCA for the primary channel must go busy with a sensitivity of at least -62 dBm to any signal on the primary channel.
- A station that does not support the reception of Greenfield format PPDU must still indicate CCA busy with a sensitivity of at least -72 dBm for any valid 20 MHz Greenfield format PPDU on the primary channel when the secondary channel is idle.
- CCA for both the primary and secondary channel must go busy with a sensitivity of at least -79 dBm to a valid 40 MHz PPDU transmission.
- CCA for both the primary and secondary channel must go busy with a sensitivity of at least -59 dBm to any signal present on both channels.
- A station that does not support reception of Greenfield format PPDU must still indicate CCA busy for both the primary and secondary channel with a sensitivity of at least -69 dBm for any valid 40 MHz Greenfield format PPDU.
- When the primary channel is idle, CCA for the secondary channel must go busy with a sensitivity of at least -62 dBm to any signal on the secondary channel.

The 40 MHz CCA requirements are thus a simple extension of the 20 MHz requirements and essentially allow for expected 20 MHz traffic on the primary channel and 40 MHz traffic on the two bonded 20 MHz channels. Only energy detect is performed on the secondary 20 MHz channel.

10.3.6 Channel access for a 40 MHz transmission

A 20/40 MHz station gains an EDCA TXOP for transmission on slot boundaries based on the 20 MHz primary channel CCA. The station shall, however, only transmit a 40 MHz PPDU if the secondary channel has been idle for at least a PIFS interval prior to the expiry of the backoff counter. If the secondary channel has not been idle for a PIFS interval then the station must restart its backoff count by selecting a random number in the same contention window as the previous backoff count.

A TXOP obtained for the transmission of 40 MHz PPDU's may also be used for 20 MHz PPDU's. Note that this means that during the TXOP the secondary channel would be idle during the 20 MHz transmission and a 20 MHz station on the secondary channel may begin transmitting, possibly interfering with the 40 MHz transmissions.

10.3.7 NAV assertion in a 20/40 MHz BSS

An HT station updates its NAV using the Duration/ID field value in any frame received in a 20 MHz PPDU in the primary channel or received in a 40 MHz PPDU. NAV is set if the RA of the received frame does not match the MAC address of the receiving station. A station does not need to set its NAV in response to 20 MHz frames received on the secondary channel even if it is capable of receiving those frames.

10.3.8 OBSS scanning requirements

Before an AP establishes a 20/40 MHz BSS it must scan for existing BSSs in the channels in the frequency range affected by the new BSS. This type of scanning is referred to as overlapping BSS (OBSS) scanning. It is expected that the scanning will allow the AP, either autonomously or through a network operator, to select the most appropriate channels for 20/40 MHz BSS operation from the channels available in the regulatory domain. The 802.11n amendment does not dictate rules for selecting the most appropriate pair of channels, but does provide rules that prevent 20/40 MHz operation in a manner that would interfere with existing 20 MHz or 20/40 MHz BSSs.

As with legacy 20 MHz BSSs, scanning to improve coexistence is not required for 20 MHz HT BSSs. However, it is recommended that a 20 MHz BSS not be established in the secondary channel of an existing 20/40 MHz BSS.

10.3.8.1 Establishing a 20/40 MHz BSS in the 5 GHz bands

If an AP chooses to start a 20/40 MHz BSS in the 5 GHz bands that occupies the same two channels as an existing 20/40 MHz BSS then it must ensure that the primary and secondary channels of the new BSS match the primary and secondary channels of the existing BSS. If two or more 20/40 MHz BSSs are present with non-matching primary and secondary channels then the AP is free to match primary and secondary channels with any of the existing BSSs. This situation may arise if existing BSSs cannot detect each other as shown in Figure 10.4.

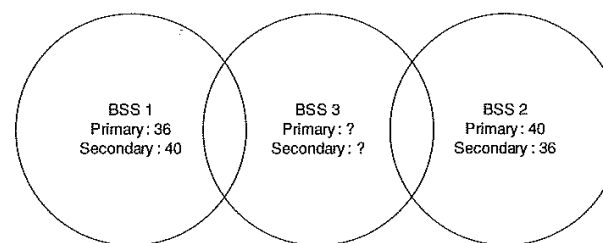


Figure 10.4 Establishing a 20/40 MHz BSS between two existing 20/40 MHz BSSs with non-matching primary and secondary channels. Reproduced with permission from Chan *et al.* (2007) © IEEE.

It is recommended that the AP not establish a 20/40 MHz BSS where the secondary channel is occupied by a 20 MHz BSS. Conversely, it is recommended that an AP not establish a 20 MHz BSS in the secondary channel of an existing 20/40 MHz BSS.

Unlike the 2.4 GHz band, there are no overlapping 20 MHz channels in the 5 GHz band. Once the 20/40 MHz BSS has been established the AP may periodically scan the secondary channel for 20 MHz OBSSs.

10.3.8.2 Establishing a 20/40 MHz BSS in the 2.4 GHz band

Because of the limited bandwidth and overlapping 20 MHz channels, the rules for establishing a 20/40 MHz BSS in the 2.4 GHz band are more restrictive than for the 5 GHz bands. Before an AP can establish a 20/40 MHz BSS in the 2.4 GHz band, it must determine that the following conditions are true over the frequency range affected by the new BSS:

1. The primary and secondary channels of the new BSS are the same as the primary and secondary channels of all existing 20/40 MHz BSSs.
2. The primary channel is the same as the operating channel of all existing 20 MHz BSSs.

The frequency range affected includes all channels with center frequencies that are within 25 MHz of the center frequency of the 40 MHz channel, which could be up to 11 channels. Figure 10.5 shows an example of a permitted 20/40 MHz BSS under these rules.

In the USA where there are 11 channels in the 2.4 GHz band, only one non-overlapping 20/40 MHz BSS can be established. Thus the first rule implies that if there is an existing 20/40 MHz BSS in the 2.4 GHz band the AP must ensure that the primary and secondary channels of the new BSS be the same as the primary and secondary channels of the existing BSS.

In Europe where there are 13 channels available in the 2.4 GHz band, it is possible to establish two non-overlapping 20/40 MHz BSSs, one using the channel pair (1, 5) and another using the channel pair (9, 13). However, if an existing 20/40 MHz BSS is not

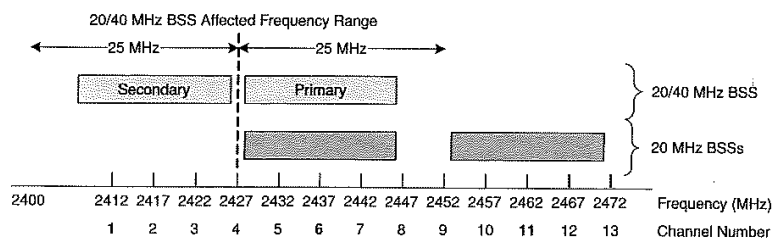


Figure 10.5 An example of permitted 20/40 MHz BSS operation.

using one of these two channel pairs, then the new BSS must use the same primary and secondary channel as the existing BSS.

There are some instances where establishing a 20/40 MHz BSS may not be possible at all. For example, if there are existing 20 MHz BSSs on each of the widely used channels 1, 6, and 11 or even when there are just two 20 MHz BSSs, for example using channels 4 and 11.

In a situation similar to that in Figure 10.4, where two existing 20/40 MHz are hidden from each other and using non-matching primary and secondary channels, it is not permitted to establish a 20/40 MHz BSS.

0.3.8.3 OBSS scanning during 20/40 MHz BSS operation

During the operation of a 20/40 MHz BSS in the 2.4 GHz band, active 20/40 MHz HT stations associated with the BSS are required to monitor the channels affected by the frequency range of the BSS to ensure that the conditions that allowed the establishment of the BSS do not change. The AP itself is not required to perform any monitoring, although it may still do so.

Monitoring stations keep a local record of the channels that are in use by other BSSs. Monitoring stations that receive Beacons determine the primary channel on which the BSS is operating by examining the DS Parameter Set element in the Beacon frame rather than from the channel number of the channel on which the Beacon was received. This is because it is possible for a station to receive a DSSS format Beacon with a 5 MHz center frequency offset, i.e. from an adjacent channel in the 2.4 GHz band, thus the center frequency of the receiver is not a reliable indicator of the channel. The existence of the secondary channel is determined by examining the Secondary Channel Offset field in the HT Information element, if present.

If a change occurs to the local record that would prohibit the operation of the 20/40 MHz BSS, or if the channels in use by neighboring BSSs change, then the station sends a 20/40 Coexistence Management frame to the AP. If the 20/40 Coexistence Management frame is sent because conditions prohibit the operation of the 20/40 MHz BSS, then the 20 MHz BSS Width Request field is set to 1. On receiving a 20/40 MHz Coexistence Management frame with the 20 MHz BSS Width Request field set to 1, the AP must immediately switch the BSS to 20 MHz operation.

With the 20/40 Coexistence Management frame, the monitoring station also reports the list of occupied channels. The channel list would include channel numbers for the channels, other than the primary channel of the BSS to which the station is associated, on which at least one 20 MHz BSS has been detected. The list would also include the channel number of the primary channel of any 20/40 MHz BSS with a primary channel and secondary channel that does not match the primary and secondary channel of the BSS to which the station is associated.

If the channels in use by neighboring BSSs change, the monitoring station reports the change in the channel list in the 20/40 Coexistence Management frame. The AP uses the in-use channel information reported by all monitoring stations as well as its own information to determine if conditions permit the operation of a 20/40 MHz BSS. If such conditions persist for $\text{dot11BSSWidthChannelTransitionDelayFactor} \times \text{dot11BSSWidthTriggerScanInterval}$ seconds (see MIB variables for 20/40 MHz scanning in Table 10.1), the AP may resume 20/40 MHz BSS operation.

An AP switches from 20/40 MHz BSS operation to 20 MHz BSS operation by setting the Secondary Channel Offset field to 0 in the next Beacon transmission. The AP will also set the STA Channel Width to 0 indicating that it can only receive 20 MHz transmissions. Similarly, the AP may switch from 20 MHz BSS operation to 20/40 MHz BSS operation by setting the Secondary Channel Offset field to +1 or -1 and updating its STA Channel Width field. An AP may also use the Channel Switch Announcement mechanism to change the BSS operating width or switch channels.

10.3.8.4 Scanning requirements for 20/40 MHz stations

An OBSS scan is a passive or active scan of a set of channels that may be affected by 20/40 MHz BSS operation. In the 5 GHz bands the affected channels are simply the two channels the 20/40 MHz BSS occupies or will occupy. In the 2.4 GHz band the affected channels would be any channels in the regulatory domain that are within 25 MHz of the center frequency of the paired 20 MHz channels the 20/40 MHz BSS will occupy or occupies.

OBSS scans are performed by an AP before it establishes a 20/40 MHz BSS and by active 20/40 MHz HT stations associated with a 20/40 MHz BSS. The parameters for OBSS scanning are listed in Table 10.2 together with their default values. The AP may override the default values in HT stations associated with the BSS by transmitting an Overlapping BSS Scan Parameters element (see Section 11.3.2.4).

An active 20/40 MHz HT station must perform at least 2 scans of each channel every $\text{dot11BSSWidthTriggerScanInterval}$. The scan of each channel may be an active scan or a passive scan. Passive scanning involves simply listening for Beacons while active scanning involves sending Probe Request frames and processing any received Probe Response frames. Active scanning is prohibited in some regulatory domains, for example in some radar bands, necessitating passive scanning.

Each time the station scans a particular channel it must examine that channel for a minimum duration of $\text{dot11OBSSScanPassiveDwell}$ for a passive scan and $\text{dot11OBSSScanActiveDwell}$ for an active scan. Typically the station will not stay in the scanned channel for much longer since by doing so it may miss traffic on the

Table 10.2 Parameters for 20/40 MHz scanning

Parameter	Default Value
dot11BSSWidthTriggerScanInterval	300 seconds
dot11OBSSScanPassiveDwell	20 TU (where TU = 1024 μ s)
dot11OBSSScanActiveDwell	10 TU
dot11OBSSScanPassiveTotalPerChannel	200 TU
dot11OBSSScanActiveTotalPerChannel	20 TU
dot11BSSWidthChannelTransitionDelayFactor	5
dot11OBSSScanActivityThreshold	25 $\times$ 1/100 percent (= 0.25%)

BSS channels. Thus the station will revisit the channel periodically over the scan interval. The station must scan each channel for a minimum total time of dot11OBSSScanPassiveTotalPerChannel or dot11OBSSScanActiveTotalPerChannel respectively for a passive or active scan respectively.

Not all stations are required to scan, only the AP prior to establishing a 20/40 MHz BSS and active stations. An active station is a station for which the fraction of time spent transmitting MSDUs and receiving unicast MSDUs directed to that station over the prior dot11BSSWidthChannelTransitionDelayFactor scan intervals exceeds the threshold dot11OBSSScanActivityThreshold/100. In other words, if

$$\text{STA activity} = \frac{\text{time spent transmitting MSDUs} + \text{time spent receiving unicast MSDUs}}{\text{activity period}} \quad (10.1)$$

where

$$\text{activity period} = \text{dot11BSSWidthChannelTransitionDelayFactor} \times \text{dot11BSSWidthTriggerScanInterval} \quad (10.2)$$

and

$$\text{STA activity} > \frac{\text{dot11OBSSActivityThreshold}}{100} \quad (10.3)$$

then the station is considered active and must perform OBSS scanning over the next scan interval.

During the development of the 802.11n amendment there was some controversy surrounding the requirement that stations perform scanning. Companies with an interest in handhelds wanted those devices to be exempt from the scanning requirements citing concern regarding power consumption. However, an analysis of the additional power consumption likely from the scanning requirements (Perahia, 2007) shows that the burden is minimal, even for a handheld device in standby (waiting for a call). The compromise reached was to set an activity threshold below which a station would not be required to scan. The default activity threshold of 0.25% would require devices with an active voice over IP (VoIP) call to scan (during a VoIP call a station typically receives or sends a short frame every 20 ms), but not if the station was in standby where there is only

a small amount of data transfer to maintain connectivity or where short text messages may be sent and received. Stations can also always re-associate with the AP as 20 MHz stations (more than adequate for VoIP) to avoid the scanning requirement.

10.3.9 Signaling 40 MHz intolerance

A station operating in the 2.4 GHz band may set the Forty MHz Intolerant bit in the HT Capabilities Information element to 1 to indicate to the AP with which it is associating that it may not operate a 20/40 MHz BSS. A station may also periodically broadcast a 20/40 BSS Coexistence Management frame containing a 20/40 BSS Coexistence element with the Forty MHz Intolerant bit set to indicate to any neighboring BSSs that they may not operate a 20/40 MHz BSS. The concept of 40 MHz intolerance does not apply to stations operating in the 5 GHz bands and such stations must always set this bit to 0.

An AP with a member station that has the Forty MHz Intolerant bit set in its HT Capabilities element will switch to 20 MHz BSS operation if it is operating a 20/40 MHz BSS at the time of the association. The AP will continue with 20 MHz BSS operation as long as the station remains associated.

An AP that receives a 20/40 BSS Coexistence element with the Forty MHz Intolerant bit set must also switch to 20 MHz BSS operation. The prohibition on 20/40 MHz BSS operation continues for dot11BSSWidthChannelTransitionTime $\times$ dot11WidthTriggerScanInterval seconds after the Forty MHz Intolerant bit was last detected and an AP may resume 20/40 MHz operation if it no longer receives a Forty MHz Intolerant indication after that period.

The net effect of these rules is that a station that cannot tolerate 40 MHz operation is able to disable 40 MHz operation in the BSS to which it is associated and any neighboring BSSs that detect its periodic 20/40 BSS Coexistence Management frame transmission.

10.3.10 Channel management at the AP

The 802.11h amendment (Spectrum and Transmit Power Management Enhancements in the 5 GHz band in Europe) introduced a mechanism for moving a BSS to a different channel. A Channel Switch Announcement element and a Channel Switch Announcement frame were defined along with a procedure for performing the channel switch.

The ability to switch channels dynamically is considered an important mechanism for supporting 40 MHz coexistence with neighboring 20 MHz BSSs. In addition, it was thought that the mechanism could be used to switch channel operating widths. An AP that detects significant traffic from neighboring BSSs on the secondary channel, or significant traffic on the primary channel for that matter, could move the BSS to a channel pair with less traffic and/or narrow the operating channel width. However, the original Channel Switch Announcement element, designed for 20 MHz channels, could not handle the channel pairing used to create 40 MHz channels.

As it turned out, another task group working on the 802.11y amendment (3650-3700 MHz Operation in the USA (IEEE, 2007b)) encountered a similar limitation in that the Channel Switch Announcement element did not support switching to a new regulatory domain. To overcome these limitation and maintain backward compatibility, a new Extended Channel Switch Announcement element and Extended Channel Switch Announcement frame are defined. The new element and frame are flexible enough to support the requirements of both task groups. The Extended Channel Switch Announcement element includes a New Regulatory Class field to indicate the regulatory class of the new channel to which the BSS is switching. For 802.11n, the regulatory class codes (see Appendix 5.1) have been defined so that the channel width (20 MHz or 40 MHz) can be determined from it and, if 40 MHz wide, which 20 MHz channel is the primary and which the secondary.

The basic procedure for switching channels is as follows. The decision to switch channels is made by the AP and the AP should select a new channel that is supported by all associated stations. The AP informs associated stations that the BSS is moving to a new channel and/or changing operating channel width at some point in the future by including the Extended Channel Switch Announcement element in Beacon frames and Probe Response frames. It may also send out one or more Extended Channel Switch Announcement frames that include the Extended Channel Switch Announcement element.

The AP will attempt to schedule the channel switch so that all stations in the BSS, including stations in power save mode, have an opportunity to receive at least one Extended Channel Switch Announcement element before the switch. A scheduled channel switch occurs just before a target beacon transmission time (TBTT). The Channel Switch Count field indicates the number of TBTTs until the switch, including the TBTT just before which the switch occurs. A value of 1 indicates that the switch occurs just before the next TBTT.

An AP may indicate that the switch is to occur immediately (or anytime in the near future) by setting the Channel Switch Count value to 0. The AP may force stations in the BSS to stop transmissions until the channel switch occurs by setting the Channel Switch Mode field to 1.

0.4 A summary of fields controlling 40 MHz operation

A number of fields in various management frames control the use of 40 MHz transmissions. These fields are as follows:

- The Supported Channel Width Set field in the HT Capabilities element.
- The Secondary Channel Offset field in the HT Information element.
- The STA Channel Width field in the HT Information element.
- The STA Channel Width field of the Notify Channel Width action frame.
- The Extended Channel Switch Announcement element.

The Supported Channel Width Set field is used to indicate whether the station is capable of transmitting and receiving 40 MHz PPDU. Once a station has associated with the BSS it cannot change this value without first disassociating. The AP may change this value to indicate a new channel width for the BSS.

The Secondary Channel Offset field is used by the AP to indicate whether or not the BSS is occupying a pair of 20 MHz channels, and if it is, whether the secondary channel is above or below the primary channel in frequency. This field will only have a non-zero value, indicating the presence of a secondary channel, if the Supported Channel Width field is set to 1.

The STA Channel Width field is used to indicate a station's current operating width (both the AP and BSS member stations). A value of 0 indicates that the station is not capable of receiving or transmitting a 40 MHz PPDU. A station can only set this field to 1 if the Supported Channel Width Set field is set to 1.

The Extended Channel Width Switch Announcement element can be used to indicate a change in the operating width of the BSS. Following the change in operating width, subsequent management frames will reflect the new operating width in the Supported Channel Width Set and Secondary Channel Offset fields.

A station that is associated with a BSS or the AP itself can only send a 40 MHz PPDU if the following conditions are met:

- The Supported Channel Width Set field of both stations is set to 1.
- The Secondary Channel Offset field of the most recently received HT Information element is set to a non-zero value.
- The AP has not executed a channel switch to 20 MHz channel width. Specifically, the most recently received Extended Channel Switch Announcement element transmitted by the AP did not have the Channel Switch Count field set to 0 and a New Regulatory Class value that does not indicate a 40 MHz channel.

A station may also dynamically change its channel width from 40 MHz to 20 MHz (or vice versa). So a station will also not transmit a 40 MHz PPDU if either of the following occurred:

- The most recently received Notify Channel Width Action frame from the peer station had the STA Channel Width field set to 0.
- The most recently sent Notify Channel Width Action frame to the peer station had the STA Channel Width field set to 0.

10.5 Phased coexistence operation (PCO)

Phased coexistence operation (PCO) is an optional BSS mode with alternating 20 MHz and 40 MHz phases of operation controlled by the AP. PCO allows a 20/40 MHz BSS to operate when there are neighboring 20 MHz BSSs on both the primary and secondary 20 MHz channels making up the 40 MHz channel. In practice, because of the overhead and potential real-time disruption associated with PCO, it is likely to only be used if

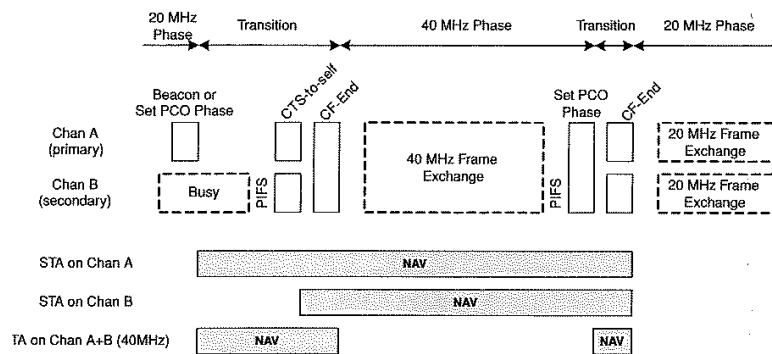


Figure 10.6 PCO 20 and 40 MHz phase transitions. Reproduced with permission from IEEE (2007a) © IEEE.

it is not possible to find a 40 MHz channel without neighboring traffic on both of the 20 MHz channels. Real-time disruption occurs with PCO because the 20 MHz stations are prevented from transmitting at all during the relatively long 40 MHz phases of operation. Also, if the neighboring 20 MHz BSSs are heavily loaded, then it may be more efficient from an overall network throughput perspective to operate in 20 MHz only.

A PCO capable AP advertises its BSS as 20/40 MHz BSS Channel Width and PCO Capable. A non-PCO capable station may associate with the BSS as it would any 20/40 MHz BSS.

A PCO capable station sets the PCO field in the HT Extended Capabilities information element to 1.

0.5.1 Basic operation

The PCO AP periodically transitions between 20 MHz and 40 MHz phases of operation using the sequences shown in Figure 10.6. To transition to 40 MHz operation, the AP sends a Beacon or Set PCO Phase management frame on the primary channel and uses it to set the NAV of 20 MHz stations operating in that channel as well as 20/40 MHz stations. The AP then waits for the secondary channel to go idle. If the secondary channel does not go idle for an excessively long time, the AP may cancel the transition to 40 MHz operation by sending another Set PCO Phase management frame.

If the secondary channel goes idle then the AP transmits a CTS-to-self on both the primary and secondary channels using the non-HT duplicate mode. The CTS-to-self sets the NAV of stations on both 20 MHz channels to cover the expected 40 MHz phase of operation. Finally, the AP sends a CF-End in a 40 MHz HT PPDU to reset the NAV of 40 MHz capable stations. The 40 MHz stations then begin contention for channel access.

The AP may extend the 40 MHz phase by transmitting a Set PCO Phase management frame in a non-HT duplicate PPDU followed by a CF-End sent in a 40 MHz HT PPDU. The Set PCO Phase management frame will set the NAV in all stations and the CF-End will reset the NAV in the 40 MHz stations so that they can compete for access again.

At some point the PCO AP begins a transition to 20 MHz operation by gaining priority channel access to send a Set PCO Phase management frame in a 40 MHz HT PPDU. It follows this with a CTS-to-self sent in a non-HT duplicate mode PPDU, which resets the NAV of 20 MHz stations and allows 20 MHz operation to resume.

During the 40 MHz phase, a 20/40 MHz PCO station transmits data frames using a 40 MHz HT PPDU and control frames using a non-HT duplicate or a 40 MHz HT PPDU, except for any CF-End frame, which can only be sent using a 40 MHz HT PPDU.

10.5.2 Minimizing real-time disruption

To minimize the access delay 20 MHz stations experience because of the 40 MHz phase of operation, the AP has a management variable `dot11PCO40MaxDuration`, which defines the maximum duration of the 40 MHz phase. Similarly, `dot11PCO20MaxDuration` defines the maximum duration of the 20 MHz phase.

Also, in order for the PCO AP to provide sufficient opportunity for stations to send frames, a minimum duration for the 40 MHz and 20 MHz phases is also defined in `dot11PCO40MinDuration` and `dot11PCO20MinDuration`, respectively.

10.6 Protection

Mechanisms are necessary to protect HT transmissions and certain HT sequences from stations that may not recognize these formats and thus not defer correctly. However, protection mechanisms add significant overhead and an attempt should be made to only use these mechanisms when needed. As with previous 802.11 amendments, it is expected that early networks will consist of a heterogeneous mix of legacy devices and 802.11n conformant devices. To complicate things, the 802.11n amendment includes many optional features and it is expected that early implementations will include varying support for many of the 802.11n optional features. Over time it is expected that implementations will become more consistent in the features supported as devices become more fully featured. It is expected that while early networks will be heterogeneous and require protection mechanisms, these mechanisms may be required less and less as networks become more fully featured and homogeneous.

The protection mechanisms described here essentially ensure that a potential interferer defers transmission for a known period allowing an HT station to complete its HT frame sequence. Multiple mechanisms are available to perform this function, some of which depend on the capabilities of the devices directly involved in the exchange, the capabilities of third party devices, and the devices against which protection is sought.

10.6.1 Protection with 802.11b stations present

Protecting new frame formats and sequences is not new to 802.11. The 802.11g amendment, which introduced the enhanced rate PHY (ERP) for OFDM operation in the 2.4 GHz ISM band, required a protection mechanism against the widely deployed 802.11b (DSSS/CCK) devices present in the band at the time. Two mechanisms were available to protect OFDM transmissions: the RTS/CTS exchange and the lower overhead CTS-to-Self transmission. Both mechanisms use legacy compatible frames to set the NAV in the legacy station(s) to cover the OFDM transmission sequence.

Because of the inherent overhead associated with both these mechanisms, it is beneficial to only require protection in the presence of legacy 802.11b stations. The mechanism used to enable protection is for the AP to set the Use_Protection bit in the ERP Information element in the Beacon and Probe Response frames. The AP sets this bit if at least one 802.11b station is associated with BSS. The AP may, but is not required to, also set this bit if it detects the presence of 802.11b stations in a neighboring BSS. A station associated with the BSS, seeing this bit set, knows to begin a sequence containing an OFDM frame with a DSSS/CCK format RTS/CTS exchange or CTS-to-Self transmission.

This protection mechanism is carried forward to HT BSSs. If an HT station sees that the Use_Protection bit is set to 1 in the ERP Information element, then it must protect non-HT and HT OFDM sequences with a DSSS/CCK format RTS/CTS exchange or CTS-to-Self transmission. In the case of a 40 MHz HT sequence the RTS/CTS exchange or CTS-to-Self transmission occurs on the primary channel.

The RTS/CTS and CTS-to-Self mechanisms are described in more detail below.

10.6.2 Protection with 802.11g or 802.11a stations present

HT transmissions are inherently protected against 802.11g and 802.11a (non-HT) stations through the use of the HT mixed format frame. This frame format includes a legacy compatible preamble, which allows non-HT stations to defer correctly for the duration of the frame.

However, some HT transmissions may not be interpreted correctly by non-HT stations and even some HT stations. In particular, non-HT stations and some HT stations may not be able to defer correctly for HT Greenfield format PPDUs. Also, certain non-HT stations may not defer correctly for burst sequence where the shorter RIFS spacing is used between frames. For these situations, additional mechanisms are needed to protect these sequences.

A broader set of protection mechanisms are available for protecting HT Greenfield format and RIFS burst transmissions when only 802.11g or 802.11a non-HT and HT stations are present, than is available for protection when 802.11b stations are present. These protection mechanisms are:

- RTS/CTS
- CTS-to-Self

- a non-HT or HT mixed format PPDU soliciting a non-HT response PPDU
- L-SIG TXOP protection.

These mechanisms offer protection with varying degrees of robustness and overhead and are discussed in more detail later in this section. To begin with though we consider first the conditions under which RIFS bursts and Greenfield format transmissions require protection.

10.6.3 Protection for OBSS legacy stations

The AP may set the OBSS Non-HT STAs Present bit in the HT Information element if it detects non-HT stations on the primary or secondary channel that are not members of the BSS. On seeing this bit set, member stations of the BSS may optionally use protection mechanisms to protect sequences for which a non-HT station may not defer correctly.

10.6.4 RIFS burst protection

Burst sequences that use RIFS may not be receivable by some legacy stations. A legacy station, designed expecting frames to be spaced at least SIFS apart, may not re-arm its acquisition circuit in time following a received frame to detect a second frame if the second frame arrives less than SIFS after the first frame. If the legacy station fails to detect the frame then it may not defer correctly for the duration frame and may begin transmitting before the frame completes.

An AP can prevent RIFS from being used in a sequence by any station associated with the BSS by setting the RIFS Mode subfield in the HT Information element to 0. A station may only use RIFS bursting if the RIFS Mode subfield is set to 1.

If RIFS bursting is permitted then the station may, but is not required to, protect RIFS sequences if there are legacy stations present. Given the overhead associated with protection it is likely that implementations would use SIFS rather than provide protection unless protection was needed for other reasons. A station knows that legacy (non-HT) stations may be present when the HT Protection subfield in the HT Information field is set to 1 (HT BSS members, non-HT non-members) or 3 (Non-HT BSS members present).

10.6.5 Greenfield format protection

The Greenfield format preamble is shorter and thus more efficient than the mixed format preamble. However, the Greenfield format preamble is not compatible with legacy stations and, support being optional, is also not receivable by some HT stations. An HT station indicates that it is not able to receive the Greenfield format by setting the Greenfield bit in the HT Capabilities element to 0.

If there are HT stations associated with the AP that are not able to receive Greenfield format PPDUs then the AP will set the Non-Greenfield HT STAs Present bit to 1 in the

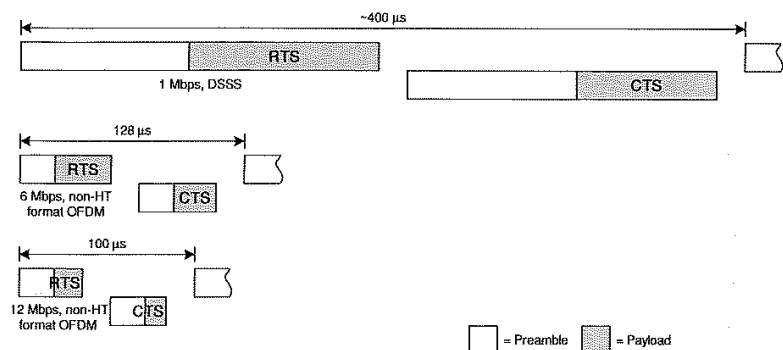


Figure 10.7 RTS/CTS overhead with DSSS and non-HT format PPDU.

HT Information element. If there are non-HT stations associated with the BSS the AP will set the HT Protection field to 3. When the Non-Greenfield HT STAs Present bit is set to 1 or the HT Protection field is set to 3 in the HT Information element then stations associated with the BSS must protect Greenfield format PPDU.

There was some opposition in the development of the 802.11n amendment with the requirement that protection is needed when non-Greenfield HT stations are present since these stations are still required to defer for Greenfield format transmissions at signal levels lower than energy detect threshold (see Sections 10.3.4 and 10.3.5).

10.6.6 RTS/CTS protection

An RTS/CTS exchange at the beginning of a TXOP sets the NAV of stations in the vicinity of both the initiator and recipient. The RTS and CTS frames are transmitted at a basic rate and are thus widely received. The robust modulation used and the fact that the widely received frames are transmitted from both ends of the link makes the RTS/CTS exchange the most robust mechanism for establishing protection.

The initiator begins a TXOP by transmitting a RTS frame, setting the Duration field in the RTS to the expected duration of the TXOP less the duration of the RTS frame itself. The responder sends a CTS frame setting the Duration field to the value seen in the RTS frame less SIFS less the duration of the CTS frame. The CTS frame is transmitted with the same modulation and coding as the RTS frame. Stations that successfully demodulate either the RTS or CTS frame or both frames have their NAV set for the duration of the TXOP and will not transmit during that time.

Figure 10.7 shows the relative overhead of an RTS/CTS exchange using 1 Mbps DSSS format, 6 Mbps non-HT OFDM format and 12 Mbps non-HT OFDM format, PPDU. The DSSS format PPDU may be used in the presence of 802.11b stations. However,

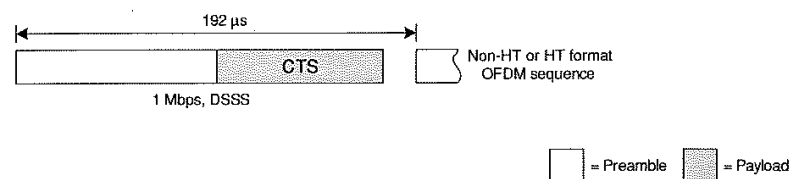


Figure 10.8 CTS-to-Self overhead with 1 Mbps DSSS (short preamble) format PPDU.

because of the high overhead of this exchange, it is more likely that the CTS-to-Self mechanism will be used since it is rare that protection needs to be established from both ends of the link, especially at DSSS rates. The non-HT format exchange may be used to protect HT Greenfield format and RIFS burst sequences in the presence of 802.11a or 802.11g stations. The 12 Mbps rate offers slightly reduced robustness over the 6 Mbps rate, but with lower overhead.

10.6.7 CTS-to-Self protection

CTS-to-Self (Figure 10.8) is a MAC level mechanism for protecting frame sequences from stations in the vicinity of the initiator. This was originally introduced with the 802.11g amendment to protect OFDM transmissions in the 2.4 GHz ISM band from 802.11b stations which were only capable of detecting DSSS/CCK transmissions. In most networks all stations are able to detect transmissions of all other stations and hidden nodes are rare. For such situations, the CTS-to-Self mechanism offered significantly reduced overhead compared with a full RTS/CTS exchange.

At the beginning of a sequence, the initiator sends a CTS frame with the RA field set to its own MAC address and the Duration field set to the expected duration of the sequence less the duration of the CTS frame itself. The CTS frame is transmitted using a basic rate PPDU compatible with the legacy stations present and against which protection is desired.

In practice, CTS-to-Self may only be used to protect a Data/ACK or MMPDU/ACK sequence since TXOP rules require that burst sequences and A-MPDU sequences have a short frame exchange performed at the start of the TXOP as a collision detect mechanism.

CTS-to-Self sent in a non-HT OFDM format PPDU may be used by HT stations that wish to protect Greenfield data/ACK sequences in the presence of 802.11g or 802.11a stations, however, the mandatory existence of the HT mixed format preamble and its lower overhead obviate this use. Thus it is likely that CTS-to-Self will continue to only be used when 802.11b stations are present.

10.6.8 Protection using a non-HT or HT mixed PPDU with non-HT response

The NAV of nearby stations is set by any frame that is correctly received for which the RA is not the address of the station receiving the frame. Thus a sequence that begins

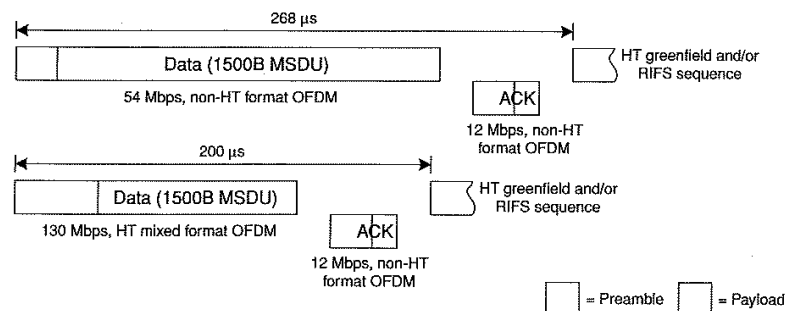


Figure 10.9 A non-HT or HT mixed format PPDU soliciting a non-HT PPDU response.

with a legacy compatible frame, for example a data frame, and that solicits a response frame, an ACK frame in this example, could be used to set NAV in nearby stations. In the example sequence in Figure 10.9, the data frame is likely transmitted using a high order MCS, however, the ACK frame is robustly modulated using a non-HT format PPDU and should be widely received. The ACK frame could thus be used to carry the NAV setting for nearby stations.

The initiator would set the Duration field of the data frame to the expected duration of the TXOP less the duration of the data frame itself. The ACK response would have its Duration field set to the value of the Duration field in the data frame less SIFS less the duration of the ACK frame itself. With NAV established in the surrounding stations, the remainder of the TXOP could include HT Greenfield format PPDU and/or RIFS burst sequences.

10.6.9 Non-HT station deferral with HT mixed format PPDU

The HT mixed format preamble is designed to be compatible with legacy 802.11g and 802.11a stations in the sense that a legacy station receiving the frame will defer for the duration indicated by the length (L_LENGTH) and rate (L_Datarate) given in the signaling field (L-SIG) of the legacy compatible portion of the preamble. This is illustrated in Figure 10.10.

The L_Datarate field is set to indicate a rate of 6 Mbps, which is the lowest OFDM PHY data rate. The L_LENGTH field is set to indicate the desired duration. A legacy station would interpret the L_Datarate and L_LENGTH fields as follows. At 6 Mbps each 4 μs symbol carries 3 octets (24 bits). At this rate, the overhead in the payload in the 16-bit SERVICE field (which carries the scrambler initialization) and 6-bit Tail (used to flush the decoder) account for one symbol.¹ The payload duration would thus

¹ Actually 22 bits, but since the PSDU is an integral number of octets the rounding works.

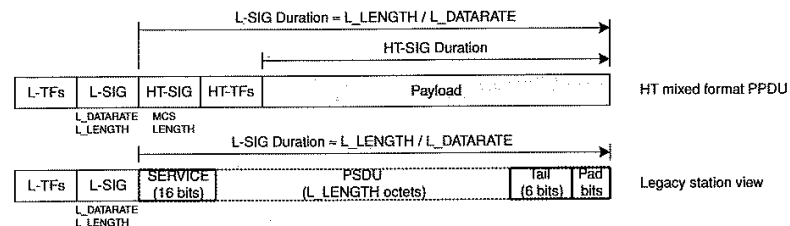


Figure 10.10 Non-HT station defer with HT mixed format PPDU.

be:

$$L\text{-SIG duration} = \left(1 + \left\lceil \frac{L_LENGTH}{3} \right\rceil \right) \times 4 \mu s \quad (10.4)$$

where $\lceil \cdot \rceil$ indicates a rounding up to the next integer value. Since the maximum L_LENGTH value is 4095, an HT station could indicate an L-SIG duration of up to 5.464 ms on 4 μs boundaries. An HT station must set the L_LENGTH and L_Datarate to a value that indicates an L-SIG duration of at least the duration of the frame.

It is worth noting that some legacy devices may not defer correctly for L_LENGTH field values greater than 2340. The legacy signaling field is weakly protected by a single parity bit. Many implementations perform additional checks to ensure that the frame has been correctly detected and the signaling field successfully demodulated, including checking that the length field is less than 2340 (the maximum MPDU size prior to 802.11n). These implementations would interpret a large length value as a false detect and not assert CCA if the signal strength is below the energy detect threshold.

10.6.10 L-SIG TXOP protection

For an HT station receiving an HT mixed format PPDU, the L_LENGTH and L_Datarate in the legacy signaling field are not needed to demodulate the HT portion of the frame since the actual PSDU length and MCS used are specified in the HT signaling field (HT-SIG). The L_Datarate and L_LENGTH may thus be used to signal a duration that is longer than the actual frame duration and in so doing protect a sequence rather than just a single frame. This is the basis for L-SIG protection and is illustrated in Figure 10.11.

An HT station sets the L-SIG duration to at least cover the duration of the HT PPDU. If the HT station is sending the PPDU to an HT station that supports L-SIG TXOP protection, the station may set the L-SIG duration to cover a sequence of frames. Since a legacy station is not able to receive another frame for the full L-SIG duration, none of the frames in the protected sequence should be addressed to a legacy station.

An HT station that supports L-SIG TXOP protection and that receives an HT mixed format PPDU will set its NAV at the end of the PPDU to the remaining L-SIG duration

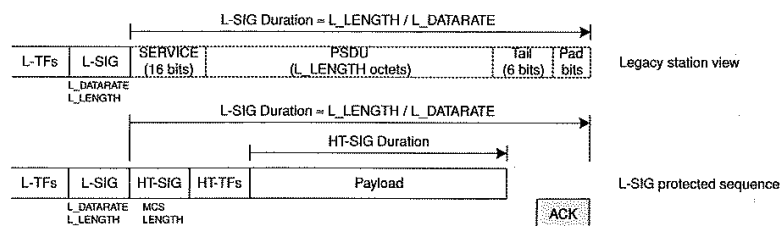


Figure 10.11 L-SIG protection.

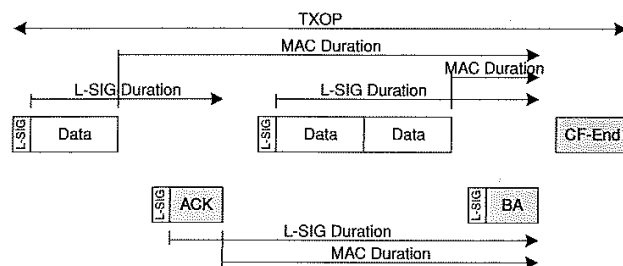


Figure 10.12 Example of an L-SIG TXOP protected sequence. Reproduced with permission from IEEE (2007a) © IEEE.

if the HT station could not successfully extract an MPDU from the PPDU with which it would normally update its NAV using the MPDU's Duration/ID field.

An example of an L-SIG protected sequence is given in Figure 10.12. An L-SIG protected sequence begins with an initial handshake, which is the exchange of two short frames each of which are in HT mixed format PPDUs. Any initial frame sequence may be used that is valid for the start of a TXOP provided the duration of the response frame is known beforehand by the initiator.

The L-SIG duration of the initial PPDU is set to cover the duration of the initial PPDU less the legacy compatible preamble plus SIFS plus the duration of the response frame. The Duration field in the MAC frame carried by the PPDU is set to the expected duration of the remainder of the sequence.

The L-SIG duration in the response PPDU will be set to the duration indicated in the Duration field from the MAC frame less SIFS less the duration of the legacy compatible portion of the preamble.

Following the successful completion of the initial handshake the sequence proceeds similarly using HT mixed format PPDUs. The L-SIG Duration of the subsequent PPDU from the initiator has its L-SIG set to cover the remainder of the sequence. The Duration fields in the MAC frames carried by the PPDU also indicate a value that extends to the

end of the sequence. The response frames use the Duration field in the MAC frames to set the L-SIG duration.

A drawback to the use of L-SIG TXOP protection is that legacy stations do not successfully demodulate any of the frames and will use an EIFS defer for channel access at the end of the protected period. This would put them at a disadvantage in gaining channel access as compared to HT stations which may successfully demodulate the final HT mixed format PPDU. To prevent this, the initiator should follow the protected sequence with a CF-End frame using a basic rate non-HT PPDU.

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Octets	1	1	1	Variable
	Element ID	Length	Regulatory Class	Channel List

Figure 11.26 20/40 BSS Intolerant Channel Report element. Reproduced with permission from IEEE (2007b) © IEEE.

Octets	1	1	2	2	2	2	2	2	2
	Element ID	Length	OBSS Scan Passive Dwell	OBSS Scan Active Dwell	BSS Width Trigger Scan Interval	OBSS Scan Passive Total Per Channel	OBSS Scan Active Total Per Channel	BSS Channel Transition Delay Factor	OBSS Scan Activity Threshold

Figure 11.27 Overlapping BSS Scan Parameters element. Reproduced with permission from IEEE (2007b) © IEEE.

11.3.2.5 20/40 BSS Intolerant Channel Report element

The format of the 20/40 BSS Intolerant Channel Report element is shown in Figure 11.26. This element contains a list of channels on which the station has found conditions that disallow the operation of a 20/40 MHz BSS. The Regulatory Class field contains the regulatory class in which the channel list is valid. The Channel List field contains one or more octets where each octet provides a channel number. The regulatory classes and associated channel numbers are listed in Appendix 5.1.

This element may appear zero or more times in the 20/40 BSS Coexistence Management frame.

11.3.2.6 Overlapping BSS Scan Parameters element

The format of the Overlapping BSS Scan Parameters element is shown in Figure 11.27. This element is used by the AP in a 20/40 MHz BSS to set the parameters controlling OBSS scanning in member stations.

References

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IEEE (2007c). *IEEE P802.11y™/D6.0, Draft Amendment to STANDARD for Information Technology – Telecommunications and Information Exchange Between Systems – Local and Metropolitan Networks – Specific Requirements. Part 11: Wireless LAN Medium Access Control (MAC) and Physical Layer (PHY) Specifications: Amendment 3: 3650–3700 MHz Operation in USA.*

To request feedback, a station sets the MRQ (MCS request) field to 1 in the HT Control field and chooses a value between 0 and 6 for the MSI (MCS request sequence identifier). The MSI is used to correlate the response with the request in the case of a delayed response and the value chosen is implementation dependent. In the MCS feedback response, the responder sets the MFSI (MCS feedback sequence identifier) to the value of the MSI in the corresponding MCS request. When the responder provides unsolicited MCS feedback, the MFSI value is set to 7.

If the HT Control field is included in more than one MPDU in an aggregate then the MRQ and MSI fields are set to the same values and act effectively as a single request. The HT Control field should be included in all frames making up an aggregate to improve robustness.

The MCS request should be sent in a staggered sounding PPDU or it should be sent with the NDP Announcement field set to 1 with an NDP transmission to follow. The number of HT-LTFs in the sounding PPDU or in the NDP is determined by the total number of spatial dimensions to be sounded, including any extra spatial dimensions beyond those used by the data portion of the PPDU.

On receipt of an MCS request, the responder should compute an MCS estimate. The responder may choose to send the response frame with any of the following combination of MFB (MCS feedback) and MFSI:

- MFB = 127, MFSI = 7. No information is provided for the immediately proceeding request or for any other pending request.
- MFB = 127, MFSI in range 0 to 6. The responder is unable to provide feedback.
- MFB in range 0 to 126, MFSI in range 0 to 6. The responder is providing feedback for the previously received request with MSI equal to MFSI.
- MFB in range 0 to 126, MFSI = 7. The responder is providing unsolicited feedback.

Hardware constraints may limit the number of outstanding MCS requests a responder can handle. When a new MCS request arrives, either from a different requester or from the same requester but with a different MSI value, the responder may choose to ignore the request or discard a current request and begin computation on the new request. If the responder discards a pending MCS estimate computation it should return a response with MFB set to 127 and MFSI set to the same value as the corresponding MSI in the MCS request.

The responder is constrained in the MCS suggestion it can make. It cannot make an MCS suggestion that includes more special streams than are supported by the requester. It should also not make an MCS suggestion for unequal modulation unless the requester indicates that it is capable of unequal modulation in the Tx Unequal Modulation Supported bit in the Supported MCS Set field (see the HT Capabilities element and 'Supported MCS Set field', p. 291).

References

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Appendix 12.1: Unequal MCS

This appendix gives unequal MCS tables for 20 MHz and 40 MHz. For 20 MHz, the number of data subcarriers is 52 and the number of pilot subcarriers is 4 in an OFDM symbol for an individual spatial stream. For 40 MHz, the number of data subcarriers is 108 and the number of pilot subcarriers is 6 in an OFDM symbol for an individual spatial stream.

Table 12.6 describes the symbols used in the subsequent tables.

Table 12.6 Symbols used for MCS parameters (IEEE, 2007)

Symbol	Explanation
R	Code rate
N_{TBPS}	Total bits per subcarrier
N_{CBPS}	Number of coded bits per OFDM symbol
N_{DBPS}	Number of data bits per OFDM symbol
N_{ES}	Number of BCC encoders

Unequal MCS for 20 MHz

Table 12.7 defines the 20 MHz unequal MCS parameters and data rates for two spatial stream transmission. The data rate is calculated by dividing N_{DBPS} by the symbol time of 4 μs for 800 ns GI, and dividing by 3.6 μs for 400 ns GI.

Table 12.8 defines the 20 MHz unequal MCS parameters and data rates for three spatial stream transmission.

Table 12.9 defines the 20 MHz unequal MCS parameters and data rates for four spatial stream transmission. All 20 MHz MCSs use a single encoder.

Table 12.7 20 MHz unequal MCS parameters for two spatial streams (IEEE, 2007)

MCS Index	Modulation		R	N_{TBPS}	N_{CBPS}	N_{DBPS}	Data rate (Mbps)	
	Stream 1	Stream 2					800 ns GI	400 ns GI
33	16-QAM	QPSK	$\frac{1}{2}$	6	312	156	39	43.3
34	64-QAM	QPSK	$\frac{1}{2}$	8	416	208	52	57.8
35	64-QAM	16-QAM	$\frac{1}{2}$	10	520	260	65	72.2
36	16-QAM	QPSK	$\frac{3}{4}$	6	312	234	58.5	65.0
37	64-QAM	QPSK	$\frac{3}{4}$	8	416	312	78	86.7
38	64-QAM	16-QAM	$\frac{3}{4}$	10	520	390	97.5	108.3

Table 12.8 20 MHz unequal MCS parameters for three spatial streams (IEEE, 2007)

MCS Index	Modulation			R	N_{TBPS}	N_{CBPS}	N_{DBPS}	Data rate (Mbps)	
	Stream 1	Stream 2	Stream 3					800 ns GI	400 ns GI
39	16-QAM	QPSK	QPSK	$\frac{1}{2}$	8	416	208	52	57.8
40	16-QAM	16-QAM	QPSK	$\frac{1}{2}$	10	520	260	65	72.2
41	64-QAM	QPSK	QPSK	$\frac{1}{2}$	10	520	260	65	72.2
42	64-QAM	16-QAM	QPSK	$\frac{1}{2}$	12	624	312	78	86.7
43	64-QAM	16-QAM	16-QAM	$\frac{1}{2}$	14	728	364	91	101.1
44	64-QAM	64-QAM	QPSK	$\frac{1}{2}$	14	728	364	91	101.1
45	64-QAM	64-QAM	16-QAM	$\frac{1}{2}$	16	832	416	104	115.6
46	16-QAM	QPSK	QPSK	$\frac{3}{4}$	8	416	312	78	86.7
47	16-QAM	16-QAM	QPSK	$\frac{3}{4}$	10	520	390	97.5	108.3
48	64-QAM	QPSK	QPSK	$\frac{3}{4}$	10	520	390	97.5	108.3
49	64-QAM	16-QAM	QPSK	$\frac{3}{4}$	12	624	468	117	130.0
50	64-QAM	16-QAM	16-QAM	$\frac{3}{4}$	14	728	546	136.5	151.7
51	64-QAM	64-QAM	QPSK	$\frac{3}{4}$	14	728	546	136.5	151.7
52	64-QAM	64-QAM	16-QAM	$\frac{3}{4}$	16	832	624	156	173.3

Table 12.9 20 MHz unequal MCS parameters for four spatial streams (IEEE, 2007)

MCS Index	Modulation				R	N_{TBPS}	N_{CBPS}	N_{DBPS}	Data rate (Mbps)	
	Stream 1	Stream 2	Stream 3	Stream 4					800 ns GI	400 ns GI
53	16-QAM	QPSK	QPSK	QPSK	$\frac{1}{2}$	10	520	260	65	72.2
54	16-QAM	16-QAM	QPSK	QPSK	$\frac{1}{2}$	12	624	312	78	86.7
55	16-QAM	16-QAM	16-QAM	QPSK	$\frac{1}{2}$	14	728	364	91	101.1
56	64-QAM	QPSK	QPSK	QPSK	$\frac{1}{2}$	12	624	312	78	86.7
57	64-QAM	16-QAM	QPSK	QPSK	$\frac{1}{2}$	14	728	364	91	101.1
58	64-QAM	16-QAM	16-QAM	QPSK	$\frac{1}{2}$	16	832	416	104	115.6
59	64-QAM	16-QAM	16-QAM	16-QAM	$\frac{1}{2}$	18	936	468	117	130.0
60	64-QAM	64-QAM	QPSK	QPSK	$\frac{1}{2}$	16	832	416	104	115.6

Table 12.9 (cont.)

MCS Index	Modulation				R	N_{TBPS}	N_{CBPS}	N_{DBPS}	Data rate (Mbps)	
	Stream 1	Stream 2	Stream 3	Stream 4					800 ns GI	400 ns GI
61	64-QAM	64-QAM	16-QAM	QPSK	$\frac{1}{2}$	18	936	468	117	130.0
62	64-QAM	64-QAM	16-QAM	16-QAM	$\frac{1}{2}$	20	1040	520	130	144.4
63	64-QAM	64-QAM	64-QAM	QPSK	$\frac{1}{2}$	20	1040	520	130	144.4
64	64-QAM	64-QAM	64-QAM	16-QAM	$\frac{1}{2}$	22	1144	572	143	158.9
65	16-QAM	QPSK	QPSK	QPSK	$\frac{3}{4}$	10	520	390	97.5	108.3
66	16-QAM	16-QAM	QPSK	QPSK	$\frac{3}{4}$	12	624	468	117	130.0
67	16-QAM	16-QAM	16-QAM	QPSK	$\frac{3}{4}$	14	728	546	136.5	151.7
68	64-QAM	QPSK	QPSK	QPSK	$\frac{3}{4}$	12	624	468	117	130.0
69	64-QAM	16-QAM	QPSK	QPSK	$\frac{3}{4}$	14	728	546	136.5	151.7
70	64-QAM	16-QAM	16-QAM	QPSK	$\frac{3}{4}$	16	832	624	156	173.3
71	64-QAM	16-QAM	16-QAM	16-QAM	$\frac{3}{4}$	18	936	702	175.5	195.0
72	64-QAM	64-QAM	QPSK	QPSK	$\frac{3}{4}$	16	832	624	156	173.3
73	64-QAM	64-QAM	16-QAM	QPSK	$\frac{3}{4}$	18	936	702	175.5	195.0
74	64-QAM	64-QAM	16-QAM	16-QAM	$\frac{3}{4}$	20	1040	780	195	216.7
75	64-QAM	64-QAM	64-QAM	QPSK	$\frac{3}{4}$	20	1040	780	195	216.7
76	64-QAM	64-QAM	64-QAM	16-QAM	$\frac{3}{4}$	22	1144	858	214.5	238.3

Unequal MCS for 40 MHz

Table 12.10 defines the 40 MHz unequal MCS parameters and data rates for two spatial stream transmission. A single encoder is used for all two stream MCSs.

Table 12.10 40 MHz unequal MCS parameters for two spatial streams (IEEE, 2007)

MCS Index	Modulation				R	N_{BPS}	N_{SD}	N_{SP}	N_{CBPS}	N_{DBPS}	Data rate (Mb/s)	
	Stream 1	Stream 2	R	N_{BPS}							800 ns GI	400 nsec GI
33	16-QAM	QPSK	$\frac{1}{2}$	6	108	6	648	324	81	90		
34	64-QAM	QPSK	$\frac{1}{2}$	8	108	6	864	432	108	120		
35	64-QAM	16-QAM	$\frac{1}{2}$	10	108	6	1080	540	135	150		
36	16-QAM	QPSK	$\frac{3}{4}$	6	108	6	648	486	121.5	135		
37	64-QAM	QPSK	$\frac{3}{4}$	8	108	6	864	648	162	180		
38	64-QAM	16-QAM	$\frac{3}{4}$	10	108	6	1080	810	202.5	225		

Table 12.11 defines the 40 MHz unequal MCS parameters and data rates for three spatial stream transmission. A single encoder is used for every three stream MCS, except for MCS 52. Table 12.12 defines the 40 MHz unequal MCS parameters and data rates for two spatial stream transmission. A single encoder is used for MCS 53–69 and two encoders are used for MCS 70–76.

Table 12.7 20 MHz unequal MCS parameters for two spatial streams (IEEE, 2007)

MCS Index	Modulation		R	N_{TBPS}	N_{CBPS}	N_{DBPS}	Data rate (Mbps)	
	Stream 1	Stream 2					800 ns GI	400 ns GI
33	16-QAM	QPSK	$\frac{1}{2}$	6	312	156	39	43.3
34	64-QAM	QPSK	$\frac{1}{2}$	8	416	208	52	57.8
35	64-QAM	16-QAM	$\frac{1}{2}$	10	520	260	65	72.2
36	16-QAM	QPSK	$\frac{3}{4}$	6	312	234	58.5	65.0
37	64-QAM	QPSK	$\frac{3}{4}$	8	416	312	78	86.7
38	64-QAM	16-QAM	$\frac{3}{4}$	10	520	390	97.5	108.3

Table 12.8 20 MHz unequal MCS parameters for three spatial streams (IEEE, 2007)

MCS Index	Modulation			R	N_{TBPS}	N_{CBPS}	N_{DBPS}	Data rate (Mbps)	
	Stream 1	Stream 2	Stream 3					800 ns GI	400 ns GI
39	16-QAM	QPSK	QPSK	$\frac{1}{2}$	8	416	208	52	57.8
40	16-QAM	16-QAM	QPSK	$\frac{1}{2}$	10	520	260	65	72.2
41	64-QAM	QPSK	QPSK	$\frac{1}{2}$	10	520	260	65	72.2
42	64-QAM	16-QAM	QPSK	$\frac{1}{2}$	12	624	312	78	86.7
43	64-QAM	16-QAM	16-QAM	$\frac{1}{2}$	14	728	364	91	101.1
44	64-QAM	64-QAM	QPSK	$\frac{1}{2}$	14	728	364	91	101.1
45	64-QAM	64-QAM	16-QAM	$\frac{1}{2}$	16	832	416	104	115.6
46	16-QAM	QPSK	QPSK	$\frac{3}{4}$	8	416	312	78	86.7
47	16-QAM	16-QAM	QPSK	$\frac{3}{4}$	10	520	390	97.5	108.3
48	64-QAM	QPSK	QPSK	$\frac{3}{4}$	10	520	390	97.5	108.3
49	64-QAM	16-QAM	QPSK	$\frac{3}{4}$	12	624	468	117	130.0
50	64-QAM	16-QAM	16-QAM	$\frac{3}{4}$	14	728	546	136.5	151.7
51	64-QAM	64-QAM	QPSK	$\frac{3}{4}$	14	728	546	136.5	151.7
52	64-QAM	64-QAM	16-QAM	$\frac{3}{4}$	16	832	624	156	173.3

Table 12.9 20 MHz unequal MCS parameters for four spatial streams (IEEE, 2007)

MCS Index	Modulation				R	N_{TBPS}	N_{CBPS}	N_{DBPS}	Data rate (Mbps)	
	Stream 1	Stream 2	Stream 3	Stream 4					800 ns GI	400 ns GI
53	16-QAM	QPSK	QPSK	QPSK	$\frac{1}{2}$	10	520	260	65	72.2
54	16-QAM	16-QAM	QPSK	QPSK	$\frac{1}{2}$	12	624	312	78	86.7
55	16-QAM	16-QAM	16-QAM	QPSK	$\frac{1}{2}$	14	728	364	91	101.1
56	64-QAM	QPSK	QPSK	QPSK	$\frac{1}{2}$	12	624	312	78	86.7
57	64-QAM	16-QAM	QPSK	QPSK	$\frac{1}{2}$	14	728	364	91	101.1
58	64-QAM	16-QAM	16-QAM	QPSK	$\frac{1}{2}$	16	832	416	104	115.6
59	64-QAM	16-QAM	16-QAM	16-QAM	$\frac{1}{2}$	18	936	468	117	130.0
60	64-QAM	64-QAM	QPSK	QPSK	$\frac{1}{2}$	16	832	416	104	115.6

Table 12.9 (cont.)

MCS Index	Modulation				R	N_{TBPS}	N_{CBPS}	N_{DBPS}	Data rate (Mbps)	
	Stream 1	Stream 2	Stream 3	Stream 4					800 ns GI	400 ns GI
61	64-QAM	64-QAM	16-QAM	QPSK	$\frac{1}{2}$	18	936	468	117	130.0
62	64-QAM	64-QAM	16-QAM	16-QAM	$\frac{1}{2}$	20	1040	520	130	144.4
63	64-QAM	64-QAM	64-QAM	QPSK	$\frac{1}{2}$	20	1040	520	130	144.4
64	64-QAM	64-QAM	64-QAM	16-QAM	$\frac{1}{2}$	22	1144	572	143	158.9
65	16-QAM	QPSK	QPSK	QPSK	$\frac{3}{4}$	10	520	390	97.5	108.3
66	16-QAM	16-QAM	QPSK	QPSK	$\frac{3}{4}$	12	624	468	117	130.0
67	16-QAM	16-QAM	16-QAM	QPSK	$\frac{3}{4}$	14	728	546	136.5	151.7
68	64-QAM	QPSK	QPSK	QPSK	$\frac{3}{4}$	12	624	468	117	130.0
69	64-QAM	16-QAM	QPSK	QPSK	$\frac{3}{4}$	14	728	546	136.5	151.7
70	64-QAM	16-QAM	16-QAM	QPSK	$\frac{3}{4}$	16	832	624	156	173.3
71	64-QAM	16-QAM	16-QAM	16-QAM	$\frac{3}{4}$	18	936	702	175.5	195.0
72	64-QAM	64-QAM	QPSK	QPSK	$\frac{3}{4}$	16	832	624	156	173.3
73	64-QAM	64-QAM	16-QAM	QPSK	$\frac{3}{4}$	18	936	702	175.5	195.0
74	64-QAM	64-QAM	16-QAM	16-QAM	$\frac{3}{4}$	20	1040	780	195	216.7
75	64-QAM	64-QAM	64-QAM	QPSK	$\frac{3}{4}$	20	1040	780	195	216.7
76	64-QAM	64-QAM	64-QAM	16-QAM	$\frac{3}{4}$	22	1144	858	214.5	238.3

Unequal MCS for 40 MHz

Table 12.10 defines the 40 MHz unequal MCS parameters and data rates for two spatial stream transmission. A single encoder is used for all two stream MCSs.

Table 12.10 40 MHz unequal MCS parameters for two spatial streams (IEEE, 2007)

MCS Index	Modulation				R	N_{BPSK}	N_{SD}	N_{SP}	N_{CBPS}	N_{DBPS}	Data rate (Mb/s)	
	Stream 1	Stream 2									800 ns GI	400 nsec GI
33	16-QAM	QPSK	$\frac{1}{2}$	6	108	6	648	324	81	90		
34	64-QAM	QPSK	$\frac{1}{2}$	8	108	6	864	432	108	120		
35	64-QAM	16-QAM	$\frac{1}{2}$	10	108	6	1080	540	135	150		
36	16-QAM	QPSK	$\frac{3}{4}$	6	108	6	648	486	121.5	135		
37	64-QAM	QPSK	$\frac{3}{4}$	8	108	6	864	648	162	180		
38	64-QAM	16-QAM	$\frac{3}{4}$	10	108	6	1080	810	202.5	225		

Table 12.11 defines the 40 MHz unequal MCS parameters and data rates for three spatial stream transmission. A single encoder is used for every three stream MCS, except for MCS 52. Table 12.12 defines the 40 MHz unequal MCS parameters and data rates for two spatial stream transmission. A single encoder is used for MCS 53–69 and two encoders are used for MCS 70–76.

Table 12.11 40 MHz unequal MCS parameters for three spatial streams (IEEE, 2007)

MCS Index	Modulation			R	N _{BPSK}	N _{SD}	N _{SP}	N _{Cbps}	N _{Dbps}	N _{FES}	Data rate (Mbps)	
	Stream 1	Stream 2	Stream 3								800 ns GI	400 ns GI
39	16-QAM	QPSK	QPSK	1/2	8	108	6	864	432	1	108	120
40	16-QAM	16-QAM	QPSK	1/2	10	108	6	1080	540	1	135	150
41	64-QAM	QPSK	QPSK	1/2	10	108	6	1080	540	1	135	150
42	64-QAM	16-QAM	QPSK	1/2	12	108	6	1296	648	1	162	180
43	64-QAM	16-QAM	16-QAM	1/2	14	108	6	1512	756	1	189	210
44	64-QAM	64-QAM	QPSK	1/2	14	108	6	1512	756	1	189	210
45	64-QAM	64-QAM	16-QAM	1/2	16	108	6	1728	864	1	216	240
46	16-QAM	QPSK	QPSK	3/4	8	108	6	864	432	1	162	180
47	16-QAM	16-QAM	QPSK	3/4	10	108	6	1080	540	1	202.5	225
48	64-QAM	QPSK	QPSK	3/4	10	108	6	1080	540	1	202.5	225
49	64-QAM	16-QAM	QPSK	3/4	12	108	6	1296	648	1	243	270
50	64-QAM	16-QAM	16-QAM	3/4	14	108	6	1512	756	1	283.5	315
51	64-QAM	64-QAM	QPSK	3/4	14	108	6	1512	756	1	283.5	315
52	64-QAM	64-QAM	16-QAM	3/4	16	108	6	1728	864	2	324	360

Table 12.12 40 MHz unequal MCS parameters for four spatial streams (IEEE, 2007)

MCS Index	Modulation				R	N _{BPSK}	N _{Cbps}	N _{Dbps}	N _{FSS}	Data rate (Mbps)	
	Stream 1	Stream 2	Stream 3	Stream 4						800 ns GI	400 ns GI
53	16-QAM	QPSK	QPSK	QPSK	$\frac{1}{2}$	10	1080	540	1	135	150
54	16-QAM	16-QAM	QPSK	QPSK	$\frac{1}{2}$	12	1296	648	1	162	180
55	16-QAM	16-QAM	16-QAM	QPSK	$\frac{1}{2}$	14	1512	756	1	189	210
56	64-QAM	QPSK	QPSK	QPSK	$\frac{1}{2}$	12	1296	648	1	162	180
57	64-QAM	16-QAM	QPSK	QPSK	$\frac{1}{2}$	14	1512	756	1	189	210
58	64-QAM	16-QAM	16-QAM	QPSK	$\frac{1}{2}$	16	1728	864	1	216	240
59	64-QAM	16-QAM	16-QAM	16-QAM	$\frac{1}{2}$	18	1944	972	1	243	270
60	64-QAM	64-QAM	QPSK	QPSK	$\frac{1}{2}$	16	1728	864	1	216	240
61	64-QAM	64-QAM	16-QAM	QPSK	$\frac{1}{2}$	18	1944	972	1	243	270
62	64-QAM	64-QAM	16-QAM	16-QAM	$\frac{1}{2}$	20	2160	1080	1	270	300
63	64-QAM	64-QAM	64-QAM	QPSK	$\frac{1}{2}$	20	2160	1080	1	270	300
64	64-QAM	64-QAM	64-QAM	16-QAM	$\frac{1}{2}$	22	2376	1188	1	297	330
65	16-QAM	QPSK	QPSK	QPSK	$\frac{3}{4}$	10	1080	540	1	202.5	225
66	16-QAM	16-QAM	QPSK	QPSK	$\frac{3}{4}$	12	1296	648	1	243	270
67	16-QAM	16-QAM	16-QAM	QPSK	$\frac{3}{4}$	14	1512	756	1	283.5	315
68	64-QAM	QPSK	QPSK	QPSK	$\frac{3}{4}$	12	1296	648	1	243	270
69	64-QAM	16-QAM	QPSK	QPSK	$\frac{3}{4}$	14	1512	756	1	283.5	315
70	64-QAM	16-QAM	16-QAM	QPSK	$\frac{3}{4}$	16	1728	864	1	324	360
71	64-QAM	16-QAM	16-QAM	16-QAM	$\frac{3}{4}$	18	1944	972	2	364.5	405
72	64-QAM	64-QAM	QPSK	QPSK	$\frac{3}{4}$	16	1728	864	2	324	360
73	64-QAM	64-QAM	16-QAM	QPSK	$\frac{3}{4}$	18	1944	972	2	364.5	405
74	64-QAM	64-QAM	16-QAM	16-QAM	$\frac{3}{4}$	20	2160	1080	2	405	450
75	64-QAM	64-QAM	64-QAM	QPSK	$\frac{3}{4}$	20	2160	1080	2	405	450
76	64-QAM	64-QAM	64-QAM	16-QAM	$\frac{3}{4}$	22	2376	1188	2	445.5	495

If you've been searching for a way to get up to speed quickly on IEEE 802.11n without having to wade through the entire standard, then look no further. This comprehensive overview describes the underlying principles, implementation details, and key enhancing features of 802.11n. For many of these features, the authors outline the motivation and history behind their adoption into the standard. A detailed discussion of the key throughput, robustness, and reliability enhancing features (such as MIMO, 40 MHz channels, and packet aggregation) is given, in addition to a clear summary of the issues surrounding legacy interoperability and coexistence. Advanced topics such as beamforming and fast link adaption are also covered. With numerous MAC and physical layer examples and simulation results included to highlight the benefits of the new features, this is an ideal reference for designers of WLAN equipment, and network managers whose systems adopt the new standard. It is also a useful distillation of 802.11n technology for graduate students and researchers in the field of wireless communication.

"Books like this one are invaluable because they provide the details and the background that allow readers to answer the questions, "What is likely to be in the final standard and how does it work?" Eldad and Robert should be congratulated on taking up the challenge."

Andrew Myles, Chairman of WiFi Alliance.

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