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First Report: State of the art and research on AMC for concatenated coded multi-carrier systems and related technologies

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Abstract:

The report aims to present the current status in the research area of Adaptive Modulation and Coding and propose directions for future common research among the partners of WPR.3. It includes a literature review of the state of the art along with examples of adaptivity implemented in past and emerging systems, and reviews link level modeling methodologies which are of great interest for the design of optimization algorithms. The last section presents the challenges to be addressed by the WPR.3 partners classified in four categories based on the target system of each effort.

Keyword list: adaptive, modulation, channel coding, link level modeling, PHY abstraction, OFDM, non-orthogonal, MIMO, STC, CPM

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1. INTRODUCTION

The increasing demand for better spectral utilization and higher QoS requirements motivate the design of increasingly more intelligent and agile communication systems, able to adapt and adjust (in real-time) the transmission parameters based on the current link quality for the ultimate goal of reaching, to the degree possible, the inherent capacity of the underlying channel. Referred to as Adaptive Modulation and Coding (AMC), many such techniques have been proposed that target the best use of available resources. This report reviews the State of the Art in AMC research and proposes few key topics for further investigation within WPR.3.

More specifically, in chapter 2 a historical review of AMC is presented, highlighting the philosophy, the targets, the considerations/limitations and the main achievements in AMC in the past years. In chapter 3 AMC-related functionalities of past and emerging systems and standards are presented.

As the new upcoming standards/systems are becoming ever more complex so as to enable multi-modal operation in a variety of operational scenarios (typical examples of this trend are the WINNER project and the WiMax standard presented in Chapter 3), the AMC design problem becomes correspondingly more complex. In such systems, compact physical-layer abstraction models have been developed (as part of an Evaluation Methodology procedure - EVM) for the purpose of predicting the performance of a given link in the presence of a large number of impacting parameters. Such methodologies accelerate the corresponding system-level simulations whenever a large number of physical-layer-related links need to be factored in the simulation. These PHY abstraction methodologies are highly relevant to the AMC algorithmic design process, namely an optimization process which employs this link-level modeling functionality either as the main objective function under optimization or as one of the design constraints. The respective methodology for link-level performance modeling of Wimax and WINNER is reviewed in chapter 4. Additionally, the OpenAirInterface Simulation/Emulation Methodology (developed by Eurecom) is presented which, among other things, proposes PHY-layer abstraction models similar to those developed in the aforementioned two.

In chapter 5 the challenges to be addressed in WPR3 are presented, following a classification on the basis of the targeted system (in the sense of the underlying modulation type). More specifically, the targeted systems for AMC design exploration are: (a) Coded OFDM systems, (b) Non-orthogonal multicarrier systems, (c) Continuous-Phase Modulation (CPM)-based systems, and (d) single-carrier systems employing coherent QAM modulation. In each of these sections selected research topics are presented, each incorporating a small review of the current state of the art, the proposed methodology and the potential objectives.

1.1 - Glossary

2G	2nd Generation
3G	3rd Generation
ACM	Adaptive Coding and Modulation
AM	Adaptive Modulation
AMC	Adaptive Modulation and Coding
AP	Access point
ARPAnet	Advanced Research Projects Agency Network
ARQ	Automatic Repeat reQuest
ASINR	Average Signal to Interference plus Noise Ratio
AWGN	Additive White Gaussian Noise
BCC	Binary convolutional code
B-EFDMA	Block-Equidistant Frequency Division Multiple Access
BEO	Bit Error Outage
BEP	Bit Error Probability
BER	Bit Error Rate
BICM	Bit-Interleaved Coded Modulation
B-IFDMA	Block-Interleaved Frequency Division Multiple Access
BLDPC	Block-circulant Low-Density Parity-Check (codes)
BLER	Block Error Rate
BPSK	Binary phase shift keying
BS	Base Station
CC	Chase Combining
CDMA	Code Division Multiple Access
CER	Codeword Error Rate
CP	Cyclic Prefix
CPE	Continuous-Phase Encoder
CPM	Continuous-Phase Modulation
CQI	Channel Quality Information
CQI	Channel Quality Indicator
CQICH	Channel Quality Indicator CHannel
CRC	Cyclic Redundancy Check
CSD	Cyclic shift delay
CSI	Channel State Information
CU	Control Unit
CWER	CodeWord Error Rate
D8PSK	Differential 8-PSK
DAB	Digital Audio Broadcasting
DFT	Discrete Fourier Transform
DQPSK	Differential QPSK
DVB-RCT	Digital Video Broadcasting - Return Channel Terrestrial
DVB-T	Digital Video Broadcasting Terrestrial
e.g.	exempli gratia
EESM	Exponential-Effective SINR Mapping
ESM	Effective SINR Mapping
ETSI	European Telecommunications Standards Institute
EVM	EValuation Methodologies
FAM	Fast Adaptive Modulation
FBMCM	Filter-Bank Multicarrier Modulation
FDMA	Frequency Division Multiple Access
FEC	Forward Error Correction
FER	Frame Error Rate

GI	Guard interval
GMC	Generalized Multicarrier
GSM	Global System for Mobile radio
HARQ	Hybrid ARQ
HDSPA	High-Speed Downlink Packet Access
HPA	High-Power Amplifier
HT	Hilly Terrain
HT	High Throughput
i.e.	id est
ICI	Interchannel Interference
IDFT	Inverse Discrete Fourier Transform
IEEE	Institute of Electrical and Electronics Engineers
IOTA	Isotropic Orthogonal Transform Algorithm
IP	Internet Protocol
IR	Incremental Redundancy
ISI	Intersymbol Interference
L2S	Link to System (Interface)
LA	Link Adaptation
LAN	Local Area Network
LDPC	Low density parity check code
LLC	Logical Link Control
LLR	Log-Likelihood Ratio
MAC	Medium Access Control
MAI	Multiple Access Interference
MAP	Maximum A-Posteriori
MC	Multi-Carrier
MCS	Modulation and coding scheme
MI-ACM	Mutual Information based Adaptive Coding and Modulation
MIESM	Mutual Information Effective SINR Metric
MIMO	Multiple Input Multiple Output
ML	Maximum Likelihood
MM	Memoryless Modulator
MRC	Maximal Ratio Combining
MS	Mobile Station
MSK	Minimum-Shift Keying
MT	Mobile Terminal
MUD	Multi-User Diversity
NACK	No-ACKnowledgment
NES	Number of encoder streams
NOMCM	Non Orthogonal Multicarrier Modulations
OFDM	Orthogonal Frequency Division Multiplexing
OFDMA	Orthogonal Frequency Division Multiple Access
OQAM	Offset Quadrature Amplitude Modulation
PAMR	Public Access Mobile Radio
PAPR	Peak-to-Average Power Ratio
P-CPM	Pragmatic Continuous-Phase Modulation
PDU	Protocol Data Unit
PEP	Pair-wise Error Probability
PER	Packet Error Rate
PHY	PHYSical layer
PLCP	Physical layer convergence protocol
PMR	Personal Mobile Radio
PPDU	PLCP protocol data unit

PSK	Phase-Shift Keying
QAM	Quadrature Amplitude Modulation
QoS	Quality of Service
QPSK	Quadrature Phase-Shift Keying
RCP	Rate-Compatible Punctured
RCPC	Rate Compatible Punctured Convolutional codes
RCPT	Rate Compatible Punctured Turbo codes
RF	Radio Frequency
RLC	Radio Link Control
RSSI	Received Signal Strength Indicator
SAM	Slow Adaptive Modulation
SAP	Service access point
SB	Statistical Beamforming
SC	Single-Carrier
SC-CPM	Serially-Concatenated Continuous-Phase Modulation
SE	Spectral Efficiency
SINR	Signal to Interference plus Noise Ratio
SISO	Single Input Single Output
SNR	Signal to Noise Ratio
SRRC	Square Root Raised Cosine
SSD	Subset Diversity
STA	Station
STBC	Space-time block coding
TDD	Time-Division Duplex
TDMA	Time Division Multiple Access
TEDS	TETRA Enhanced Data Service
TETRA	TERrestrial Trunked RADio
TF	Time-Frequency
TU	Typical Urban
vs.	versus
WF	Water-Filling
WINNER	Wireless World Initiative New Radio
WLAN	Wireless Local Area Network
WSSUS	Wide-Sense Stationary Uncorrelated Scattering
ZF	Zero-Forcing

2. A LITERATURE REVIEW OF ADAPTIVE MODULATION AND CODING

Adaptive Modulation and Coding (AMC) techniques target the best use of available resources (system bandwidth, channelization, transmit power, time slots, computational power of executing platform) in order to achieve a specified Quality of Service (QoS) level [Big02]. Based on inputs from higher layers (namely, target-QoS parameters pertaining to the physical layer) plus Channel State Information--CSI (or its estimate), the optimized transmission parameters (in this case, a pair of modulation constellation size and the coding-scheme parameters such as block size and code rate) are computed and then employed in the transmission. The choice of the values of the proper parameters is often not trivial and depends strongly on the defined cost function (the so-called "adaptation criterion") underlying the optimization algorithm. Moreover, the AMC algorithm performance depends also strongly on the accuracy of the CSI, which is in turn affected by the estimation errors and the delay incurred in the feedback channel.

The AMC strategy is defined by the following set: an adaptation criterion or cost function, the target QoS parameters, the CSI (often imperfect, erroneous or obsolete), and the resulting outputs from the AMC algorithm (namely, a set of transmission parameters). A study of possible adaptation strategies can be found in [Kla03] with extensive bibliography on the subject, as well as a review on the link adaptation research history. As far as the adaptation criteria are concerned, there are multiple possibilities and combinations of AMC schemes, in which some particular metrics and quantities may be inputs (requirements and constraints) for one strategy or the criterion/output for the other. Some aspects (like data throughput) may be introduced as the criterion (objective function), while others (like transmission power limit and delay) may be introduced as constraints. By varying these choices, a large number of adaptation strategies is obtained. For example, in the throughput-oriented strategy, the AMC algorithm aims in providing the highest bit rate (or spectral efficiency) for a required BER and fixed radiated power limit. This was one of the the first adaptive transmission scheme proposed by Steele and Webb for single-carrier QAM modulation and narrow-band fading channels [Stee91]. Following this idea of exploiting the time-variant channel capacity, various concatenated coded schemes with an adaptive coding rate have been investigated in [Mor97]; variable coding rate and power schemes in [Soo97], [Gol97a], [Gol97b], [Seo01], [Sye01]; latency and interference aspects with turbo-coded adaptation in [Han02], [Kel00], as well as many other important aspects of link adaptation. The concepts elaborated for adaptive QAM modulation and coding have been invoked for OFDM QAM in [Cho95], [Czy96], [Fis96]. Adaptive subcarrier selection for OFDM TDMA dynamic links has been investigated in [Roh96]. Further significant advances have been made in the direction of adaptive subcarrier allocation in multi-user adaptive OFDM [Gmun02], [Pfl02] space-time diversity [Cat01], [Lie01] and multi-coded systems [Dlu03], [Wes02], as well as in the investigation of the key agents affecting AMC performance [She01], [Hol02], [Lam02a], [Lam02b], [Mmun02], [Fal03], [Fal04]. Apart from the techniques proposed in the above mentioned publications, interesting proposals for throughput-oriented AMC algorithms for various systems and standards can be found in [Ior01], [Cha01], [Fei01], [Tra01], [Sen01].

The more general approach to AMC provides more than just these two possibilities (the constellation size and the code rate) for parameter adjustment. For example, transmit power, symbol rate, the number of subcarriers, the number of antennas and other parameters can be adjusted to achieve any given optimization goal. The key issue is which of these parameters should be adapted to obtain the required performance at reasonable complexity. For example, the Shannon capacity limit of a flat fading channel can be achieved by varying both the radiated power plus the bit rate [Soo97]. Interestingly, the achievable capacity is negligibly smaller if only one of these two parameters is varied alone [Soo97], [Cai99]. An interesting study on maximizing the spectral efficiency by optimally varying combinations of the transmission rate and radiated power with average power and instantaneous (or average) BER constraints has been published in [Seo01]. There, both continuous-rate adaptation (where the set of signal constellation points is unrestricted) as well as discrete-rate adaptation (where only a discrete finite set of constellation points is available) are considered. There, again, the conclusion has been reached that the use of only one or two degrees of freedom in the adaptive technique yields close to the maximum possible spectral efficiency that would be obtained by utilizing all degrees of freedom.

Reference DR.3.1

The overall power consumption of a transceiver for a negotiated level of service is a typical and popular adjustable parameter within the framework of AMC, especially for energy-constrained wireless networks. The major motivation behind this type of optimization is to prolong the battery life and to minimize the impact of electromagnetic radiation on human health. For this criterion, a terminal should be configured and the transmission parameters should be adapted in order to minimize both the power needed for the baseband signal processing plus the power radiated by the transmitting antenna. Thus, the philosophy of the AMC algorithm design in this case should take the computational complexity and required processing power into account. To this end, some trade-offs in choosing the optimization criteria are needed, such as minimization of the transmitted power within a given limit for the baseband processing, as well as adoption of some simplifications in the link description in order to perform less complex computing towards a sub-optimal but acceptable solution. Interesting propositions for power-oriented AMC algorithms and strategies can be found in [Soo97], [Eph02], [Min02], [Pol03], [Dlu02], [Won99].

The above considerations are related mostly to the single-link and single-user case. There exist in literature interesting proposals concerning multi-user adaptive schemes. In multi-user systems (for instance, in adaptive FDMA) subcarriers may be adaptively allocated to users in an optimal (or in a fair) manner, taking the quality of each user's channel into account. As an example we may consider the algorithm described in [Won99] which aims to minimize the overall transmit power by adaptive multi-user subcarrier, bit, and power allocation. Furthermore, in a multi-user scenario, handling the delay-sensitive traffic becomes a scheduling issue. The most straightforward approach is to prioritize delay-sensitive traffic such that the delay constraints are fulfilled [Joh04]. In situations where the channel quality might be poor for extended periods over the whole bandwidth, more resources such as transmit power or the number of sub-carriers must be allocated to delay-sensitive streams to guarantee on-time delivery of data.

3. AMC IN CURRENT AND UPCOMING SYSTEMS

Research in the AMC area in the last two decades resulted in the successful implementation of optimization procedures in commercial standards/systems. In this section the AMC procedures of selected systems/standards are reviewed.

3.1 WiMax (IEEE 802.16e – 2005)

3.1.1 Overview

The WiMAX collection of standards supports a variety of modulation and coding schemes and allows for the scheme to change on a burst-by-burst basis per link, depending on channel conditions. Using the channel quality feedback indicator, the mobile can provide the base station with feedback on the downlink channel quality. For the uplink, the base station can estimate the channel quality based on the received signal quality. The base station scheduler can take into account the channel quality of each user's uplink and downlink and assign a modulation and coding scheme that maximizes the throughput for the available signal-to-noise ratio. AMC methods significantly increase the overall system capacity as they allow real-time trade-offs between throughput and robustness on each link.

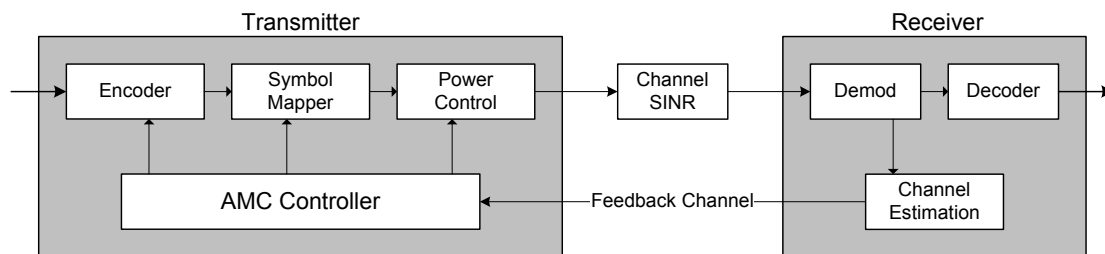


Figure 3.1.1: AMC block diagram

The challenge here is to efficiently control three quantities at once: transmit power, constellation size/type used, and the overall coding scheme/rate. This translates to developing an appropriate policy for the AMC controller (Figure 3.1.1). In the idealized case, perfect channel knowledge will be available to the transmitter. However, in practice the information feedback will incur some delay and perhaps also be degraded due to imperfect channel estimation or errors and noise in the feedback channel. Although reasonable guidelines can be developed for adaptive modulation from a theoretical standpoint, in practice the system engineer needs to develop and fine-tune the algorithm based on extensive and realistic simulations, since true performance depends on many factors. The content of this section is based on the references [And07], [IEEE802.16e05], [Yag04].

3.1.2 Supported modes of operation

In the last released WiMAX version (IEEE 802.16e – 2005) the following modes of operation are supported: in the downlink, QPSK, 16 QAM, and 64 QAM are mandatory for both fixed and mobile WiMAX; 64 QAM is optional in the uplink. FEC coding using convolutional codes is mandatory. Convolutional codes are combined with an outer Reed-Solomon code in the downlink for OFDM-PHY. The standard optionally supports turbo codes and low-density parity check (LDPC) codes at a variety of code rates as well. A total of 52 combinations of modulation and coding schemes are defined in WiMAX as burst profiles. Most practical implementations of WiMAX offer only a fraction of these.

Burst profiles in IEEE 802.16e-2005					
0	QPSK CC 1/2	19	64 QAM CTC 2/3	38	QPSK LDPC(B) 2/3
1	QPSK CC 3/4	20	64 QAM CTC 3/4	39	QPSK LDPC(B) 3/4
2	16 QAM CC 1/2	21	64 QAM CTC 5/6	40	16 QAM LDPC(B) 2/3
3	16 QAM CC 3/4	22	QPSK ZCC 1/2	41	16 QAM LDPC(B) 3/4
4	64 QAM CC 1/2	23	QPSK ZCC 3/4	42	64 QAM LDPC(B) 2/3

5	64 QAM CC 2/3	24	16 QAM ZCC 1/2	43	64 QAM LDPC(B) 3/4
6	64 QAM CC 3/4	25	16 QAM ZCC 3/4	44	QPSK CC(I) 1/2
7	QPSK BTC 1/2	26	64 QAM ZCC 1/2	45	QPSK CC(I) 3/4
8	QPSK BTC 3/4	27	64 QAM ZCC 2/3	46	16 QAM CC(I) 1/2
9	16 QAM BTC 3/5	28	64 QAM ZCC 3/4	47	16 QAM CC(I) 3/4
10	16 QAM BTC 4/5	29	QPSK LDPC 1/2	48	64 QAM CC(I) 2/3
11	64 QAM BTC 5/8	30	QPSK LDPC 2/3	49	64 QAM CC(I) 3/4
12	64 QAM BTC 4/5	31	QPSK LDPC 3/4	50	QPSK LDPC 5/6
13	QPSK CTC 1/2	32	16 QAM LDPC 1/2	51	16 QAM LDPC 5/6
14	reserved	33	16 QAM LDPC 2/3	52	64 QAM LDPC 5/6
15	QPSK CTC 3/4	34	16 QAM LDPC 3/4		>52 reserved
16	16 QAM CTC 1/2	35	64 QAM LDPC 1/2		
17	16 QAM CTC 3/4	36	64 QAM LDPC 2/3		
18	64 QAM CTC 1/2	37	64 QAM LDPC 3/4		

CC: Convolutional Code

BTC: Block Turbo Codes

CTC: Convolutional Turbo Code

ZCC: Zero-terminating Convolutional Code

LDPC: LDPC with use of A code

LDPC(B): LDPC with use of B code

CC(I): CC with use of optional Interleaver

3.1.3 Feedback mechanisms

Feedback is critical for adaptive modulation and coding: The transmitter needs to know the “channel SINR” (signal-to-interference-plus-noise ratio) γ , which is defined as the received SINR γ_r , divided by the transmit power P_t . Thus the received SINR is $\gamma_r = P_t \gamma$. More specifically in IEEE 802.16e the channel-quality measurements for the AMC are RSSI (received signal strength indicator) and SINR that the MS (mobile station) is required to provide to the BS (base station) on request. The MS uses the Channel Quality feedback Indicator channel (CQI) to provide the BS with this information. Based on the CQI, the BS can either:

- Change modulation and/or coding rate for the transmissions: change the *burst profile*
- Change the power level of the associated DL transmissions

or do both. Due to the dynamic nature of the wireless channel, both the mean and the standard deviation of the RSSI and SINR are included in the definition of CQI. The RSSI measurement as defined by the IEEE 802.16e-2005 standard does not require the receiver to actively demodulate the signal, thus reducing the amount of processing power required. When requested by the BS, the MS measures the instantaneous RSSI. A series of measured instantaneous RSSI values are used to derive the mean and standard deviation of the RSSI. The SINR measurements, unlike the RSSI measurement, require active demodulation of the signal and are usually a better indicator of true channel quality. The mean and the standard deviation of the SINR are converted to the dB scale before being reported to the BS.

The receiver calculates the effective SNR per subchannel (a group of subcarriers) and provides that information to the transmitter, using the 6-bit CQI channel (CQICH). The periodicity of the SNR feedback can be varied according to the Doppler spread of the channel. When closed-loop MIMO (see below) is used, the receiver also uses the CQICH or the fast feedback channel to provide feedback needed for closed-loop MIMO or beamforming. WiMAX systems heavily protect the feedback channel with error correction, so the main source of degradation is usually mobility which causes channel estimates to rapidly become obsolete. Empirically, with speeds greater than about 30 km/hr on a 2,100MHz carrier, even the faster feedback configurations do not allow timely and accurate channel state information to be available at the transmitter.

3.1.4 Closed-Loop MIMO

Closed-Loop MIMO operation is included in IEEE 802.16e - 2005, and it involves a signalling framework where partial or full Channel state information (CSI) is used at the BS which could be also used by the AMC, if needed (it does not belong to the AMC framework).

CSI at the transmitter can be used to select the appropriate MIMO mode (number of transmit antennas, number of simultaneous streams, and space/time encoding matrix) as well as to calculate an optimum precoding matrix that maximizes system capacity. The CSI can be known at the transmitter due to channel reciprocity, in the case of TDD, or by having a feedback channel, in the case of FDD. The uplink bandwidth required to provide the full CSI to the transmitter, i.e the MIMO channel matrix for each subcarrier in a multiuser FDD MIMO-OFDM system, is too large and thus impractical for a closed-loop FDD MIMO system. For practical systems, it is possible only to send some form of quantized information in the uplink. The framework for closed-loop MIMO, as shown in Figure 3.1.2, consists of a space/time encoding stage identical to an open-loop system and a MIMO precoding stage. The MIMO precoding matrix in general is a complex matrix, with the number of rows equal to the number of transmit antennas and the number of columns equal to the output of the space/time encoding block. The linear precoding matrix spatially mixes the various parallel streams among the various antennas, with appropriate amplitude and phase adjustment.

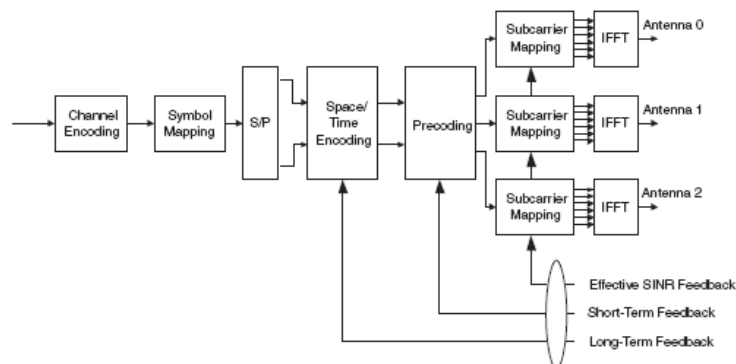


Figure 3.1.2: Closed-loop MIMO framework in 802.16e-2005

In order to determine the appropriate amplitude and phases of the various weights, the transmitter requires some feedback from the MS. In the case of closed-loop MIMO, the feedback falls broadly into two categories: long-term feedback and short-term feedback. The long-term feedback provides information related to the maximum number of parallel streams: the rank of the precoding matrix to be used for DL transmissions. The short-term feedback provides information about the precoding matrix weights to be used. The IEEE 802.16e-2005 standard defines the following five mechanisms so that the BS can estimate the optimum precoding matrix for closed-loop MIMO operations:

1. **Antenna selection:** The MS indicates to the BS which transmit antenna(s) should be used for transmission in order to maximize the channel capacity and/or improve the link reliability.
2. **Antenna grouping:** The MS indicates to the BS the optimum permutation of the order of the various antennas to be used with the current space/time encoding matrix.
3. **Codebook based feedback:** The MS indicates to the BS the optimum precoding matrix to be used, based on the entries of a predefined codebook.
4. **Quantized channel feedback:** The MS quantizes the MIMO channel and sends this information to the BS, using the MIMO_FEEDBACK message. The BS can use the quantized MIMO channel to calculate an optimum precoding matrix.
5. **Channel sounding:** The BS obtains exact information about the CSI of the MS by using a dedicated and predetermined signal intended for channel sounding. The channel-sounding option for closed-loop MIMO operation is the most bandwidth intensive MIMO channel-feedback mechanism, but it provides the BS with the most accurate estimate of the DL channel, thus providing maximum capacity gain over.

3.2 WiFi (IEEE 802.11n)

3.2.1 Overview

The focus is on the 802.11n Draft Amendment to IEEE 802.11 WLAN standard [IEEE802.11n]. The amendment incorporates PHY and MAC features to support a throughput of 100 Mb/s and greater in the downlink, as measured at the MAC data Service Access Point (SAP). The HT PHY uses an OFDM air interface.

Some PHY features that distinguish an HT STA from a non-HT STA and an HT AP from a non-HT AP include: MIMO operation (spatial multiplexing), transmit beamforming, STBC encoding, LDPC encoding, and antenna selection. The amendment provides a new HT greenfield PPDU format. The PPDU may be transmitted with 20 MHz or 40 MHz bandwidth. The HT PHY supports operation in 2.4 GHz and 5 GHz bands. The HT PHY can support 4 spatial streams, operating in 20 MHz bandwidth. Additionally, transmission using 1 to 4 spatial streams is defined for operation in 40 MHz bandwidth. These features are capable of supporting data rates up to 600 Mb/s (4 spatial streams, 40 MHz bandwidth).

The transmitter block diagram for a HT PHY system is shown in Fig.3.2.1.

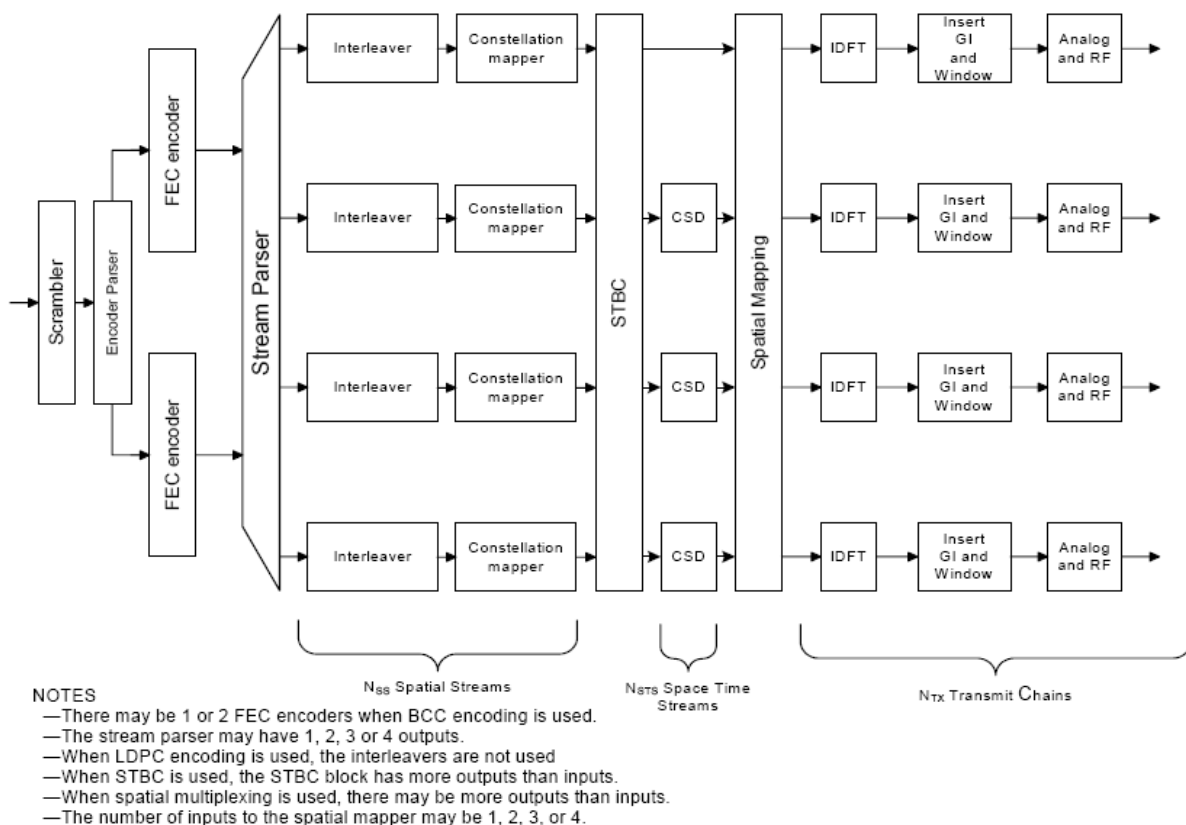


Figure 3.2.1: 802.11n PHY block diagram.

The scrambling block scrambles the data to reduce the probability of long sequences of zeros or ones. The scrambled data is encoded using either a binary convolutional code (BCC) or a low density parity check (LDPC) code. A single FEC encoder is always used when LDPC coding is used. When the BCC FEC encoder is used, a single encoder is used, except that two encoders shall be used when the selected MCS has a PHY rate greater than 300 Mb/s. The encoder parser de-multiplexes the scrambled bits into NES (1 or 2) streams depending on the number of BCC encoders to be used. The stream parser distributes the outputs of FEC blocks into spatial streams that are sent to different interleaver and mapping blocks. Interleaving is only applied when BCC encoding is used. Next, constellation mapping is used, followed by space time block encoding and spatial mapping to transmit chains. The Reference DR.3.1

options for spatial mapping are: 1) Direct mapping: constellation points from each space time stream are mapped directly onto the transmit chains (one-to-one mapping), 2) Spatial expansion: vectors of constellation points from all the space time streams are expanded via matrix multiplication to produce the input to all the transmit chains, 3) Beamforming: similar to spatial expansion: each vector of constellation points from all the space time streams is multiplied by a matrix of steering vectors to produce the input to the transmit chains. The IDFT, CSD, GI and windowing blocks give the final OFDM signals. The CSD insertion prevents unintentional beamforming.

3.2.2 Supported modes of operation

The HT PHY data subcarriers are modulated using binary phase shift keying (BPSK), quadrature phase shift keying (QPSK), 16-quadrature amplitude modulation (16-QAM), or 64-QAM. Forward error correction coding (convolutional coding) is used with a coding rate of 1/2, 2/3, 3/4, or 5/6. Low-density parity-check (LDPC) codes are added. Other features at both transmit and receive sides include transmit beamforming, antenna selection and space-time block codes (STBC). There are a total of 77 MCS options. Some parameters that are used in the MCS options are listed in the following table.

MCS parameters

Symbol	Explanation
NSS	Number of spatial streams
R	Coding rate
NBPSC	Number of coded bits per single carrier (total accross all streams)
NBPSCS	Number of coded bits per single carrier for each spatial stream
NSD	Number of data subcarriers
NSP	Number of pilot subcarriers
NCBPS	Number of coded bits per OFDM symbol
NDBPS	Number of data bits per OFDM symbol
NES	Number of BCC encoders for the DATA field
NTBPS	Total bits per subcarrier

Some MCS options for mandatory 20 MHz, $N_{SS} = 1$, $N_{ES} = 1$, EQM are shown in the following table.

MCS Index	Modulation	R	$N_{BPSCS}(i_{SS})$	N_{SD}	N_{SP}	N_{CBPS}	N_{DBPS}	Data rate (Mb/s)	
								800 ns GI	400 ns GI see NOTE
0	BPSK	1/2	1	52	4	52	26	6.5	7.2
1	QPSK	1/2	2	52	4	104	52	13.0	14.4
2	QPSK	3/4	2	52	4	104	78	19.5	21.7
3	16-QAM	1/2	4	52	4	208	104	26.0	28.9
4	16-QAM	3/4	4	52	4	208	156	39.0	43.3
5	64-QAM	2/3	6	52	4	312	208	52.0	57.8
6	64-QAM	3/4	6	52	4	312	234	58.5	65.0
7	64-QAM	5/6	6	52	4	312	260	65.0	72.2

NOTE—Support of 400 ns guard interval is optional on transmit and receive

Some MCS options for optional 40 MHz, $N_{ss} = 4$ MCSs, with unequal modulation of the spatial streams are shown in the following table.

MCS Index	Modulation				R	N_{BPSC}	N_{SD}	N_{SP}	N_{CBPS}	N_{DBPS}	N_{ES}	Data rate (Mb/s)	
	Stream 1	Stream 2	Stream 3	Stream 4								800 ns GI	400 ns GI
53	16-QAM	QPSK	QPSK	QPSK	1/2	10	108	6	1080	540	1	135	150
54	16-QAM	16-QAM	QPSK	QPSK	1/2	12	108	6	1296	648	1	162	180
55	16-QAM	16-QAM	16-QAM	QPSK	1/2	14	108	6	1512	756	1	189	210
56	64-QAM	QPSK	QPSK	QPSK	1/2	12	108	6	1296	648	1	162	180
57	64-QAM	16-QAM	QPSK	QPSK	1/2	14	108	6	1512	756	1	189	210
58	64-QAM	16-QAM	16-QAM	QPSK	1/2	16	108	6	1728	864	1	216	240

3.2.3 Feedback mechanisms

Several types of feedback mechanisms are provided for link adaptation. To fully exploit MIMO channel variations and transmit beamforming on a MIMO link, a STA can request that another STA provide MIMO channel sounding and MCS feedback, of which there are three types: Immediate, Delayed, and Unsolicited.

In order for a beamformer to calculate an appropriate steering matrix for transmit spatial processing when transmitting to a specific beamformee, the beamformer needs to have an accurate estimate of the channel that it is transmitting over. There are two methods defined as follows:

— *Implicit feedback*: When using implicit feedback, the beamformer receives long training symbols transmitted by the beamformee, which allow the MIMO channel between the beamformee and beamformer to be estimated.

— *Explicit feedback*: When operating explicit feedback, the beamformee makes a direct estimate of the channel from training symbols sent to the beamformee by the beamformer. The beamformee may prepare Channel State Information or Steering feedback based on an observation of these training symbols.

Feedback mechanisms exist also for implementing antenna-selection algorithms. Antenna selection is a time-variant mapping of the signals at the RF chains onto a set of antenna elements, when the number of RF chains is smaller than the number of antenna elements. Antenna selection requires the training of the full size channel associated with all antenna elements, which is obtained by transmitting or receiving sounding signals over all antennas.

3.3 3GPP-HSDPA

3.3.1 Overview

The 3GPP standardization group has introduced a new air interface to increase the spectral efficiency of current 3G cellular network. The new air interface is called HSDPA (High Speed Downlink Packet Access) [3GPP TR 25.855]. To maximize the system capacity, HSDPA enables several smart AMC techniques to adapt the transmission parameters to the changing channel propagation conditions. As

Reference DR.3.1

described in [3GPP TR 25.855], [3GPP TR 25.848], the objective of AMC is to change the modulation and coding format in accordance with the fluctuations of the mobile channel conditions. In general, the propagation channel can be estimated exploiting a feedback from the receiver. Thus, in an AMC system, terminals in favourable positions, e.g. users close to the cell site or in a good channel state, are typically served with higher order modulation and higher code rates, while users in unfavourable positions or in bad channel state are assigned lower order modulation with lower code rates. The main benefits of using AMC in this air interface are:

- higher data rates are available for users in favourable positions which in turn increases the average throughput of the cell
- reduced interference variation due to link adaptation based on variations in the modulation/coding scheme instead of variations in transmit power

AMC combined with time domain scheduling offers the opportunity to take advantage of short term channel variations, so that the terminal is always being served on a constructive fade event. In particular, the figure below shows the Rayleigh fading envelope correlation vs. time delay for different values of Doppler frequency. It shall be noted that for lower Doppler frequencies it is possible to schedule a user on a constructive fade provided that the scheduling interval (i.e. frame size) is small and the measurement reports are timely (i.e. distributed scheduling).

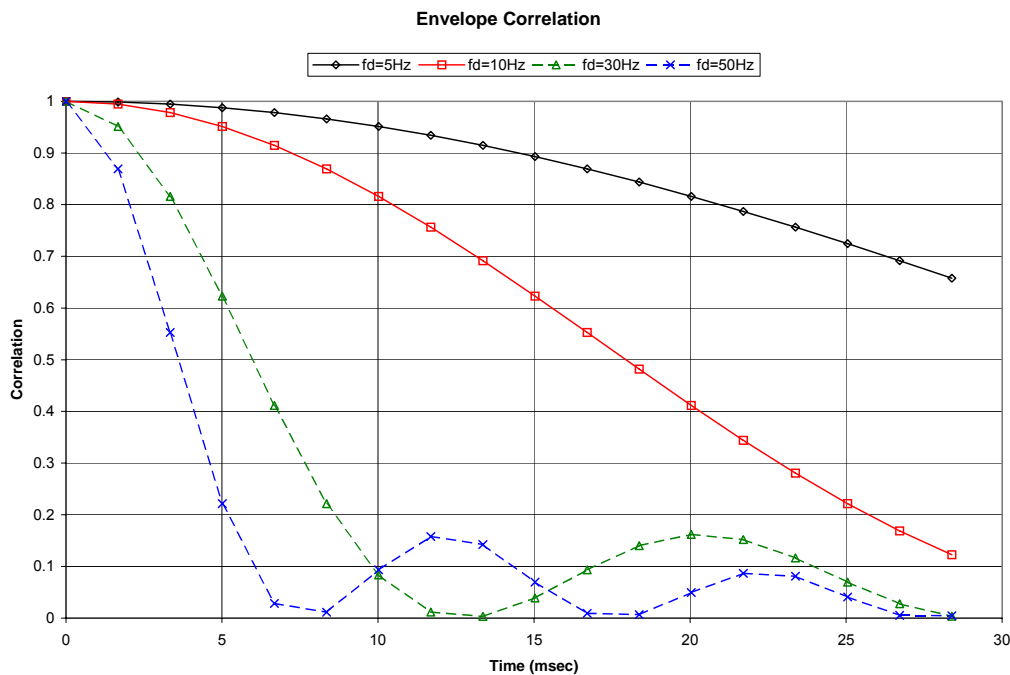


Figure 3.3.1: Envelope Correlation as a function of different Doppler (extracted from [3GPP TR 25.848]).

3.3.2 Supported modes of operation

A block diagram showing the AMC principle in the HSDPA transmitter is reported below.

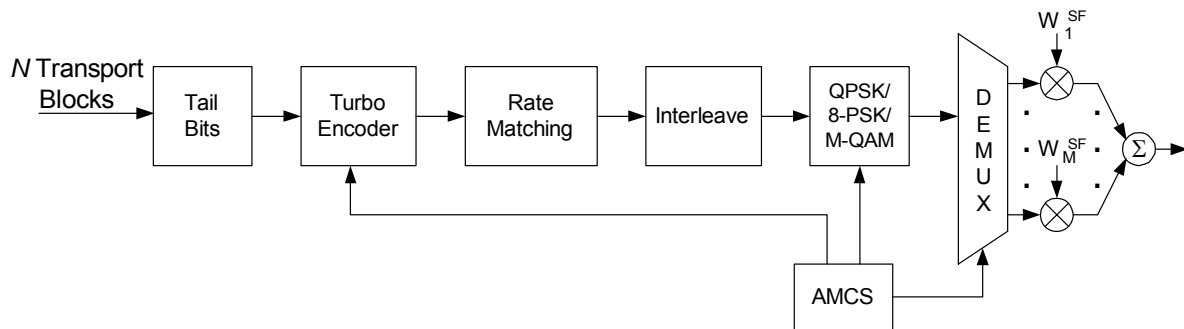


Figure 3.3.2: HSDPA physical layer structure (extracted from [3GPP TR 25.848]).

In the following table peak data rates for different parameter settings (modulation, code rate, number of codes, and number of timeslots) for the TDD mode are presented as they were calculated in [3GPP TR 25.848]:

		1 Timeslot	1 Timeslot	12 Timeslots	13 Timeslots
		1 Code	12 Codes	12 Codes	14 Codes
Modulation	Coderate	(kbps)	(Mbps)	(Mbps)	(Mbps)
64	$\frac{3}{4}$	62,1	0,745	8,94	11,3
16	$\frac{3}{4}$	41,4	0,497	5,96	7,53
16	$\frac{1}{2}$	27,6	0,331	3,97	5,02
8	$\frac{3}{4}$	31	0,372	4,46	5,64
4	$\frac{3}{4}$	20,7	0,248	2,97	3,76
4	$\frac{1}{2}$	13,8	0,166	1,99	2,51
4	$\frac{1}{4}$	6,9	0,083	1	1,26

3.4 TETRA 2

3.4.1 Overview

The core of the TERrestrial Trunked RADio (TETRA) standard developed by the European Telecommunications Standards Institute (ETSI) has been available since 1994. From then on, the ETSI Technical Committee has been continuously enhancing the standard (the IP packet data capability was added, for example, in 2000), with the aim of providing a state-of-the-art digital mobile radio system to professional users. The rapid upwards trend in the data speed of mobile radio networks and the increasing popularity of Internet “contents” and multimedia services prompted ETSI to develop a Release 2 of the TETRA standard, with the aim of reaching at least a ten-fold increase with respect to the current uncoded data rate limit of 28.8 kbit/s. The major outcome of this standardization work, completed at the end of 2005, is the TETRA Enhanced Data Service (TEDS) which became available from September 2006 as the ETSI Technical Standard TS 100 392-2 V 3.1.1 [ETS06].

In order to support wideband multimedia services whilst ensuring reliable link performance over harsh time-frequency selective fading mobile channels, the Release 2 of the TETRA standard [ETS06] features FBMCM-based signalling jointly with a combination of spectral-efficient multilevel modulation schemes (M -QAM and π/M -DMPSK) and powerful turbo channel encoding. Moreover, link adaptation techniques are also included so as to improve the system performance (e.g., the overall message throughput), based on choosing adaptively the modulation level, the coding rate and possibly the RF channel bandwidth according to the varying channel propagation conditions.

More specifically the up-to-date technological choices at the physical layer level with respect to the current TETRA 1 standard are the following [Nou06]:

- adoption of FBMCM-based signalling to achieve robust performance even in frequency-selective fading channels, for a total number of (2.7 kHz spaced) subcarriers ranging from 8 (25 kHz channel) to 48 (150 kHz channel);
- spectral-efficient multilevel modulation schemes, i.e., 4-QAM, 16-QAM and 64-QAM, to boost the system data throughput; these QAM schemes plus the $\pi/8$ -D8PSK scheme add to the $\pi/4$ -DQPSK modulation scheme used in the current TETRA 1 standard;
- powerful turbo-code for payload channel encoding with rates 1/2 and 2/3, plus rate 1 (uncoded case);
- separate channel encoding for short “header” blocks to exceed the payload performance and enable reliable slot decoding and network operations;
- flexibility of selecting the required data throughput from a wide range extending to beyond 500 kbit/s;
- link adaptation techniques to improve the system performance (e.g., overall message throughput), based on choosing adaptively the modulation level, the coding rate and possibly the RF channel bandwidth according to the varying channel propagation conditions.

3.4.2 Supported modes of operation

Table 3.4.1 presents the estimated throughput (in kbit/s) for all the modulation/coding combinations envisaged in TETRA 2 standard. Each modulation/coding scheme is associated to one or more channel widths. The first row refers to the current TETRA system.

Table 3.4.1: Estimated throughput (kbit/s) for TETRA 2 modulation/coding and channel setting.

<i>Modulation</i>	<i>25 kHz</i>	<i>50 kHz</i>	<i>100 kHz</i>	<i>150 kHz</i>
$\pi/4$ -DQPSK $r = 2/3$	15			
$\pi/8$ -D8PSK $r = 2/3$	24			
4-QAM $r = 1/2$	10	26	55	86
16-QAM $r = 1/2$	20	52	110	173
64-QAM $r = 1/2$	30	77	164	259
64-QAM $r = 2/3$	40	103	219	345
64-QAM $r = 1$	60	154	329	518

Note: Rates in kbit/s for a downlink transmitting data in 17 frames per multiframe

3.4.3 AMC implementation in TETRA 2

In TETRA 2 link adaptation may be employed in order to improve the link efficiency. In each time slot, the MAC (Medium Access Control) of the transmitting base station (BS) or mobile station (MS) selects the appropriate modulation level and coding rate (which in turn determine the available bit rate) based on the current link conditions and the “data category” parameter provided by the LLC (Logical Link Control) with the data. This parameter indicates the type of data (i.e., “background”, “telemetry”, “real-time”, “non-classified”) and the required reliability level. For “background” or “telemetry” class data, the reliability level may vary for transmissions of one data segment according to the number of times that segment has been transmitted, starting with low reliability (enabling higher throughput when successful) and increasing to higher reliability if the first transmission(s) of that segment are not successful. In a simple link adaptation algorithm, the choice of bit rate for each data category may be predefined. For example, for “background” or “telemetry” class data on a QAM channel, an MS or BS could send:

- one transmission of each segment at a low reliability level, using 64-QAM and coding rate 2/3;
- if retransmission of a segment is needed, one transmission of that segment at a medium reliability level, using 16-QAM and coding rate 1/2;

- then, if retransmission of the segment is still needed, further transmission(s) of that segment at a high reliability level, using 4-QAM and coding rate 1/2.

The graphs in Figure 3.4.1 present the performance achieved over fading channels for the M -QAM modulation schemes envisaged by TETRA 2, i.e., 4-, 16- and 64-QAM. The channels considered are the standard six-path GSM typical urban (TU) and hilly terrain (HT) models. In particular, the results refer to the following two scenarios: *i*) TU with mobile speed 50 km/h and carrier frequency $f_c = 400$ MHz (referred to as “TU50-400”); *ii*) HT with mobile speed 200 km/h and carrier frequency $f_c = 800$ MHz (referred to as “HT200-800”). The signal bandwidth is fixed at 50 kHz and the coding rate at $r = 1/2$. The charts present the frame error rate (FER) plotted vs. E_b/N_0 for 4-, 16- and 64-QAM over the TU50-400 and HT200-800 channels, together with those relevant to single-path Gaussian channel as reference. It is seen that at medium-to-high E_b/N_0 the receiver performs better on the HT200-800 channel rather than on TU50-400. This can be explained observing that the fast-varying context has inherently a higher degree of time diversity, and this provides a beneficial impact on the decoder, analogous to inner and outer interleaving.

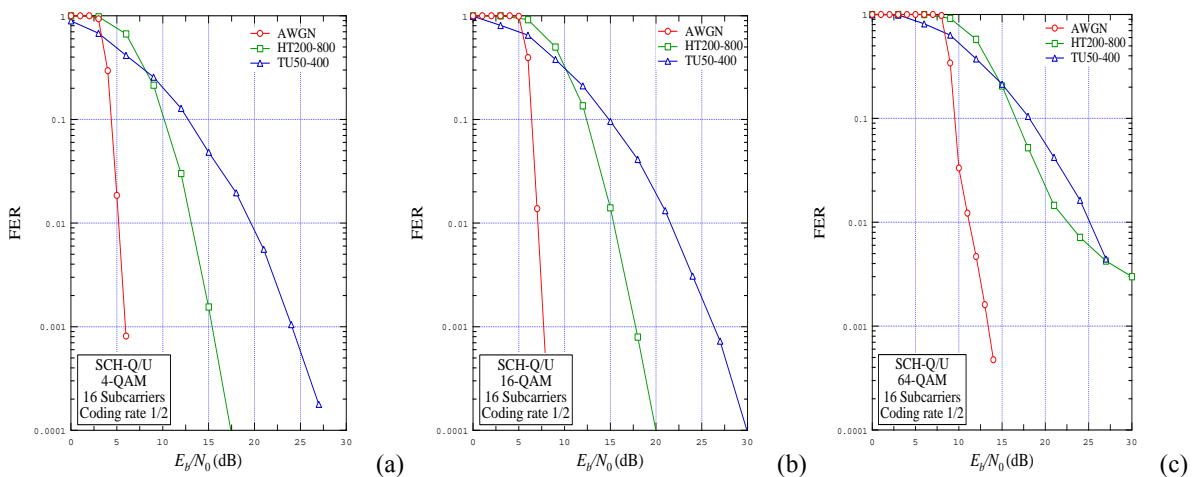


Figure 3.4.1: Curves of FER vs. E_b/N_0 over the TU50-400, HT200-800 and Gaussian channels with 4-QAM (a), 16-QAM (b) and 64-QAM (c).

The charts also highlights the benefits provided in terms of FER performance by the adaptation of the modulation scheme to the propagation conditions. To this aim, in each time slot, the MAC of the transmitter selects the appropriate modulation level and coding rate based on the current link conditions and the “data category” parameter provided by the LLC with the data. This parameter indicates the type of data category and the required reliability level. In a simple link adaptation algorithm, the choice of bit rate for each data category may be predefined. In a more complex link adaptation algorithm, the MAC’s choice of the appropriate bit rate for each of the data categories may vary adaptively according to the current channel conditions. The choice may be based on the link adaptation feedback messages, advanced link performance information provided locally by the LLC (derived from recent segment success/failure statistics) and/or measurements of the reverse channel. For example, use of downlink measurements, together with knowledge of the BS link imbalance and an MS correction factor, may enable the MS to estimate the uplink slot error rates. The link adaptation feedback messages may indicate the preferred modulation level and coding rate for a specified data class; alternatively they may provide the other party’s estimate of the received signal-to-noise ratio and an estimate of the channel model and speed.

3.5 DVB-S2

3.5.1 Overview

The DVB-S2 standard [ETSI EN 302-307] has adopted AMC, allowing different modulation formats and error protection levels to be used and modified on a frame-by-frame basis. By exploiting a return (feedback) channel to inform the transmitter of current receiving conditions, the transmission parameters can be optimized for each individual user according to its own channel propagation conditions. One of the key factors which can heavily affect the improvement of system capacity is the *granularity* of physical layer schemes supported by the system. In the current DVB-S2 standard the forward error correction rates have been designed in order to obtain a granularity of about 1 dB in C/N (Carrier to Noise ratio), covering a wide range of spectral efficiencies (Figure 3.5.1).

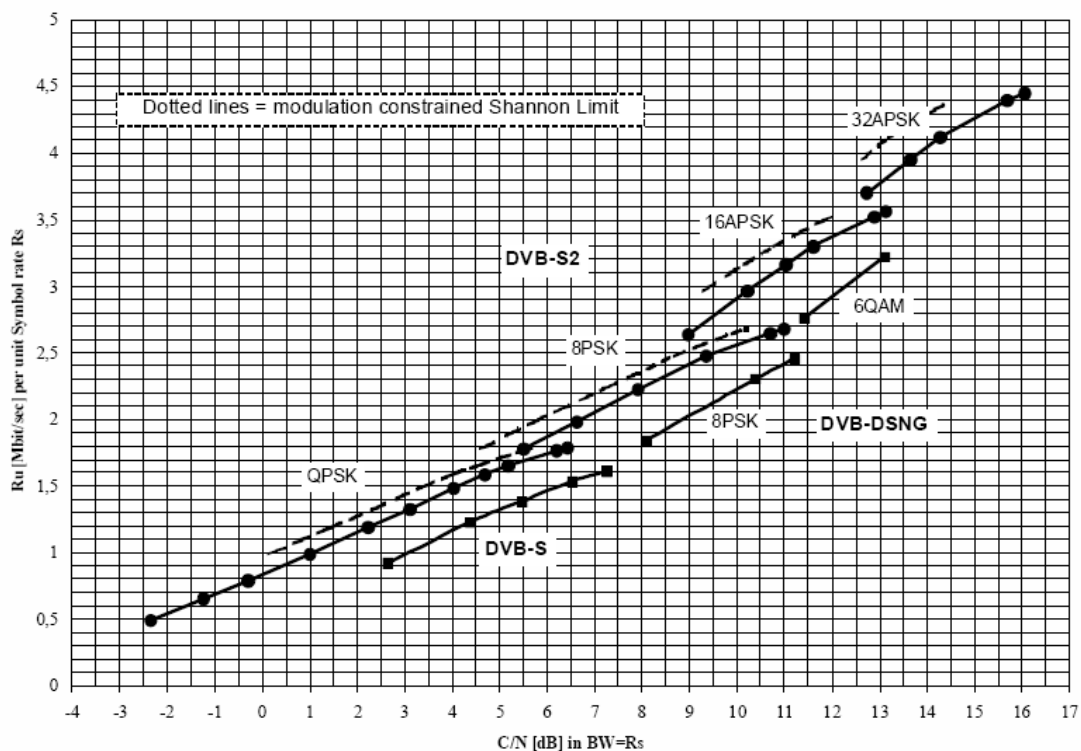


Figure 3.5.1: Required C/N versus spectrum efficiency in AWGN channel condition (extracted from [ETSI TR 102 376]).

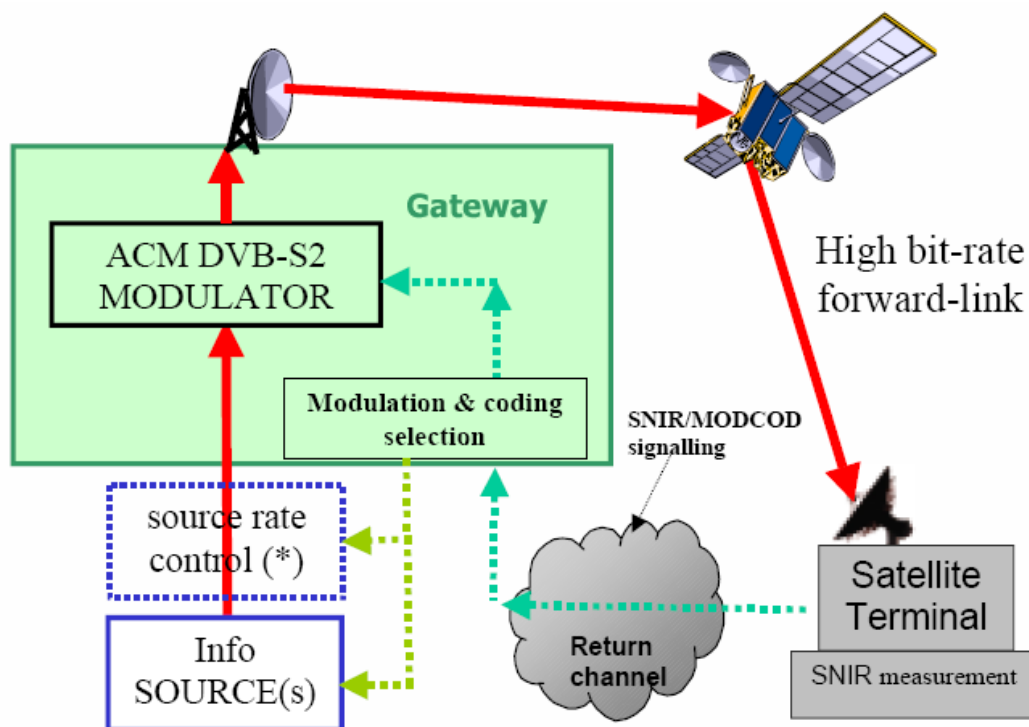
3.5.2 Supported modes of operation

In the table below the complete set of all possible coding rate and modulation pairs (called MODCOD) available in the DVB-S2 standard are presented.

Mode	MOD COD	Mode	MOD COD	Mode	MOD COD	Mode	MOD COD
QPSK 1/4	1 _D	QPSK 5/6	9 _D	8PSK 9/10	17 _D	32APSK 4/5	25 _D
QPSK 1/3	2 _D	QPSK 8/9	10 _D	16APSK 2/3	18 _D	32APSK 5/6	26 _D
QPSK 2/5	3 _D	QPSK 9/10	11 _D	16APSK 3/4	19 _D	32APSK 8/9	27 _D
QPSK 1/2	4 _D	8PSK 3/5	12 _D	16APSK 4/5	20 _D	32APSK 9/10	28 _D
QPSK 3/5	5 _D	8PSK 2/3	13 _D	16APSK 5/6	21 _D	Reserved	29 _D
QPSK 2/3	6 _D	8PSK 3/4	14 _D	16APSK 8/9	22 _D	Reserved	30 _D
QPSK 3/4	7 _D	8PSK 5/6	15 _D	16APSK 9/10	23 _D	Reserved	31 _D
QPSK 4/5	8 _D	8PSK 8/9	16 _D	32APSK 3/4	24 _D	DUMMY PLFRAME	0 _D

3.5.3 AMC implementation in DVB-S2

In Figure 3.5.2, a conceptual block diagram for the AMC implementation in the DVB-S2 standard is depicted. The DVB-S2 ACM modulator operates at a constant symbol rate because the available transponder bandwidth is assumed constant. During atmospheric fading events, the service continuity and QoS are achieved by reducing the number of transmitted bits while at the same time increasing the FEC redundancy and, thus, the overall robustness. This way, the channel symbol rate is kept constant.



(*) Source rate control may be directly applied to source(s) or locally at the GTW input or via network traffic control

Figure 3.5.2: Conceptual block diagram for a DVB-S2 link with AMC (extracted from [ETSI TR 102 376]).

In general, the physical layer adaptivity is obtained as follows:

- via a return (feedback) channel, individual satellite terminals provide to the gateway information on their channel status through the reported SNIR. Since periodic reports could increase the return link signaling, a preferable approach is for the terminal to send a message whenever channel variations imply a change in the modulation and coding rate pair.

- in order to avoid information overflow during fades, a source bit rate control mechanism has to be implemented, adapting the offered traffic to the available channel capacity.

A crucial issue in all AMC systems is the physical layer adaptation loop delay, since it is strictly related to the system capability of tracking channel propagation variations. The adaptation loop can be defined as the set of operations which occur between the start of channel estimation at the user terminal side and the end of the reception of the information encoded/modulated signal. In general, the longer the loop delay the higher the impact physical layer adaptation has on the overall user QoS, because of the propagation channel dynamics. The satellite loop delay is much more challenging in comparison to terrestrial networks due to the fixed 250 ms propagation delay from the gateway to the user terminal and vice versa (for geosynchronous satellites). Of course, different strategies for information processing and buffering at the transmitter side lead, in general, to different QoS performance for the same channel variation rate.

3.6 Winner/Winner II

3.6.1 Overview

The WINNER (Wireless World Initiative New Radio) radio interface for next generation (beyond 3G) wireless systems has been designed as a packet-oriented, user-centric, ubiquitous, *always-best* concept [WIN2D61314]. An always-best solution, which provides competitive performance in a variety of scenarios is a challenging design goal. Therefore, one of the key new features of the proposed WINNER concept, which helps in fulfilling this goal, is a flexible protocol architecture with incorporated mechanisms for both long-term and short-term adaptation.

There are two methods of link adaptation (LA) in the WINNER concept in the form of AMC [Ste07]:

- **Frequency-adaptive transmission**, where user data flows are given exclusive access to *chunks*¹ (or *chunk layers*) and individual link adaptation is performed within the chunks (or chunk layers). The adaptation utilizes the frequency selectivity of the channel and uses a very fast feedback loop, working on the time-scale of the frame to follow the short-term fading.
- **Non-frequency-adaptive transmission**, which averages over the frequency variations of the frequency-selective channels. A code block is interleaved and mapped onto a wide frequency range. The whole code block utilizes the same modulation and coding scheme. Modulation and coding is adapted to the shadow fading and path loss, but not to the fine frequency-selective fading.

The above two methods are based on different concepts. The former utilizes the frequency-selective channel variations whereas the latter averages over them by diversity techniques. The frequency-adaptive transmission should typically be combined with MIMO schemes that preserve the channel variability (e.g. spatial multiplexing), while the non-frequency-adaptive transmission is more suitable for schemes with multi-antenna diversity techniques (such as *space-time coding*), which reduce the variability of the perceived channel.

Figure 3.6.1 presents the WINNER scheduling architecture with the two parallel methods of adaptive transmission. An input data flow in the form of protocol data units (PDUs) coming from the radio link control (RLC) layer is first segmented into packets and FEC encoded. The encoding is done with the use of efficient block-circulant low-density parity-check (BLDPC) codes and the encoded packets are next interleaved and optionally punctured. Depending on the current mode of operation, which is selected according to a set of input parameters (like channel quality information (CQI), mobile terminal (MT) velocity, requested resources, MT functionality, etc.), the encoded packets are either

¹ A “*Chunk*” is the smallest time-frequency unit of the WINNER frequency-adaptive transmission scheme; it consists of n_f adjacent subcarriers and n_t following OFDM symbols. In case of multi-antenna transmission, there are multiple independent *chunk layers*, which can be independently loaded with some data (cf. Figure 1.3 in [WIND210])

mapped onto dispersed blocks of time-frequency grid (non-frequency-adaptive transmission) or mapped onto optimum chunks according to the output of the adaptation in the frequency.

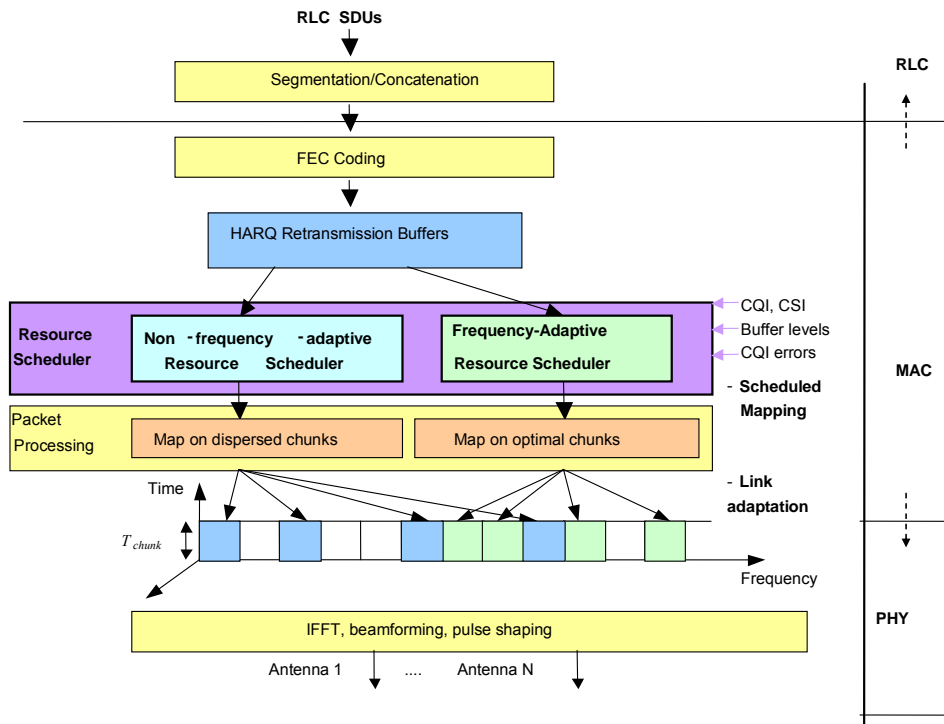


Figure 3.6.1: WINNER scheduling architecture for downlinks [Ste07]

3.6.2 Frequency-adaptive transmission

The OFDM system scheme for the frequency-adaptive transmission method is depicted in Figure 3.6.2. In the transmitter, an input data flow from one user is first grouped into N_{cw} packets of K bits, which are encoded by the BLDPCC code encoder and punctured according to an adapted code rate R . Such a group of encoded packets, each of length $N = K/R$, is then bit-interleaved and mapped onto a set of N_{ch} chunks, which have been already assigned to this particular user flow by the resource scheduler. Next, the loaded chunks are modulated according to the MI-ACM (*mutual information based adaptive coding and modulation*) algorithm output. The modulation order, i.e. the number of bits per QAM symbol, determined by the algorithm is denoted as $b = [b_1, b_2, \dots, b_{N_{ch}}]$, where $b \in B = \{1, 2, 4, 6, 8\}$. The number of coded bits at the output of the interleaver is thus

$$N_{cb} = N_q \sum_{n=1}^{N_{ch}} b_n$$

where $N_q = n_t n_f$ stands for the number of QAM symbols per chunk. Since N_{cb} is generally not an integer multiple of the codeword length, the number of packets in one frame is calculated as

$$N_{cw} = \left\lfloor \frac{N_{cb} R}{K} \right\rfloor$$

The missing $N_{pad} = N_{cb} - N_{cw} N$ bits are inserted before interleaving as a cyclic repetition of the first N_{pad} bits of the codeword block.

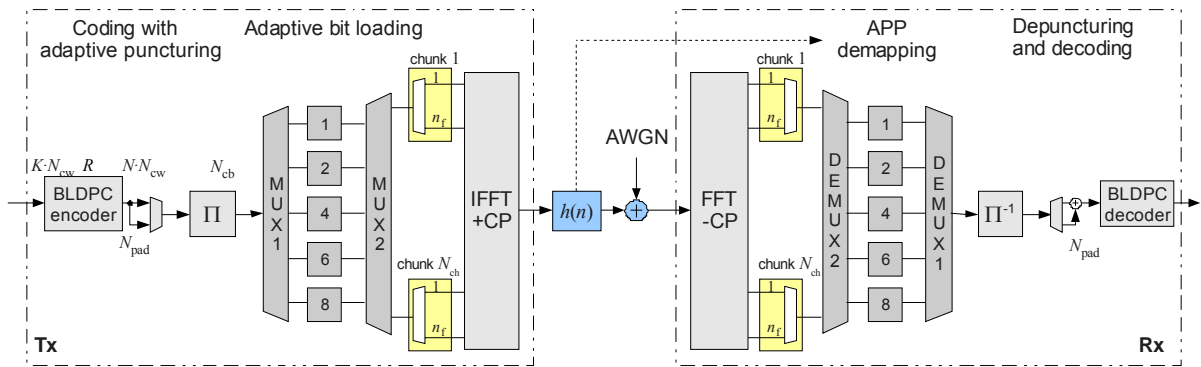


Figure 3.6.2: Simplified system model for the WINNER frequency-adaptive transmission [Pfl07]

MI-ACM algorithm

The core of the frequency-adaptive transmission scheme in WINNER is the *mutual information based adaptive coding and modulation* (MI-ACM) algorithm, which was proposed by S. Stiglmayr et al. in [Sti07] [WIN2D223]. The key idea in this algorithm is the mutual-information-based averaging. This step allows the accurate consideration of channel coding without being restricted to a specific coding scheme. Hence, in spite of the fact that *rate-compatible punctured* (RCP) BLDPC codes have been chosen for the final WINNER reference design, the MI-ACM algorithm is equally applicable to any rate-compatible codes. The application of MI-ACM algorithm in WINNER has been compared with a modified Hughes-Hartogs algorithm [Hug87] [Dlu04], which is the optimum bit and power loading algorithm in terms of rate maximization or power minimization. During this investigation [Pfl07] [WIN2D223], the following conclusions have been drawn:

- the performance in terms of throughput of the MI-ACM algorithm is almost the same as the reference optimum Hughes-Hartogs algorithm, which means that the gain from power-loading is insignificant
- the signalling overhead required for proper adaptation is quite low, since only 2 bits per each chunk are required to signal the modulation size and additional 3 bits per one frame (consisting of one or more FEC packets) are used for signalling the code rate
- the use of additional intermediate non-square QAM constellations (8-, 32- and 128-QAM) in adaptation has no any effect on the throughput of the link
- the computational complexity of the MI-ACM algorithm is very low.

3.6.3 Non-frequency-adaptive transmission

The WINNER non-frequency-adaptive transmission scheme is used when the frequency-adaptive transmission is impractical or impossible. This scheme averages over the frequency domain channel properties. The same modulation and code rate scheme is used for all resources assigned to a particular codeword (packet). That means that from a coding and decoding perspective, the non-frequency-adaptive transmission is a simplified case of the more general frequency-adaptive transmission [Sve07] [WIN2D461].

The link adaptation used for non-frequency-adaptive transmission is based on the average SINR of the resource blocks to be used. Such measurements can be done on a time-scale corresponding to that of the shadow fading. Faster updates of SINR estimates are potentially advantageous in environments with fast-varying interference. This would result in better adjustment of the modulation and coding rate to the instantaneous interference level, thus reducing the need for retransmissions. It would be of particular value in the transmission of urgent control information, for which Hybrid ARQ cannot be used due to timing constraints.

The mapping of code blocks (packets) onto transmission resources for the non-frequency-adaptive transmission has been proposed in [WIN2D461]. In order to gain from the diversity in frequency domain, small packets need to be mapped onto smaller resource units than the chunks defined for frequency-adaptive transmission. Therefore, the data coming from different users is regularly spaced in the frequency and mapped onto small time-frequency blocks (sub-blocks of a chunk). This scheme

is denoted as B-EFDMA (*block-equidistant frequency division multiple access*) in downlinks and B-IFDMA (*block-interleaved frequency division multiple access*) in uplinks and is motivated by the following reasons [WIN2D461]:

- orthogonal mapping of flows from different users simplifies receivers (no code multiplexing)
- small resource blocks decrease the padding loss when mapping small packets
- regular spacing of resource blocks in the frequency domain reduces the signalling (addressing) overhead and also permits the use of DFT-precoded transmission in uplinks, which lowers the signal envelope variations
- a short block duration in time, as compared to the chunk duration, allows for micro-sleep modes of terminals, when no data is transmitted or received; this should result in less power consumption in overall
- a short block duration in time also improves the robustness w.r.t. channel time variations within blocks; this is especially interesting for wide area scenarios with the highest vehicular velocities (e.g. 250 km/h in a fast train).

3.7 LTE – Open Air Interfaces

UMTS-LTE is an evolution of the UMTS-HSPA physical layer in the sense that multicarrier-WCDMA is replaced by MIMO-OFDMA (Multiple-Input Multiple Output Orthogonal Frequency Division Multiple-Access) on the downlink and MIMO-SC-FDMA (MIMO single-carrier FDMA) on the uplink. The standard is still in the process of being specified, so the information contained here can only be considered as a high-level summary of the current release (see [3GPP TS36.211, 3GPP TS36.212]). The use of OFDM and dynamic FDMA allows for very fine spectral resource allocation which is beneficial for MAC scheduling and PHY link adaptation mechanisms.

3.7.1 Downlink LTE-PHY overview

LTE retains the basic features regarding link adaptation (type I and type II HARQ) and adds additional features for controlling the tradeoff between spatial multiplexing and diversity in MIMO systems. The choice of whether spatial multiplexing (e.g. by multi-stream space-time codes) or diversity transmission (e.g. by single-stream space-time codes) is used can be controlled using rapid feedback mechanisms. A schematic overview of the MIMO-OFDM feedback-based strategy (downlink) is shown in Figure 3.7.1.

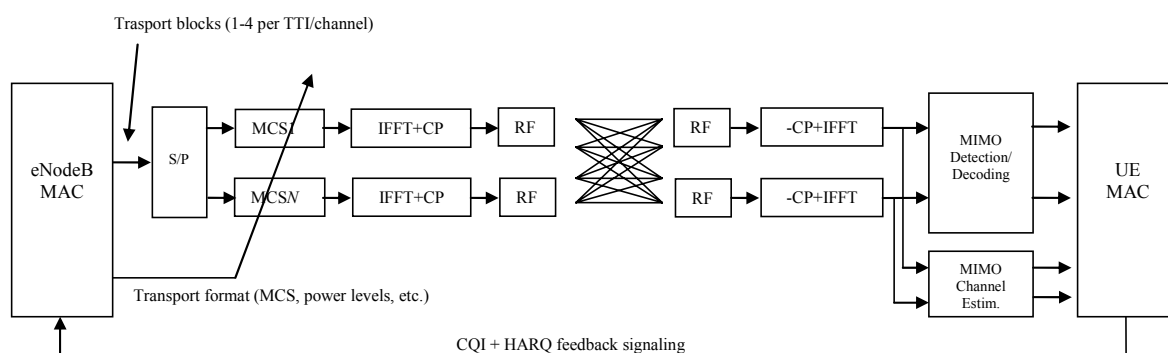


Figure 3.7.1: High-level View of LTE AMC

In Figure 3.7.1 we see that transport blocks (codewords) arrive at the PHY interface along with format information for MCS and power selection. Up to 4 transport blocks per transport channel can be sent to the PHY, corresponding to the maximum number of spatial streams provided. Very fine granularity can be accommodated in terms of channel code block size and code rate using rate-matching (puncturing or repetition) of two basic binary codes. Each transport block is passed to the processing chain in Figure 3.7.2.

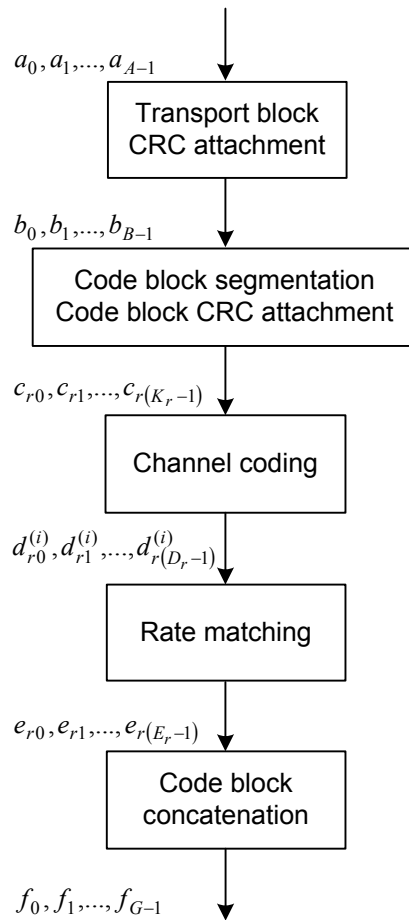


Figure 3.7.2: Transport channel processing for data channels

Each transport block has a 24-bit CRC attached as a error-detection mechanism for the HARQ process controller. The extended blocks are then segmented into code blocks in the event that they are longer than the maximum code block size of the underlying channel code, and each segment receives an addition CRC. For data channels a rate-matched rate 1/3 parallel concatenated turbo code is used and for the downlink broadcast control channel a 1/3 convolutional code is used. Rate matching is based on random repetition or puncturing of the coded data stream. Random puncturing is achieved through pseudo-random interleaving and predefined repetition/puncturing rules. In each round of a HARQ process, the code segments are punctured differently so that an incremental redundancy effect is guaranteed (type II HARQ). The receiver must therefore perform advanced combining across HARQ rounds. It is worth noting that the block size of the codewords input to the PHY layer can change dynamically from TTI to TTI (1ms intervals) in order to allow for fine-grain scheduling of user traffic.

The front-end of an LTE downlink transmitter (modulation and precoding) is shown in 3.7.3.

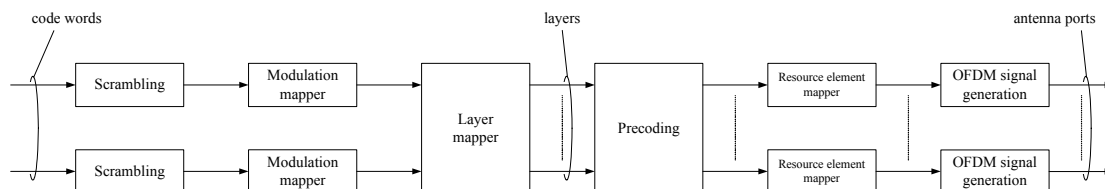


Figure 3.7.3: Front-end LTE Transmitter

Coded bits for each downlink spatial stream are first scrambled and modulated using either QPSK, 16-QAM or 64-QAM modulation. They are then mapped to different layers and input to a precoding matrix which creates the true spatial streams. Each stream is then mapped to the appropriate spectral

resource blocks determined by the MAC-layer scheduler and passed to the OFDM signal generator and the corresponding antenna port of the transmitter.

3.7.2 *Downlink Link Adaptation*

Link adaptation can be very fine-grain in UMTS-LTE due to the advanced measurement reports and granularity of frequency/space measurement information provided by the PHY and high-rate feedback signaling channels. The resource manager in the basestation can manage power and time/frequency/space resources and MCS at the granularity of the TTI (1ms) in order to maintain the BLER seen by the MAC layer and delay seen by the RLC layer (i.e. after HARQ). In other words, targets for link adaptation can be not only rate but delay as well, which can be important for future application scenarios. Dynamic choice of MIMO transmission scheme (diversity/multiplexing) should also be possible.

3.7.3 *Downlink LTE-HARQ overview*

Two levels of ARQ are provided by UMTS-LTE; an N -channel stop-and-wait protocol is used by eNodeB to control PHY transport blocks with an incremental redundancy mechanism. Residual errors after a fixed number of retransmission rounds are potentially corrected using an outer selective-repeat ARQ protocol in the RLC layer, when using an acknowledged-mode (AM) radio bearer. UMTS-LTE uses an asynchronous HARQ protocol, as opposed to that of HSPA. This offers the possibility of dynamically scheduling retransmissions based on air interface conditions. Moreover, the protocol can be adaptive in the sense that MCS parameters and can be dynamically modified during a retransmission round for a particular codeword. The latter include the modulation size, spectral allocation and transmission duration. The HARQ process manager in eNodeB is intimately integrated with the downlink packet scheduler which controls the MCS and spectral resource allocation.

3.7.4 *OpenAirInterface AMC*

The physical (PHY) layer of the OpenAirInterface platform targets WiMax and UMTS-LTE-like networks and thus uses multiple-input multiple-output orthogonal frequency division multiple access (MIMO-OFDMA) as a modulation and multiple access technique. One of the main targets of this research initiative is to explore innovative aspects in PHY design (in particular with respect to AMC) on an experimental set of specifications which broadly resemble those of UMTS-LTE and WiMAX in spirit but are flexible enough to permit experimentation in algorithm development. It is hoped that the experimental techniques could eventually find their way into industrial air interface standards.

OpenAirInterface's MIMO-OFDMA system provides the means for transmitting several multiple-bitrate streams (multiplexed over subcarriers and antennas) in parallel. Moreover, PHY signaling strategies are included to provide the means for exploiting channel state feedback at the transmitters in order to allow for advanced PHY allocation of OFDMA resources via the MAC as in UMTS-LTE.

OpenAirMesh makes use of punctured binary codes (64-state rate 1/2 convolutional or 8-state rate 1/3 3GPP/LTE Turbo code). Puncturing uses random puncturing as in LTE, although based on pseudo-random number generators, in order to fine tune the coding rate to adapt to configurable transport block sizes delivered to PHY by the MAC. The overall coding sub-system is shown in Figure 3. New transport blocks arriving from the MAC layer (based on multi-user scheduling) are coded using a CRC extension and the chosen binary code. These are then fed to the active transport block buffer along with those that are to be retransmitted. Each transmitted block is punctured and then passed to a bit-interleaver and modulation mapper (BICM). OpenAirMesh supports QPSK, 16-QAM and 64-QAM modulation.

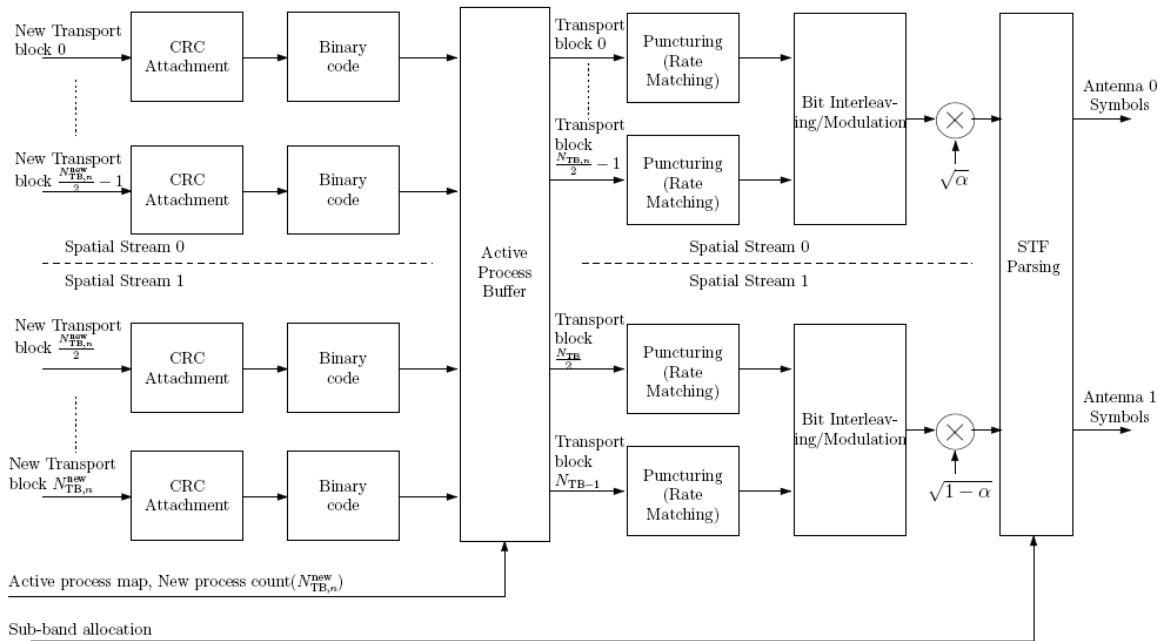


Figure 3.7.4: OpenAirInterface Coded Modulation

The main difference in OpenAirInterface's AMC system with respect to UMTS-LTE is that the transport block size remains fixed, but that a variable number of transport blocks can be sent in each TTI. This allows for more fine grain control of the HARQ protocol, since each transport block contains its own CRC. It is also envisaged to consider very small blocks without appending a CRC through the use of an error-detection mechanism based on bounded-distance decoding [For69, ElG06]. This will allow us to investigate the potential benefit that increasing the feedback rate on the uplink provides to downlink rates by very finely controlling the retransmission of small blocks of data. In addition, the possibility of using very small block sizes provides more flexibility to the resource scheduler in the eNodeB.

Another commonality with UMTS-LTE is that the transmitted transport blocks can be split into two spatial streams in the case of point-to-point MIMO transmission. Each stream receives an adjustable amplitude and then each is passed to a different (orthogonal) space-time parser which guarantees that both streams use different antennas in the same time/frequency dimension. The space-time mechanism is simpler in OpenAirInterface than that of UMTS-LTE but allows for low-complexity successive detection at the receiver and maximizes diversity against fading. This is a form of superposition coding since the two streams are combined additively in the air through the use of multiple transmit antennas. A second design objective for this coding strategy, in addition to low-complexity point-to-point MIMO operation, is that the same transmitter and receiver structure can be used in a distributed MIMO scenario. Here one spatial stream is used at each source and the second stream originates in another part of the network, either in the same cluster or an adjacent cluster. Co-operation is needed in order to guarantee different STF parsing for the two streams so that they can be decoupled at the SIC receiver. A particular user can decode both streams or simply select the one it requires. This provides a means for either unequal error-protection or single-frequency reuse across cells with interference cancellation at the receiver [Gha08a, Gha08b].

4. LINK LEVEL PERFORMANCE MODELING

4.1 Introduction

The *AMC design problem* can be viewed as a constrained optimization task with objective functions and constraints properly defined so as to accommodate specific system scenarios. A simple-to-use (thus, of reduced complexity) yet sufficiently accurate link-performance functional model that accounts for the combination of the transmitted-signal mode choice plus channel and interference conditions is crucial for efficient AMC algorithmic design, since it serves as a performance-prediction tool that guides the AMC optimization sub-routine to reach meaningful decisions. Such a model must be detailed enough to include channel modeling issues such as, for instance, the effect of multiple antennas at the transmitter and/or the receiver in combination with the applied MIMO technique (beamforming or other spatial multiplexing). We note that similar compact-description models are also of great interest in the context of Evaluation Methodologies (EVM) which are currently being developed for various systems in the respective standardization bodies. However, these have a different goal in mind, namely to use this type of *physical-layer abstraction* in order to determine the performance of a given link and thus avoid the need for extensive simulation. This “simulation-shortcut” in turn accelerates the corresponding system-level simulations where a large number of physical-layer-related links need to be taken into account. The abstraction should be accurate, computationally simple, relatively independent of channel models, and extensible to interference models and multi-antenna processing. It follows that, although our reasons for a PHY-abstraction methodology are primarily for AMC algorithm design and performance estimation, the related PHY abstraction methodologies developed for such specific EVM’s are quite relevant to our work. Hence, in this chapter, a survey of link-level models used for PHY-layer abstraction in specific standards/systems will be presented.

It is worth elaborating a bit further, at this point, on the role of performance evaluation for a given link. In older and simpler systems where multi-modality was not an option, the role of performance evaluation (analytically or by simulation) was to simply check whether a given signal design met the pre-specified performance requirements. If there were degrees of freedom (typically in parametric form) either for transmitter-based signal design or for receiver-based algorithmic choice, then the role of performance evaluation (now plotted versus ranges of parametric values) was to pick the right set of parameter values so as to optimize a performance metric (for instance, to minimize the BER). In that sense, performance evaluation started becoming an integral part of the *system design process* itself, and the motivation thus arose to have simple analytic forms for these performance results which would make them amenable to easy parametric optimization. Once the design aspect advances to become multi-modal and multi-parametric at both sides of the transmission link (which AMC definitely is), the task of link-performance evaluation becomes not only germane to the design procedure itself, but the effective and efficient representation of this parameterized performance in ways that are compact (parsimonious) yet accurate comprises a main challenge of the optimization task. Otherwise, overly complex or inaccurate representations may lead the whole design process to naught. It is in this spirit that we put great emphasis on these abstraction methodologies in this chapter. Finally, we note that there is one final “loop” of performance assessment that is needed in AMC and which is not really present (or needed) in the previously-mentioned design cases: this is the final performance assessment once the mode has been chosen and applied to a given link. Because the aforementioned performance-modeling description is, out of necessity, of reduced dimensionality and fidelity, it is important to check whether the system truly performs as expected in actual scenarios and true prevailing conditions, a final task usually taken up by simulation.

4.2 Link level modeling for WiMAX

4.2.1 Introduction

For system-level simulations in the past, the average performance of a system was quantified by using the topology and the channel macro-characteristics in order to compute a geometric (or average) SINR

distribution across the cell. Each subscriber's average SINR was then mapped to the highest modulation and coding scheme available, which could be supported based on link-level SINR tables that capture the fast-fading statistics. The link-level SINR-versus-PER look-up tables served as the PHY abstraction for predicting average link-layer performance. Examples of this static methodology may be found in [RUN03a], [RUN03b].

Current system designs (HSDPA, 3GPP-LTE, WiMAX) are based on exploiting instantaneous channel conditions for performance enhancement. Channel-dependent scheduling and adaptive coding and modulation are examples of channel-adaptive schemes employed to improve system performance. This in turn creates the need for current system level evaluation methodologies to be based on the modelling of the *dynamic* system behavior, in other words, including fast fading models within the system-level simulation. As a consequence, the system-level simulation must support a PHY abstraction capability in order to accurately predict the instantaneous link-layer performance of the PHY link layer in a computationally simple way. This abstraction should be accurate, computationally simple, relatively independent of specific channel models, and extendable to interference models and miscellaneous multi-antenna processing schemes. The content of this section is based on the ongoing work in IEEE 802.16m Task Group [IEEE802.16m-08].

4.2.2 PHY Abstraction Methodology

PHY abstraction methodology for predicting instantaneous link performance for OFDM systems has been an active area of research and has received considerable attention in the literature ([SON04], [STM04], [ATH04], [3GPP25892], [WG501], [ERI03], [NOR03], [NOR04], [BRU05], [WAN06], [DAG03]). In coded OFDM systems the encoded packet may be transmitted over a frequency selective channel, resulting in unequal channel gains for the sub-carriers, and thus non-uniform post-processing SINR values of the data streams just prior to decoding. Additionally, the channel gains of the sub-carriers can be time selective due to the fading process and possible delays involved in HARQ re-transmissions. The result on a transmission of a large encoder packet is thus that encoded symbols possess unequal SINR ratios at the input of the decoder due to the selective channel response over the encoder packet transmission. In this case, the role of a PHY abstraction method is to predict the coded Block Error Rate (BLER) for a given received channel realization across the OFDM sub-carriers used to transmit the coded FEC block. In order to predict this coded performance, the post-processing SINR values at the input to the FEC decoder are considered as input to the PHY abstraction mapping. As the link level, curves are generated assuming a frequency flat channel response at given SINR. An *effective* SINR, $SINR_{eff}$, is then required to accurately map the system-level SINR onto the link-level curves for determining the resulting BLER. This mapping is quite important and is termed *effective SINR mapping (ESM)*. The ESM PHY abstraction is thus tantamount to compressing the vector of received SINR values to a single effective SINR value, which can then be further mapped to a BLER number; the procedure is shown in Figure 4.2.1.

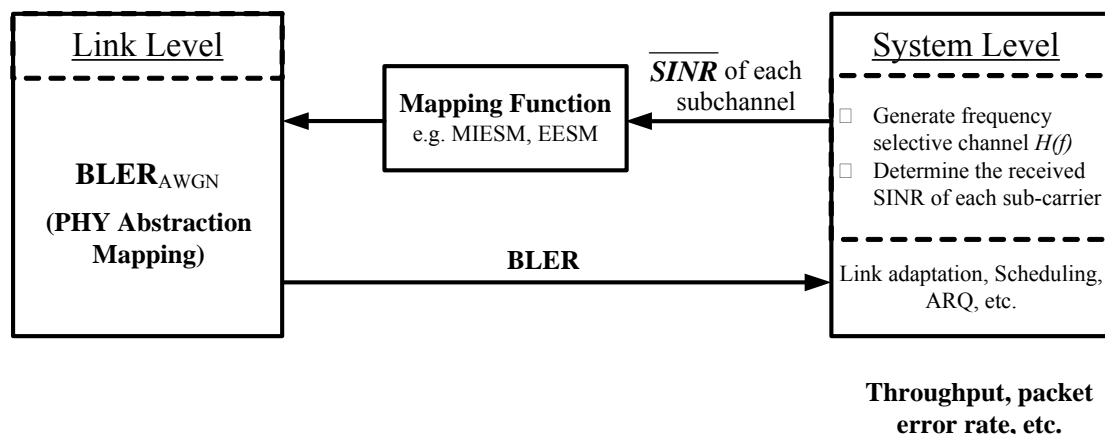


Figure 4.2.1: PHY link-to-system mapping procedure [IEEE802.16m-08]

Several ESM approaches for predicting the instantaneous link performance have been proposed in the literature, including: mean instantaneous capacity [Son04], [STM04], [ATH04], exponential-effective Reference DR.3.1

SINR Mapping (EESM) [3GPP25892], [ERI03], [NOR03], [NOR04] and Mutual Information Effective SINR Mapping (MIESM) [BRU05], [WAN06]. Within the class of MIESM there are two variants: one is based on the mutual information per received symbol normalized to yield the bit mutual information and the other directly computes the bit mutual information. Each of these PHY abstractions uses a different function to map the vector of SINR values to a single number. Given the instantaneous ESM SINR, the BLER is calculated using a suitable mapping function

In general, any ESM PHY abstraction method can be described via the following equation

$$SINR_{eff} = \Phi^{-1} \left\{ \frac{1}{N} \sum_{n=1}^N \Phi(SINR_n) \right\}$$

where $SINR_{eff}$ is the effective SINR, $SINR_n$ is the SINR in the n^{th} sub-carrier, N is the number of symbols in a coded block, or the number of sub-carriers used in an OFDM system and $\Phi(\bullet)$ is the invertible function that defines the specific ESM. In the case of the mutual information-based ESM, the function $\Phi(\bullet)$ is derived from the constrained capacity; while in the case of EESM, the function $\Phi(\bullet)$ is derived from the Chernoff bound on the probability of error. Details on those methods can be found in [IEEE802.16m-08].

Another important abstraction step is the Per-tone SINR Computation. All PHY abstraction metrics are computed as a function of post-processing per-tone SINR values across the coded block at the input to the decoder. The post-processing per-tone SINR is therefore dependent on the transmitter/receiver MIMO-STC structure used to modulate/demodulate the symbols.

4.2.3 Abstraction-related issues

In several cases the packet error rate (PER) is a more suitable metric than BLER. A packet comprises several FEC blocks. PER is the probability that an error occurs in at least one of FEC blocks comprising the packet {ap: PER used previously}. The PHY abstraction predicts the link performance, in terms of BLER, for a coded FEC block. Thus the need arises to derive the PER, given the predicted BLER. If a packet is comprised of J blocks and the predicted BLERs are given by $BLER_1, BLER_2, \dots, BLER_j$, then assuming that the block errors events are independent, the PER is given by:

$$PER = 1 - \prod_{j=1}^J (1 - BLER_j)$$

Furthermore, WiMAX includes H-ARQ coding and retransmission schemes. PHY abstraction approaches for H-ARQ depend on the H-ARQ method. The approaches are similar for all bit-based mutual information-based abstraction techniques.

The following abstraction is proposed as the baseline choice going forward:

- For Chase combining (CC): The SINR values of the corresponding sub-carriers are summed across retransmissions, and these combined SINR values will be fed into the PHY abstraction.
- For Incremental redundancy (IR): The transmission and subsequent retransmissions are regarded as a single codeword, and all the SINR values are fed into the PHY abstraction. In practice, some partial repetition occurs when part of the coded information is repeated in subsequent retransmissions. For methods combining CC and IR this approach is preferred but should be justified by link-level simulations.

For Repetition Coding the SINR values of the sub-carriers are summed across the repetition number, and these combined SINR values will be fed into the PHY abstraction.

4.3 Link level modeling in WINNER

As it is indicated in the introduction to this chapter, the modelling of link level performance of wireless systems is quite important from the computational complexity point of view of system level simulation tools. It is thus unsurprising that PHY abstraction models were evaluated and included in the final system concept of the WINNER (Wireless World Initiative New Radio) and WINNER II IST projects [BRU05] [WIN2D61310]. The concept of link-level performance modelling is known in WINNER under the name of link to system (L2S) interface. A generic modelling approach, which takes instantaneous channel and interference characteristics into account, is depicted in Figure 4.3.1. In this model, the performance of radio links is evaluated in terms of PER as a function of some input quality measures, e.g. SINR [Lam03] [3GPP25892] [3GPP2C30], raw bit error probabilities [Häm97] [Olo97], or SISO/MIMO channel capacities for Gaussian signalling [FITD331]. Many proposals found in the literature are compatible with this approach.

In the first step a set of quality measures is extracted for all resource elements² covered by the FEC packet of interest (see Figure 4.3.1). Because the number of quality measures can be quite large, usually some compression has to be performed to obtain typically one or two characteristic scalars to be later mapped to the output PER. In WINNER the so called effective SINR (SINReff) [3GPP2C30] is calculated for a larger set of input SINR measures, which realises the compression step from Figure 4.3.1. Next, the effective SINR is mapped with one-dimensional function to the output PER. This results in a link level performance model of low complexity and training requirements. The mapping of SINReff to PER is performed using a look-up table, with error rate curves in AWGN corresponding to particular modulation and coding schemes. This mapping is generally specific to code rate, code type, codeword length and modulation type [WIN2D61310].

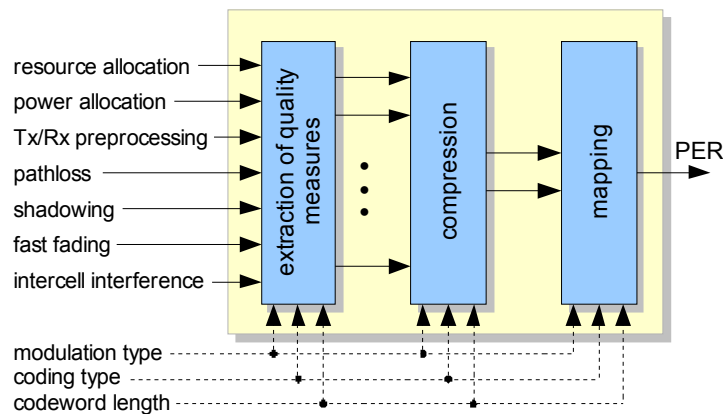


Figure 4.3.1: Generic link performance model [Bru05]

According to the evaluations presented in [Bru05], the MIESM seems to be the most accurate modelling approach in case of resource specific adaptive modulation and different spatial processing techniques. The MIESM link to system interface, which averages a set of input SINR values in the mutual information domain, can be described with the following formula [WIN2D61310]:

$$\text{SINR}_{eff} [\text{dB}] = 10 \log_{10} \left[\beta \cdot I_M^{-1} \left(\frac{1}{P} \sum_{p=1}^P I_M \left(\frac{\text{SINR}_p}{\beta} \right) \right) \right]$$

where:

- β is an optimization parameter to be derived from link level simulations
- P is the number of resource elements covered by a codeword of interests, for which SINR_p values are provided (calculated)

² In the WINNER frequency-adaptive transmission scheme, *chunks* are the smallest time-frequency resource elements

- M is an identifier of the modulation alphabet applied to all element of the packet (number of constellation points)
- I_M is the mutual information associated with the modulation alphabet M as a function of SINR, defined as follows³:

$$I_M(x) = \log_2(M) - E_Y \left\{ \frac{1}{M} \sum_{i=1}^{\log_2(M)} \sum_{b=0}^1 \sum_{z \in X_b^i} \log \frac{\sum_{\hat{x} \in X} \exp(-|Y - \sqrt{x}(\hat{x} - z)|^2)}{\sum_{\tilde{x} \in X_b^i} \exp(-|Y - \sqrt{x}(\tilde{x} - z)|^2)} \right\}$$

where: X is the set of M symbols, X_b^i is the set of symbols for which bit i equals b . Further, Y is zero mean unit variance complex Gaussian variable.

4.4 OpenAirInterface Simulation/Emulation Methodology

Eurecom is developing a system simulator based on the www.openairinterface.org development platform which is an open hardware and software initiative for innovation in emerging air interface technology. Here we provide an overview of the OpenAirInterface simulation/emulation methodology, which provides both a real-time and offline distributed validation environment comprising:

- A PHY emulation layer using IP Multicast over Ethernet or direct-memory transfer
- PHY behavioural abstraction models
- OpenAirInteface Layer two real-time protocol stack (openair2) potentially virtualized into N instances in the same physical machine.

The development can be used for the purpose of both protocol implementation validation which enables developers of layer 2 and layer 3 and applications to test their implementation in a real-time setting without the need for RF equipment. Moreover, it provides an environment for completely repeatable experiments using a scalable simulation of a real protocol stack using a PHY abstraction layer. For the purpose of system performance evaluation it allows for layer 2 and 3 algorithm assessment using accurate PHY abstraction model. Accurate PHY abstraction is required for both modes of operation.

In terms of runtime environments, the system simulator can be executed in two different forms on one or several standard Linux PCs. A typical configuration is shown in Figure 4.4.1. The system simulator comprises one kernel module, a Linux networking device driver called `nas_driver`, and a user-space executable which contains N instances of the OpenAirInterface layer 2 protocol stack (MAC,RLC,PDCCP,RRC) and a PHY_abstraction unit (green in the figure) each corresponding to a particular node in the network. These instances can be distributed on a network of PCs for a large-scale simulation (networks comprised of hundreds of nodes can be envisaged). The simulator executes in non-real time but with a true networking device so that higher-layer protocols (routing, mobility management, etc.) can be integrated into the simulation if needed. The PHY_abstraction uses memory transfer for SDU exchange between nodes residing in the same physical machine and Ethernet for exchange between instances on different physical machines.

The system emulator operates in a similar fashion, except that the layer 2 protocol stacks execute in real-time under RTAI/Linux, again with the potential of a significant number of nodes due to PHY abstraction. This requires a Linux PC with a patched kernel. Again, several instances can reside in the same physical machine. The main difference is speed of simulation and the ability to use real applications or real-time traffic generators to stimulate the protocol stack. Large scale repeatable simulations can be performed on a real protocol stack and performance evaluations with real applications can be realized. Both the layer 2 protocol stacks and the PHY_abstraction unit are real-time processes under RTAI/Linux.

³ Note that the use of the BICM capacity expression depends on the demodulator used.

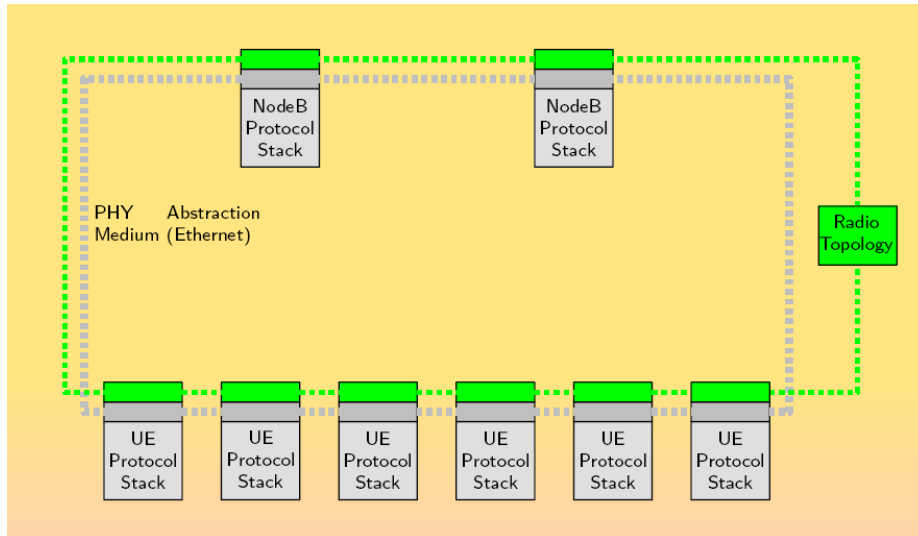


Figure 4.4.1: Typical Simulation/Emulation Setup

4.4.1 OpenAirInterface PHY_Abstraction Unit

The environment of the PHY_abstraction unit is shown in Figure 4.4.2. It takes input from the emulation medium (Ethernet or direct memory transfer) corresponding to the MAC-layer SDUs from corresponding nodes in the network. These SDUs correspond to transport blocks for different transport channels to be encoded by PHY or just decoded by PHY. It also receives information from an RF topology server regarding slowly-varying propagation parameters (mobility, path loss/shadowing models, multipath intensity profiles, etc.). The other end implements the PHY/MAC interface in order to interconnect with the true layer 2 protocol stack.

PHY abstraction is done at the receiver for each link of each node (UE and eNodeB) where wideband spatial signal and interference signal levels are computed every TTI based on the RF topology and pre-defined propagation models. The granularity of the channel simulation is at the level of the sub-carrier, or potentially groups of sub-carriers. The function of the PHY abstraction module can be system dependent (i.e. based on precomputed probability of error simulation tables for specific modulation and coding formats) or generic based on semi-analytical formulae. As in the 802.16m methodology, the output of the radio simulation is random PDU loss indicators for each transport channel block traversing the PHY/MAC interface. Alternatively, if erroneous packets are to be passed to the higher layers, in the spirit of UDP lite-type protocols, bit errors must be generated in the MAC PDUs, corresponding to the residual bit errors at the output of the channel decoder.

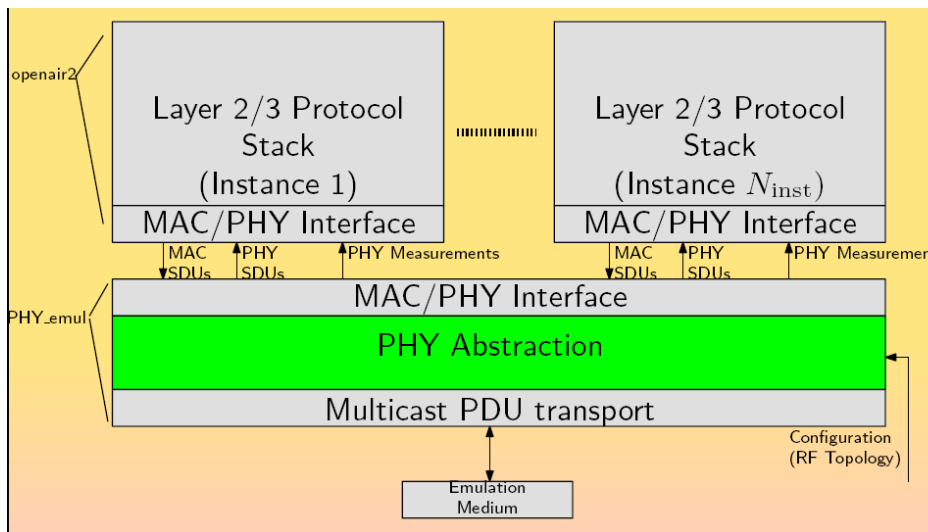


Figure 4.4.2: Environment of the OpenAirInterface PHY_abstraction Unit

In each TTI, the PHY_Abstraction unit analyzes the set of received SDUs from the layer 1 simulation/emulation medium and determines those which are sources of information and those which represent interference. The target PHY SDUs to be received are those programmed by the MAC as with the true PHY. The interferers, however, are naturally present with the true PHY and thus their impact must be simulated in the PHY abstraction unit. Since a particular node in the network is not aware of all sources of interference a priori, this is done by adding a description of the physical resource allocations to each transport block in the simulation/emulation medium which is not present in the real PHY.

A secondary task of the PHY_abstraction unit is to provide the stimulus on the measurement interface for the Layer 2 protocol stack. In system simulations this is required for validation of the Layer 2 mechanisms related to adaptive resource control (modulation and coding adaptation) and wideband resource scheduling which are typically both functions of the MAC-layer scheduling algorithm. Eventually, these measurements also make their way up to Layer 3 protocols which are responsible for connection management and admission control procedures.

4.4.2 An example of PHY Abstraction in OpenAirInterface (Wideband MIMO)

Consider the following example for PHY abstraction at node j in the network. This targets a wideband MIMO system, and most likely multi-carrier (OFDMA or dynamic FDMA in the spirit of UMTS-LTE). Let $RSSI_{i,j}[n]$ be the short-term average received strength in TTI n between node i and j . This can be generated locally in each node based on a model for mobility or can be signalled by a topology server dynamically. As in the 802.16m simulation methodology (Section 3.2) the goal of the PHY_abstraction entity is to simulate the block error rate process (BLER) of each transport block of a particular received resource. Evaluation of the BLER is a function of the statistics of the received signal and interference vectors at node j . Let $\sqrt{RSSI_{i,j}[n]}H_{ij}[n,k]$ be the spatial channel in frequency band k for the signal from node i to j and

$$K_{I,i,j}[n,k] = \sigma^2 I + \sum_{i' \neq i, i' \neq j} \sum_{m=0}^{M(i')-1} \sqrt{RSSI_{i',j}[n]} h_{i',j}[n,m,k] h_{i',j}^*[n,m,k]$$

is the auto-correlation matrix of the received interference for the link from node i to j (note we have assumed zero-mean channels for simplicity) where $h_{i',j}[n,m,k]$ is the spatial channel for transmit antenna m in band k corresponding to interferer i' . $M(i')$ is the number of transmit antennas corresponding to interferer i' . An appropriate choice for the frequency band granularity would be the number of sub-carriers making up a chunk in LTE, for example.

The random variables $\sqrt{RSSI_{i,j}[n]}H_{ij}[n,k]$ and $h_{i',j}[n,m,k]$ depend on the space/time/frequency description of the propagation environment, and more specifically a second-order description of the power-delay profile (PDP), the Ricean factor which relates the energy of direct path and the reflected paths, the antenna correlation, and the relative mobility of the pair of nodes and their environments. As is common in the literature, these are modelled using Gaussian random number generators each TTI based on previously generated channels and modifications to the propagation environment.

Characterization of the BLER

As described in earlier sections, a description of the BLER as a function of $\sqrt{RSSI_{i,j}[n]}H_{ij}[n,k]$ and $K_{I,i,j}[n,k]$ is the key issue in PHY abstraction. This must be done differently for the various coding and multiple-access techniques. Here we outline a few cases which highlight the key issues in system simulation of emerging air interface technologies.

In the case of non-HARQ based transport channels without transmit precoding (beamforming, dirty-paper coding, etc.) we must define a particular receiver structure (bit-interleaving metrics, MMSE, etc.) for which wideband SINR expressions (similar to those considered in 802.16m described in Reference DR.3.1

previous section) can be derived from the above second-order interference description. These are then used for BLER lookup based on tabulated performance of a particular coded-modulation scheme. This consists of running a series of computer simulations of BLER vs. SINR curves for the given average SINR. The difficulty in this approach is that SINR is a vector which is possibly characterized by a significant number of degrees of freedom and thus the offline procedure could be very time-consuming. Another alternative would be the use of PHY-agnostic information-theoretic bounds based on information-outage probabilities under the assumption of either Gaussian codebooks or finite QAM constellations. In the first case, reasonably simple analytical formulae can be used, in the second tabulated pre-computed data must be generated. A third PHY-agnostic possibility would be to use finite block-length error-rate expressions (upper and lower bounds) based on random coding experiments. This is the most challenging approach from a research perspective.

With HARQ-based schemes, errors at a particular time n also depend on past values of the signal and interference components. Here additional protocol information from the MAC signalling channel header are required, but well-known semi-analytical models can be used to describe the BLER on a particular transport block as a function of the round index of the HARQ protocol and the current and past SINR values. This can be used for both repetition coding with Chase combining (type I HARQ) or with Incremental redundancy (type II HARQ). Again this can be done using either tabulated BLER performance curves for the coded-modulation scheme under test or information-theoretic PHY-agnostic formulae.

When precoding is performed based on channel state feedback at the transmitting end, be it linear or non-linear, additional PHY layer information must be transported in the emulation process along with MAC information, namely the linear/non-linear spatial filtering description at the sending nodes. This is required to compute the received SINR at the nodes which now depends, in addition to the channel, on the spatial filtering done at the transmitter. Once this information is incorporated into the SINR characterization, the methods described above can be employed for PHY abstraction.

5. CHALLENGES TO BE ADDRESSED IN WPR.3

In WPR.3, research involves both design methodologies and algorithmic design for selected operational scenarios. Based on the targeted system (modulation wise) of each proposed research effort, a first categorization on four tasks has been proposed. More specifically the targeted systems for AMC design exploration are: (a) Coded OFDM systems, (b) Non orthogonal multicarrier systems, (c) Continuous Phase Modulation based systems, and (d) Single carrier systems employing coherent QAM modulation.

5.1 AMC for Coded OFDM

5.1.1 A Generic Modeling approach

In this approach a model description of the performance of a system capable to support emerging standards (for MIMO-COFDM within WPR3) will be devised, based on selected system deficiencies and environment characteristics. The target is to come up with a model description that is a compromise between simplicity (thus cost-effective implementation) and accuracy. It will incorporate different MIMO options, channel estimation noise, feedback delay and ICI noise (due to phase noise, residual frequency offset) and interference. Based on this description model, specific AMC algorithms shall be devised based on a number of optimization criteria. The target is the wide applicability of the produced algorithms to large number of different MIMO-COFDM systems that can be projected to that model.

It follows that the main target here is to create a “simple” yet sufficiently rich model for total-system description, a computational model which is amenable to run-time optimization. Proper simplification steps (what we have termed “pruning”) are thus viewed as the key to attaining such a model, which will eventually represent a compromise between wide applicability, implementational simplicity and the inevitable performance loss (since we move away from point-optimal solutions). In order to achieve that, we must identify and systematically quantify the effect on system performance of the various parameters or algorithms (computing modules) under this model.

5.1.1.1 Pruning

Pruning implies a meaningful reduction of the overall design space so that the design procedure is simplified, while the corresponding performance loss is kept at a minimum. The selection of the appropriate algorithms for the design space is based on the quantification of the aforementioned trade-off analysis for a chosen scenario. Such a basic sequence of three pruning steps in the design of the PHY layer of COFDM-based MIMO and SISO systems in fairly static environments can be found below.

First Pruning Step: This first step refers to adopting only a subset of Space-Time Codes (STC) that transform a MIMO system into an equivalent SISO channel. A well-chosen “outer” channel code, along with appropriate adaptive modulation techniques can be employed in order to approach the capacity of this equivalent SISO channel.

Recent advances on overall code design have encouraged the separation of the coding procedure into an inner and outer part [Gon02]. The inner code is represented by the STC, whereas the outer code is a classic SISO channel code. Each part of the concatenation tries to exploit different characteristics and properties of the overall channel in order to improve the overall system performance. This inner code usually provides a mixture of diversity/multiplexing /SNR gain, whereas the outer code provides diversity/coding gain. The best choice of the inner/outer code combination will depend on channel characteristics, complexity and feedback requirements (the existence of the latter being a critical system constraint). The majority of STC’s found in the literature are indeed compatible with this desired pruning step. Unfortunately, not all are, as the counter-example of the turbo iterative scheme proposed in [Hoc03] indicates, which approaches the capacity of the underlying channel but does not fall into this category. Instead, this method employs a sphere decoder to compute the bit Log-Likelihood Ratios (LLR) directly from a linear transformation of the MIMO channel inputs, and then

feeds these LLR's to the outer turbo code. On the other hand, inner codes that do fall into this category are, for example, the classic orthogonal block codes [Tar99]. These are good candidate solutions for systems where it is acceptable to sacrifice a bit of performance for lower complexity and feedback requirements. Another case of a compatible inner code is the Singular Value Decomposition (SVD) code, a solution however of high complexity and feedback requirements. SVD along with appropriate water-filling power-loading techniques yield the capacity of the equivalent SISO channel, which is identical to that of the original MIMO channel.

Second Pruning Step: The second pruning step refers to the use of an Effective SINR mapping (ESM) of the equivalent SISO channel demodulator (across one code block) as the basic parameter for uniquely determining the performance of the outer channel code.

As already mentioned in chapter 4, there are several methods for ESM such as:

- Mutual Information Based Effective SINR Mapping (MIESM)
 - Received Bit Mutual Information Rate (RBIR)
 - Mean Mutual Information per Bit (MMIB)
- Exponential ESM (EESM)

An overview of these methods is given in [Bru05], where its performance is evaluated in terms of PER prediction accuracy focusing on a single receive and transmit antenna OFDM link with different coding options and channel characteristics. The results demonstrate that a mutual information based metric which accounts for the modulation alphabet is preferable in the considered cases and, furthermore, applicable to the large class of MIMO-COFDM transmission techniques with linear pre- and post-processing.

Third Pruning Step: The third pruning step models the performance of the environment-characterization unit, e.g., channel and carrier (frequency/phase) estimators, by an appropriate Additive White Gaussian Noise (AWGN) component to the equivalent SISO noise level. All these estimators input the related information into the AMC algorithm, which is then responsible for the selection of the best mode.

5.1.1.2 Objectives

ESM tries to capture the effect of the SINR variation of effective SISO channel in the PER/BER performance of the system. When this variation is low, the average SINR (ASINR) (per codeblock) can be used since it is easily computable and there exist a large number of AMC algorithms that are based on it. Moreover the ASINR is a parameter that changes slowly in comparison with the instantaneous SINR per carrier. Since AMC design based on ASINR has been thoroughly studied in the literature, we will focus on cases where an Effective SINR is needed.

1) Identification of simple rules that indicate the need of an ESINR mapping for performance evaluation and consequently ESINR based AMC algorithms.

The SINR variation is related mainly on two things: (a) the power delay profile (in SISO), (b) the STC scheme (in MIMO). In this effort an evaluation of proposed ESINR methods will be performed, targeting to AMC design and compared to the classic ASINR method. Rules indicating the necessity of ESINR mapping must be extrapolated from the results. These rules can be used to select the best AMC approach based on MIMO-STC or channel statistics.

2) Design of low complexity AMC algorithms based on ESINR mapping methods.

Adaptive bit-loading based on mutual information for an LDPC-coded OFDM system was proposed in [Li07]. They demonstrated that the mutual information between the transmitted binary sequence and the soft information input to LDPC decoder characterizes well the performance of the LDPC-coded system, independent of the modulation type and the number of carriers. They proposed mutual information-based adaptive bit-loading and power allocation algorithms for LDPC-coded OFDM systems to maximize throughput performance under a packet-error rate constraint. To characterize the performance of the studied system, they derived the achievable information rate and a throughput upper bound for the system.

A lot of work has been based on ESM mapping. As shown in [Su99],[Dag03], the turbo-coded performance of a system can be approximated quite well as a function on the uncoded performance at the SNR of interest, whenever a channel bit interleaver is employed. As a consequence, the mode-optimization procedure can be based on this function of the uncoded performance versus the corresponding SNR per coded bit. This fact can easily be extended to other types of codes. There is a large list of bit/power loading algorithms in the literature which are compatible with this Uncoded BER (UBER) approach, under various optimization schemes. The set of those algorithms which are pruning-step-compatible includes those which minimize or guarantee the UBER based on some criteria and constraints per each sample path code block. On the other hand, algorithms that work on the average UBER over all sample paths of coded blocks are not compatible choices for this step. When compatibility does exist, for a specific code, the needed UBER for a corresponding coded BER will be stored in a Look-Up Table (LUT). We note that the entries of the LUT do not just depend on the chosen encoder, since different decoding algorithms can be applied for the same code, resulting in different coded performance.

In [Dag03], two such algorithms have been presented and their performance assessed. The first algorithm calculates the minimum Tx power that guarantees the per-sample-path required QoS in a system with no bit or power loading across the sub-carriers. If the total Tx power is non adapted per channel sample path, the same algorithm can be used to find the modes that satisfy the QoS requirements. The algorithm can work under various system constraints; the price to pay is some performance loss when the SISO channel is highly selective. Nonetheless, the agreed QoS is guaranteed. The second algorithm employs on/off bit loading (also known as Weak Sub-Carrier Excision (WSCE)). This means that a certain number of sub-carriers can be excluded from transmission. With this added feature, the optimization module can select different code rates for the same target bit rate (throughput), something that the first algorithm cannot do. For each of these pairs {code rate-constellation size}, the fixed percentage of excised carriers is computed, in order for all the pairs to provide the same final (target) throughput. The algorithm then calculates the triplet {code rate - constellation size- percentage of excised carriers} that requires the minimum Tx power for a given target BER.

In [Dag05], additional performance gain is achieved by allowing the power to vary across sub-carriers. Three Power-Allocation (PA) schemes were developed, all of which minimize the transmission power for a fixed target BER,: the Optimal (OPA), the Sub-Optimal (SOPA) and the Equal-BER PA (EBPA). These PA algorithms were combined with those proposed in [Dag03] and their performance was assessed through simulation. The use of different codes (namely, punctured patterns of a single mother code), all of which are compatible with the target throughput, on top of the WSCE and PA tools, provides an interesting and efficient mix of two different design philosophies for OFDM systems: water-filling on the one hand, which is mainly used in variable-bit-rate applications, and the PA for BER minimization on the other hand, which is mainly used in fixed-bit-rate applications. The other tool, WSCE, when combined with mode adaptation, can be viewed as a simple water-filling method, but it can also interact beneficially with the proposed PA schemes which target error-rate-performance improvement. All these developed algorithms are based on common (i.e., reusable) functional blocks that can be exploited in the implementation process.

3) Design robust AMC algorithms in the presence of strong ICI due to phase noise and residual frequency offset.

Assuming perfect common phase rotation with the help of pilot symbols or data [Nik05], a closed-form expression that measures the additive noise component due to phase noise is provided in [Wu04]. Assuming a static channel with frequency tap gains $H = [H(0)...H(N-1)]$, normalized residual frequency offset $\Delta\phi$, and phase noise of variance σ_ϕ^2 (Wiener model) the inter-carrier interference (ICI) term for carrier k is given by:

$$ICI(k) = \sum_{l=0, l \neq k}^{N-1} X(l)H(l)I(l-k)$$

where $I(l), l = 0 \dots N-1$ is the FFT of the random phase process plus the residual frequency offset component and $X(l)$ is the transmitted symbol at carrier l . The average energy of the transmitted symbols is assumed known. When power loading is employed, the resulting ICI noise is a function of the power loading profile. Here, $E[|I(p)|^2]$ is the basic term to compute for the ICI and it is given by

$$E[|I(p)|^2] = \frac{1}{N^2} \left\{ 2\Re \left(\frac{d_p^{N+1} - (N+1)d_p + N}{(d_p - 1)^2} \right) - N \right\}$$

where $d_p = e^{j(2\pi(p-\Delta\phi) - (\sigma_\phi^2/2))}$.

The modelling assumption that the noise resembles an AWGN process is very precise for the SNR region of interest.

5.1.2 AMC for BICM-OFDM system based on the goodput criterion

BICM (bit-interleaved coded modulation) was proposed in 1992 by Zehavi as a pragmatic coding scheme for bandwidth-efficient communications [Zeh92]. Specifically, the key idea underlying the BICM is the insertion of a bit-interleaver between the channel encoder and the modulator so as to provide large Hamming distances and to obtain high diversity gains. A theoretical foundation for BICM was given later by Caire, Taricco and Biglieri [Cai98]. Different methods to evaluate the performance of BICM systems have been proposed in the literature. Most of them are based on the union bound and the expurgation technique which was first proposed in [Cai98]. More recently, in [Mar06], thanks to the binary-input output-symmetric (BIOS) nature of the channel, a simple yet accurate computation of the pair-wise error probability (PEP) based on the saddlepoint approximation has been provided. In [McK06] this approach has been extended also to MIMO-BICM systems.

5.1.2.1 AMC in BICM-OFDM Systems

In order to achieve a high spectral efficiency over fading channels, adaptive modulation in BICM was first introduced in [Orm01], wherein two different types of construction for the design of adaptive BICM are proposed. In the first approach, termed “deterministic design”, the adaptive system chooses a modulation scheme under the assumption that the CSI available at the transmitter is perfect, while the second construction explicitly takes into account the variations of the fading channel between channel estimation and data transmission. In [Son06], through the extension of the conventional BICM-OFDM pair-wise error probability (PEP) analysis, a “water-filling based” bit and power allocation algorithm has been proposed to improve the total bit-rate in the presence of slow fading frequency selective channel. In particular, the problem of coding selection with subcarrier rate and power allocation that maximizes the data rate under power and packet error rate (PER) constraint is solved with multiple executions of a discrete rate water-filling algorithm based on the Greedy principle [Cho95], [Cam99]. This method has been extended to MIMO channels in [McK06] and [Sen06], and for reduced feedback information in [Sun07]. In [Sen06], by using the knowledge of the channel at the transmitter, the MIMO channel is split into a bank of scalar subchannels, and only the subchannel with the largest gain is used for the transmission. The AMC problem is then tackled as a discrete-rate capacity maximization problem with a given SNR gap, and solved with the Greedy-based-loading algorithm. In [McK06], the authors consider a MIMO bit-interleaved coded modulation in spatially-correlated Rayleigh channels assuming both statistical beamforming (SB) and spatial-multiplexing with a zero-forcing (ZF) receiver. After estimating the PEP by using the saddlepoint approximation, they propose a practical switching algorithm that jointly selects the best combination of MIMO transmission scheme (i.e., SB or ZF) and the couple of code-rate and modulation format in order to maximize the throughput, whilst satisfying a given target BER. In order to avoid the intensive traffic overheads for the feedback channel to report information on all subcarrier needed by the water-filling AMC scheme, in [Sun07], the authors propose an AMC scheme with reduced feedback information for a BICM-OFDM system.

5.1.2.2 Objective: The Goodput Criterion

In many current applications, only error-free packets are kept by the receiver, while the others are retransmitted through an automatic repeat request (ARQ) retransmission mechanism. In this case, instead of separately considering AMC at the physical layer and ARQ at the data link, a cross-layer design which judiciously combines these layers can be pursued in order to improve the actual system data rate [Liu04]. Therefore, in order to achieve the goal of a cross-layer link adaptation design, an optimized allocation strategy has necessarily to maximize the number of transmitted payload bits in the error-free packet by unit of time, or equivalently the offered layer 3 data rate, or goodput for short. Clearly, depending on the data payload length and the wireless channel conditions, the expected goodput varies with different transmission strategies. The more robust the transmission strategy, the more likely the packet will be delivered successfully, though with a lower spectral efficiency. The key idea for an efficient adaptive cross-layer design, consists in deriving a link adaptation strategy that maximizes the expected goodput as a trade-off between the probability that the packet will be delivered successfully and the shortest possible transmission time. In [Qia02], the goodput was used as a criterion to dynamically select the best physical layer mode for IEEE 802.11a wireless LANs. This has been extended to a multi-user scheme in [Rea05]. In particular, in the case of heterogeneous systems, the maximum expected goodput criterion is used for both the multi-user diversity (MUD) scheduling and the link adaptation. The paper shows that the combination of such scheduling strategy with the link adaptation based on the maximum goodput criterion is effective in improving the layer 3 data rate. Based on the above baseline, an enticing criterion for the power allocation strategy is the maximization of the goodput achievable in a packet-based BICM-OFDM system, where the theoretical foundation of the allocation algorithms lies in a simplified PEP analysis. In [Dev06] the goodput criterion is used to derive a bit and power allocation algorithm for a frame-oriented transmission with convolutional coding and hard Viterbi decoding in an OFDM system. In [Stu08], a power allocation strategy that aims at improving the goodput achievable in a packet-based BICM-OFDM systems is derived. The theoretical foundation of the allocation algorithms consists in a simplified PEP analysis based on Bhattacharyya bound. Moreover, given the optimal power allocation, an accurate estimation of the goodput, based on the PEP saddlepoint approximation, is derived.

5.1.3 Nonlinear distortions

In the context of multicarrier (MC) signaling, a major drawback is the signal vulnerability to nonlinear distortions caused by the high power amplifier (HPA) located in the base station (BS) transmitter or in the hand-held mobile terminal (MT) [Zou95], [Tel03]. The same problem is also experienced in satellite networks, due to on-board amplifiers [Ibn04]. More in detail, MC signals exhibit Gaussian-like time-domain waveforms with high peak-to-average power ratio (PAPR). Such a signal feature would ideally require a transmitter having a linear characteristic spanning over a wide dynamic range. A more pragmatic approach to tackle such an issue simply consists in reducing (i.e., “backing-off”) the signal power at the amplifier input, so as to exploit the linear part of the characteristic. However, this is done at the expense of a reduction of the signal-to-noise ratio (SNR) at the receiver. A first alternative approach consists in trading-off signal distortions with power-efficient use of the amplifier, thus allowing occasional saturations due to signal peaks. In the latter case the operating point of HPA is usually chosen as near as possible to the saturation region (due to the abovementioned power efficiency constraints). As a consequence, the large peaks of the MC signal envelope induce intermodulation products affecting both the signal itself as in-band distortions and the adjacent channels as out-of-band components [Gro94], [Nei95], [Faz98], [Kim00]. This causes a significant performance degradation that, for a reliable and cost-effective network design, has to be accurately predicted. A second alternative consists in generating low-PAPR MC signals. A number of creative methods to generate such a kind of signals have been proposed in literature (e.g. [Li98], [Och00], [Din04], [Mul97], [Gia07]). However, most of them add substantial complexity at the transmitter, require significant coding overhead, degrade the SNR of the system, or can be applied only to a specific subset of MC signals. Moreover, these methods still require a linear channel behaviour over the reduced-dynamic operating region. Assuming, instead, that the HPA operating point is set very close to the saturation region so that the nonlinear distortions cannot be left out anyway, each of the points of the generic user constellation is found to scatter according to two major effects [Kar89], [Gia08]:

- *warping*, i.e., the data symbol is multiplied by a complex-valued factor, and accordingly, each center of mass of the constellation clusters moves away from the position relevant to the linear channel case;
- *clustering* due to the so-called “non-linear distortion noise” (NLDN), which is an additional noise term caused by non-linearities that sums to the thermal noise component, thus further degrading the detection process.

Moreover, in [Gia08] the authors suggest that in case of multi-users transmissions, a further clustering contribution due to multiple access interference (MAI) should be considered, too.

In the context of AMC for MC systems, the joint “action” of warping and clustering effects could significantly modify the expected SNR per subcarrier with respect to the ideal (i.e., linear) case, thus making the estimation of the instantaneous link performance a very harsh task. In this case, an accurate theoretical characterization of the nonlinear distortions could be very helpful for a more comprehensive physical layer (PHY) abstraction.

In the technical literature, many contributions have been devoted to the theoretical characterization of nonlinear distortion effects, both in MC and in single carrier (SC) code division multiple access (CDMA) transmissions. One of the most useful tools proposed in this context is the extension of the Busgang theorem to bandpass memoryless nonlinearities with complex Gaussian nonzero-mean non-stationary inputs [Con02], [Dar00]. Such an approach yields the analytical evaluation of the error probability performance for both CDMA [Con02] and orthogonal frequency division multiplexing (OFDM) systems over nonlinear channels [Dar00], [Pup02], together with the calculation of the spectral properties of distorted OFDM signals as well [Ban00]. In [Din06], bit-error rate (BER) performance analysis is carried out for a specific MC-CDMA signaling format. In [Gia06a], [Gia06b] the MC-CDMA performance analysis has been extended in order to address some important issues, such as the level of out-of-band emissions and the effects of a bandlimited pulse-shaping.

5.1.4 HARQ modeling for AMC design purposes

A major concern in the design of wireless communication systems supporting reliable data transmission is how to control the transmission errors in order to deliver error-free data to the users. To this respect, error handling and recovery policies are implemented by resorting to a joint use of powerful error correcting and/or detection schemes with retransmission of corrupted data, which is commonly referred to as ARQ (Automatic Repeat Request) [Lin84]. There are three basic types of ARQ schemes: i) stop-and-wait, ii) go back-N, and iii) selective repeat. The stop-and-wait scheme is the simplest to implement, however, it has the lowest throughput especially when the propagation delay is long. In order to avoid the drawback of time-wasting delays, the multiple-channel (or parallel) stop-and-wait ARQ has been proposed. ARPANet (Advanced Research Projects Agency Network) supports multiplexing of 8 logical channels over a single link, and features stop-and-wait ARQ on each logical channel [Kle77]. ARQ systems are simple, easy to implement, and provide high system reliability. However they suffer from a severe decrease in throughput when the channel error rate increases. The advantage of obtaining high reliability in ARQ systems can be coupled with the advantage of Forward Error Correction (FEC) systems in order to provide a good throughput even in poor channel conditions. Such a system, called Hybrid ARQ (HARQ), is essentially a combination of Forward Error Correction (FEC) with ARQ, in an optimal manner. When HARQ is used, each erroneously received packet, instead of being discarded, is stored at the receivers, and then combined with retransmitted copies. Two well-known types of HARQ are Chase Combining (CC, also known as Type I) [Cha85] and Incremental Redundancy (IR, also known as Type II) [Man74]. In Type-I HARQ scheme, a coded packet is transmitted initially and if the packet is found to be affected by errors, usually through the use of Cyclic Redundancy Check (CRC), a retransmission request in the form of no-acknowledgment (NACK) is fed-back to the transmitter. Upon reception of this NACK, the transmitter sends the same coded packet again. If the receiver is capable of buffering previously received signals, the optimal solution is to combine these multiple signals according to the maximal ratio combining (MRC) principle, which was first discussed by Chase [Cha73]. In the Type-II HARQ scheme, when a NACK is received, instead of sending the same coded packets, the transmitter tries to construct and send additional coded parities. Another HARQ scheme presented in the literature is the

so called Partial IR (also known as Type III) [Cha85]. This method decreases the coding rate by sending additional redundancy bits while maintaining self-decodability in each retransmission. The retransmitted packet can be Chase-combined with the previous packets to increase the diversity gain. Incremental Redundancy techniques often make use of Rate Compatible Punctured Convolutional Codes (RCPC) or Rate Compatible Punctured Turbo Codes (RCPT). These codes respect the rate compatibility criterion which requires that the puncturing matrices are chosen in such a way that the coded bits of higher puncturing rates belong to the coded bits of lower puncturing rates. Among the HARQ schemes used in practical systems, WiMAX uses multi-channel stop-and-wait Type I HARQ [IEEE05] wherein each HARQ channel is independent of each other; i.e., a data burst can only be retransmitted by the HARQ channel that initially sent it. The HARQ that applies Chase Combining in the multiple channel stop-and-wait is also used in High-Speed Downlink Packet Access (HSDPA) [3GP01].

A comprehensive understanding and accurate modeling of the behaviour of HARQ mechanisms is of particular importance in evaluating the relevant benefits in a large dynamic network system. A critical component needed for an accurate system level evaluation is an effective signal-to-noise ratio (SNR) which can be used to simulate the decoding success/failure of the packet. This is accomplished by means of a proper look-up table mapping the SNR into the corresponding codeword error rate (CER), wherein the table's entries are obtained by simulating the link over an additive white Gaussian noise (AWGN) channel. In the case of CC scheme, the link performance can be modelled by summing together the effective SNRs of the involved packets, whilst in the case of the IR scheme, the link performance is modelled by providing additional coding gain on top of the combined SNR. The IR coding gain can be obtained through link simulations [Fre01], [Fre02] or by developing a suitable analytical model [Che03]. In [Che06], this model is extended to HARQ schemes based on BICM.

5.1.5 Sensitivity analysis and robust techniques

5.1.5.1 Imperfect CSI

There are various aspects of provisioning, handling and utilizing channel state information (CSI) for the purpose of AMC. A first aspect concerns the minimum amount of CSI required for the proper operation of the AMC scheme. This minimum may depend, for example, on the duplexing method and the channel reciprocity, on the number of estimated channels (SISO or MIMO), on the properties of the feedback channel, on the applied AMC scheme, and on requirements of how accurate the information should be (quantization levels, etc.). Feedback reduction techniques, yielding acceptable AMC performance deterioration, are strongly desired. For example, exploitation of the temporal correlation of the channel gain, or source coding with appropriate compression are being proposed [Cho06, Sim07, Hua06a, Hua06b]. A good trade-off between the performance and the required feedback has to be found in order to motivate the usage of the adaptive schemes.

A second important aspect is the accuracy of the obtained CSI. In the past, a large number of research studies that deal with diversity and MIMO systems in the presence of correlated fading were based on the assumption of perfect CSI at the receiver (e.g., [Fos98, Tel99, Ala98, Tar98, Ral98, Kri51, Mal03, Kar03, Ale07]). In [Con07], slow and fast adaptive modulations have been compared through a two-steps analytical methodology. However, in realistic scenarios, the assumption of perfect CSI cannot be held [Tar99]. CSI errors have destructive impact on the performance of the AMC schemes [Jus03] and make the reliability of adaptive schemes questionable (BEP above the desired target value, increasing of the outage probability, etc.). Thus it is very important to assess a framework able to capture the sensitivity of AMC schemes to imperfect CSI.

The accuracy of the CSI is affected by many different factors:

- One of the major CSI error sources is the feedback delay. Due to the time-varying nature of the radio channel and the round-trip delay, the channel status at the time of adaptation is different from the status at the time of the CSI transmission. In many practical situations, it has been observed that the round-trip delays are not negligible, and that the AMC schemes (both throughput-oriented or power-oriented designs) fail to achieve the target reliability levels even when the delay is quite small [Gol97,

Han02]. Therefore, for fast varying channels, it is advised to use channel prediction which also results in imperfect CSI [Due00].

- CSI errors may also result from imperfect channel estimation. The problem of studying wireless systems employing channel estimation has received considerable attention in the context of non-adaptive communication. Typically, the channel coefficients are estimated from known pilot symbols sent among the data (e.g., [Tar05, Tar07, Cav91, Has03]), although also blind joint channel estimation and detection can be applied [Hoc00, Guo03, Ma06]). Furthermore, the impact of channel estimation errors on the system's performance can be studied by modelling the channel estimation error as a complex Gaussian random variable [Mer00] or by considering a particular estimation scheme (e.g., pilot-symbol assisted). The former approach has been adopted to study the performance of diversity and MIMO systems in the presence of fading (e.g., [Luo03, Dia04, Kur05, Lar04, Gar05, Maa05]), whereas particular pilot-symbol assisted estimation schemes have been assumed frequently to analyze the performance of orthogonal space-time block codes (OSTBCs) on block fading channels [Gar05, Gar06, Jac08].

In WPR.3 we aim to extend the current advances in the study of non-adaptive systems in the presence of imperfect CSI to the case of AMC. However, in order to be able to take the effect of feedback delay on the AMC performance into account, more realistic channel models that incorporate the time variations of the channel must be adopted. Such models have been used in [Ma07, Naj05, Sha04] for investigation of the system performance without AMC.

5.1.5.2 Game Theory approach to robust AMC design in OFDM

As mentioned above, an AMC algorithm (which can be viewed as optimization procedure) results in optimal solution only if the CSI is actual and valid. This, however, is not usually the case in realistic systems and environments due to the CSI estimation error, CSI feedback delay, limited feedback rate and changing radio environment. As a consequence, the algorithm is being implemented based on the erroneous CSI, and therefore it does not provide the optimal solution. In some papers the consequences of this problem have been studied with respect to the resulting Bit Error Rate (BER) [Lek98], bit rate [Sou01], fundamental information rate limits [Yin05], power consumption, maximum allowable CSI error and feedback delay, etc. However, there are very few ideas for the AMC robust design, i.e. ideas to incorporate the abovementioned CSI imperfections into the adaptive algorithm, which could handle these imperfections.

In WPR.3, we propose to consider the application of the game theory as a tool to deal with erroneous, not valid or obsolete CSI. Recently, the game theory has been used to model a decisive behaviour of a radio node in a wireless network competitive environment. Depending on the environmental awareness (and its accuracy) the radio nodes can play cooperative or non-cooperative games against each other. Such games when appropriately defined have fair and efficient (in various terms) solutions. Examples of games dealing with network service pricing, quality of service handling, routing, slotted-Aloha medium access control, interference avoidance, power control and trust management in cooperative radio networks are shortly discussed in [Mac06].

We will treat the process of assigning the transmission parameters (the modulation constellation and the coding scheme) as a game in the sense of the game theory, which an intelligent Control Unit (CU) of a radio transceiver plays against the wireless channel. In the game theory such a game is classified as a repeated "game against the nature" [Str02] in which the behaviour of one player is not considered as rational. In fact, the behaviour of the nature, represented by the instantaneous channel gains estimated or measured, is not rational and not reactive to the behaviour of the other (rational) player, namely the CU. The payoffs in this game are intended only for the rational player, and should include the amount of both the achieved throughput and the transmission power saving with respect to a certain reference power level.

The game modelling will naturally involve definition of strategies and utilities in such a way that the expected game outcome will hopefully result in the near-optimum modulation-coding-power loading assignment. Some initial work in this aspect has been published in [Bog08], where only the water-

filling problem has been considered. There, the definition of the utility function provides a player (the user) with some flexibility in determining her preferences concerning the throughput maximization and power consumption. As mentioned above, the considered environment is also treated as a player, although its behaviour is not reactive to the user's adopted strategies. Such an approach is appropriate for modelling the situation of limited environmental awareness, including erroneous CSI.

The main focus of research on the application of the game theoretic models to AMC will be in the OFDM-based systems. The first objective is to suggest methods to deal with CSI imperfections for such systems, and analysis of their performance (in terms of the achieved throughput and bit-error rate), and their cost (in terms of resulting power consumption and computational complexity). The second main objective would be to compare the game-theoretic AMC approach with AMC methods based on classical constraint optimisation with respect to the abovementioned performance and cost in the case of dealing with erroneous CSI.

5.2 AMC for non-orthogonal multicarrier modulation

5.2.1 Introduction to Non-Orthogonal Multi-Carrier Systems

The class of "Non Orthogonal Multicarrier Modulations" (NOMCM) was first introduced in 1998 by W. Kozek and A.F. Molisch [Koz98] as a novel approach to multicarrier transmissions over doubly dispersive channels affected by both time-varying and frequency-selective fading phenomena. In such a kind of propagation scenarios, frequency-selectivity is caused by multipath, while time variations are due to the relative motion between transmitter and receiver.

In the conventional OFDM scheme the data bit stream is divided into many (hundreds or thousands) of substreams, and each of these substreams modulates a different carrier. Since each of these substreams has a low data rate, the intersymbol interference (ISI) caused by the time dispersion effect of the propagation channel is greatly reduced with respect to the single-carrier (SC) case. However, such a big advantage comes at the expenses of interchannel interference (ICI), which arises when the propagation channel is also frequency dispersive and the energy from a subcarrier spills over into the adjacent ones. These two detrimental effects, i.e., ISI and ICI, are thus influenced by two factors: 1) the time dispersion (due to multipath propagation) and frequency dispersion (due to the Doppler effect) of the mobile radio channel, and 2) the shape of the pulse that is used to transmit one symbol on one subcarrier. Pulses that are well localized in the time domain (and thus cause little ISI) are widely spread out in the frequency domain (and thus cause more ICI), and vice versa.

The underlying idea of NOMCM consists in finding a pulse shape that gives minimum distortion for a given Doppler spread (denoted as f_D) and delay spread (denoted as τ_{spread}). In order to better clarify the question, let us consider again a conventional OFDM system, wherein each subcarrier is modulated by pulses having duration coincident with one symbol interval T and sinc-shaped spectrum. Under ideal propagation conditions, i.e., without ISI (e.g., thanks to the use of the cyclic prefix (CP)) and ICI, the pulse spectra on the subcarriers are orthogonal, as in AWGN conditions. However, when a time-varying propagation channel induces a frequency dispersion effect, the orthogonality condition does not hold true any longer, because the subcarrier frequencies do not coincide with the nulls of the sinc-shaped spectrum, thus causing ICI. In order to solve this problem, several pulse-shape optimization techniques have been proposed in the technical literature. For instance, some solutions proposed the use of Nyquist pulses [Cha66], of the CP [Bin90], of time-limited prolate spheroidal wave functions [Vah95], and of Hermitian functions [Haa97]. However, all of these proposals suffer from a severe limitation, as they assume that the pulses must be orthogonal, or that any deviation from orthogonality is a negligible effect. Actually, orthogonal functions are optimum basis functions only in AWGN channels, while in doubly dispersive channels other basis functions turn out to be optimal. The only requirement is that these basis functions form an (incomplete) Riesz basis.

To summarize, the key advantages of non-orthogonal systems (NOFDM) when compared to standard OFDM schemes are as follows:

- a modulation scheme based on incomplete Riesz bases tends to be more robust against frequency-selective fading;
- NOFDM systems lead to better bandwidth efficiency, because the underlying pulse can be chosen with sharper frequency domain decay than that of a comparable OFDM system (thus the spectral efficiency, defined as

$$\eta = \frac{\xi}{TF} \left[\frac{\text{bit/s}}{\text{Hz}} \right],$$

can be maximized; in the equation above ξ denotes the number of bits per symbol and T and F are the time and frequency spacing, respectively);

The NOFDM systems can be treated as a general representation of all multicarrier (MC) systems. In such an approach, the well-known OFDM technique is only a special case of the generic multi-carrier signalling. In OFDM systems the waveform is chosen to be rectangular and the consecutive pulses do not overlap each other in the time domain whereas in frequency domain they are spread over many adjacent subband spectra whilst do not affecting them. In general, the designed pulses used in an NOFDM system can overlap each other in time and/or in frequency domain. It is illustrated in Fig. 5.2.1, where one circle reflects the time-frequency representation of the non-orthogonal pulse used in the NOFDM system, T is the distance between consecutive pulses (atoms) in time domain and F denotes the distance between adjacent subcarriers in frequency domain, N depicts the number of frequency bins and L denotes the number of NOFDM symbols. These pulses are called also atoms in Time-Frequency plane.

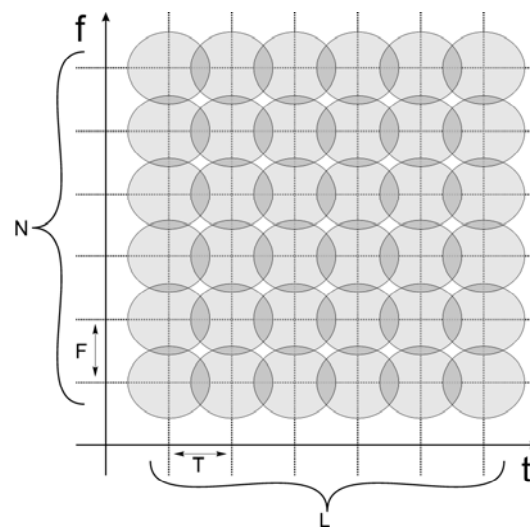


Fig. 5.2.1 - The time-frequency representation of the generic NOFDM signal

The possibility of using various pulse shapes in NOFDM systems allows for transmitter parametrization where the transmit signals are defined by the set of parameters determining e.g. the waveform, sampling rate, overlapping ratio etc. [URA D3.1]. The NOFDM systems are also known as the Generalized Multicarrier (GMC) systems.

NOFDM System Description

The use of nonorthogonal basis functions (Riesz basis) is the key idea presented in the paper by W. Kozek and A.F. Molisch [Koz98], wherein it is also explicitly stated that ACM is not considered, but this is done just for the sake of analysis simplicity and clarity. As matter of facts, ACM remains an appealing option for improving NOFDM performance. The authors also claim that, while it is difficult to draw quantitative conclusions from the uncoded to the coded bit error rate (BER) for doubly dispersive channels, it seems clear that pulse shapes that are optimum for the uncoded case will also give good performance in the coded case. Figure 5.2.2 depicts a conceptual block diagram of a NOFDM system (from [Koz98]), wherein $g(t)$ and $\gamma(t)$ are the so-called prototype pulse and the dual

prototype pulse, respectively, N is the number of subcarriers, T is the multicarrier symbol duration, F is the subcarrier spacing, \mathbf{H} is a linear transformation corresponding to the physical layer of the wireless channel (these effects are mainly due to multipath wave propagation, but also include the undesired linear effects introduced by the receiver/transmitter, such as oscillator drift or timing offsets), $c_{k,i}$ is a complex-valued information-bearing symbol at the transmitter input, and $\tilde{c}_{k,i}$ is the relevant detected symbol at the receiver side.

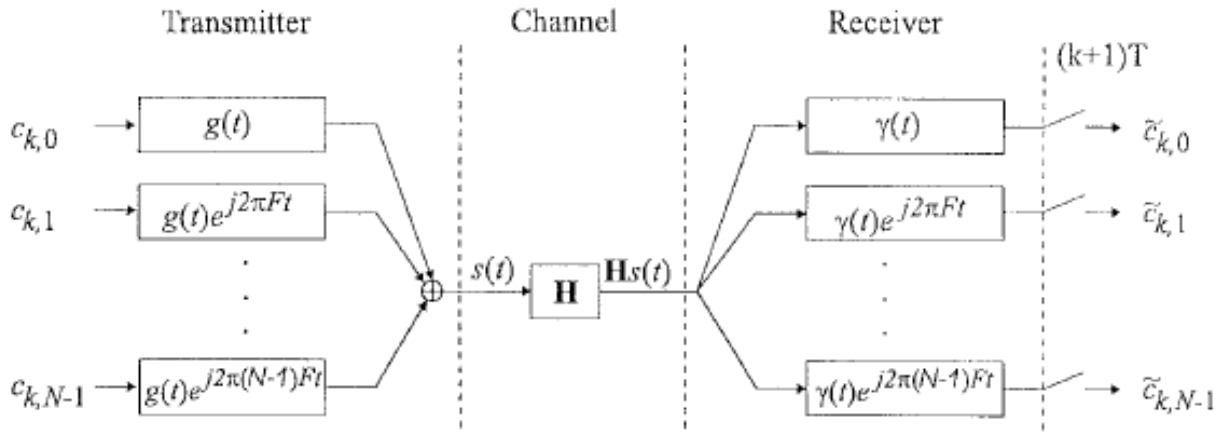


Fig.5.2.2 - The general transmitter/receiver structure of an NOFDM system [Koz98]

NOFDM Pulse Design

Basically, the key idea of NOMCM consists in designing an optimum pair of pulse shapes, as a function of the available a priori information about channel state. To this respect, wide-sense stationary uncorrelated scattering (WSSUS) is one canonical way to formulate statistical a priori knowledge about a doubly dispersive channel. For instance, a typical shape which is assumed for the channel's scattering function is: $\exp\{-\tau/\tau_{spread}\} \cdot 1/\sqrt{1-(f/f_D)^2}$, $\tau > 0$, $|f| < f_D$, where τ_{spread} is the delay spread and f_D is the Doppler bandwidth. The two prototype pulse shapes $g(t)$ and $\gamma(t)$ are then selected by solving a nonlinear optimization problem which involves the channel's scattering function. By discarding the orthogonality constraint, a wider degree of freedom in the design of the pulse shapes can be exploited for obtaining improved performance. By this way, NOFDM pulses, can be matched to arbitrary channel's scattering functions. The resulting pulses, which are not orthogonal in the frequency domain any longer, actually minimize the distortion effect due to both ISI and ICI. Let also notice that the role of the two prototype pulse shapes $g(t)$ and $\gamma(t)$ is interchangeable. Furthermore, another benefit of such an approach consists in an enhanced spectral efficiency, since the pulses can be selected so as to have a frequency support which is more compact with respect to that of the sinc-shaped spectrum used in conventional OFDM.

One can also consider other criteria in pulse designing like minimization of the mean square error of the signal decision at the receiver, or maximization of spectral efficiency.

Filter-Bank Multicarrier Modulation

Filter-Bank Multicarrier Modulation (FBMCM) was originally introduced in the scenario of high-speed wired access networks [Che02] and in the standard for the return channel of terrestrial digital video broadcasting (DVB-RCT) [ETS02]. FBMCM, which is a particular case of NOFDM, differs from conventional OFDM in that the data symbols are transmitted over the different subcarriers after proper pulse-shaping. The result is that the spectra of the subcarriers are bandlimited, bringing forth a number of advantages, namely: i) reduction of sensitivity to narrowband interferers; ii) frequency-domain equalization without the need of the cyclic extension (thus avoiding the consequent efficiency loss); iii) more flexibility to allocate groups of subchannels to different users in the case of Orthogonal Frequency Division Multiple Access (OFDMA).

Moreover, similarly, as in the case of OFDM, the NOFDM transceivers can be also efficiently implemented using the fast Fourier transform (and its inverse) but followed by the polyphase filter

block [Vai93]. The polyphase representation of the transmit filter $g[n]$ is defined as: $R_k[z] = \sum_{k=-\infty}^{\infty} g[nN+k] \cdot z^{-n}$, $0 \leq k \leq N-1$, where R_k denotes the polyphase filter designed for the k -th subcarrier.

OFDM/OQAM-IOTA systems

It is well known from the Gabor frames theory that the well-localized orthogonal set of basis functions cannot be achieved in the critical sampling case, i.e. $TF = 1$ (because of the Balian-Low theorem) [Fei98]. The OFDM/OQAM (offset QAM) systems allow for the usage of the well-localized atoms $g(t)$ in time and frequency for e.g. half lattice density (where $TF=2$) [Bol99]. As mentioned above, the good localization in the TF plane allows to minimize the ICI and the ISI between adjacent atoms.

The baseband discrete-time OFDM/OQAM transmit signal can be described as follows:

$$x_k[n] = \sum_{l=-\infty}^{\infty} \operatorname{Re}\{c_{k,l}\} g[n-lN] e^{j\frac{2\pi}{N}k(n-\alpha/2)} + \sum_{l=-\infty}^{\infty} j \operatorname{Im}\{c_{k,l}\} g[n+N/2-lN] e^{j\frac{2\pi}{N}k(n-\alpha/2)},$$

where $c_{k,l}$ denotes the data symbols, $g[n]$ is the discrete-time pulse used in NOFDM system and $\alpha \in [0, N-1]$.

Based on such an approach, the aim is to design the orthogonal set of pulses that are well-localized in the TF plane. An interesting solution has been proposed in [LeF95], where the isotropic orthogonal transform algorithm (IOTA) function has been suggested for pulse shaping. The IOTA function is the result of the orthogonalization process applied to the Gaussian function. Similar approach, but applied for other function than the Gaussian one, and using the Discrete Zak transform is presented in [Bol99]. In [Kur07] the Hermite function is proposed instead of the IOTA function (as a candidate for OFDM/OQAM systems) because it allows better spectral efficiency.

Gabor signal

The generic multicarrier signal with orthogonal or non-orthogonal modulation using synthesis $g(t)$ and analysis $\gamma(t)$ filters can be described based on the Gabor signal expansion:

$$s(t) = \sum_{l \in \mathbb{Z}} \sum_{m \in \mathbb{Z}} c_{lm} g_{lm}(t),$$

where $\{c_{lm}\}$ ($l, m \in \mathbb{Z}$) are the so called frame coefficients, $\{g_{lm}\}$ is a set of basis functions defined as [Koz98]:

$$g_{l,m}(t) = g(t-lT) \exp(j2\pi mFt)$$

Above, $g(t)$ is the pulse shape also called the *Gabor atom*. Good Time-Frequency (TF) localization of atoms results in reduced intersymbol and intercarrier interference, and allows to omit the guard period necessary in the case of OFDM as it was mentioned in the previous sections. The frame coefficients $\{c_{lm}\}$ constitute the TF representation of $s(t)$. The necessary condition for $\{g_{lm}\}$ to be a Gabor frame is that $TF \leq 1$ [Qia99]. Detailed analysis of the GMC signal generation and reception, as well as of the GMC potential application to multi-standard transceivers can be found in [URA D3.1], [Hun07], [Ste07].

5.2.2 Adaptive Coding and Modulation (ACM) for NOFDM systems - problem formulation

The idea of classical adaptive modulation suggested for OFDM can be extended in a certain way in order to be applied in NOFDM (or GMC) systems, where the signals are defined in the TF plane. Such a two-dimensional (TF) signal representation leads to the necessity to reformulate the *water level* K to *water surface* in the water-filling principle [Gol05]. In such a case, the power allocated to an atom at its particular location in time and in frequency $P(f,t)$ satisfies the following two dimensional condition:

$$P(f,t) = K - \frac{G_n(f,t)}{\alpha |H(f,t)|^2}$$

where K denotes the water surface, $G_n(f,t)$ is the power spectral density of the AWGN, $H(f,t)$ defines the channel TF transfer function, and α is defined as follows:

$$\alpha = -\frac{1,5}{\ln(5 \cdot \text{Pr}_b)},$$

where Pr_b is the assumed bit error probability [Gol05]. The number of bits (defined as the base-2 logarithm of the modulation order M) assigned to the atom localized at time t and frequency f can be calculated in the following manner:

$$\log_2 M(f, t) = \log \left(1 + \frac{\alpha P(f, t) |H(f, t)|^2}{G_n(f, t)} \right).$$

For NOFDM systems the water-filling principle as well as the bit and power loading problem are much more complex. This is because the adaptation algorithm has to incorporate the lack of orthogonality between adjacent atoms and their overlapping in the TF plane. Let us note that altering the power assigned for one atom has its repercussions on power at the adjacent atoms locations, so one can observe that:

$$P(f_m, t_l) = \sum_{l'=0}^{L-1} \sum_{m'=0}^{N-1} P_{c_{l',m'}} |g_{(l'-l), (m'-m)}(n)|^2$$

$$P(f_m, t_l) = P_{c_{l,m}} |g_{0,0}(n)|^2 + \sum_{\substack{l'=0 \\ l' \neq l}}^{L-1} \sum_{\substack{m'=0 \\ m' \neq m}}^{N-1} P_{c_{l',m'}} |g_{(l'-l), (m'-m)}(n)|^2,$$

where $P(f_m, t_l)$ denotes the power assigned to the atom at the TF point (f_m, t_l) , $t_l = lT + nT_s$ and T_s is the sampling period. In the above formula, $P_{c_{l,m}}$ is the power of the Gabor coefficient carried by the pulse $g_{l,m}(t)$. Thus, the two dimensional condition for the assigned power can be rewritten as:

$$P_{c_{l,m}} |g_{0,0}(n)|^2 + \sum_{\substack{l'=0 \\ l' \neq l}}^{L-1} \sum_{\substack{m'=0 \\ m' \neq m}}^{M-1} P_{c_{l',m'}} |g_{(l'-l), (m'-m)}(n)|^2 = K - \frac{G_{l,m}(n)}{\alpha |H_{l,m}(n)|^2}.$$

After integration (summation) of both sides of the above equation with respect to the time variable t (or n in discrete representation), the above relation can be reformulated as:

$$P_{c_{l,m}} = \frac{K}{P_g} - \frac{G_{l,m} + |H_{l,m}|^2 \sum_{\substack{l'=0 \\ l' \neq l}}^{L-1} \sum_{\substack{m'=0 \\ m' \neq m}}^{N-1} P_{c_{l',m'}} P_{g_{(l'-l), (m'-m)}}}{\alpha |H_{l,m}|^2 P_g},$$

where

$$\frac{1}{N} \sum_{n=-N/2}^{N/2} |g_{(l'-l), (m'-m)}(n)|^2 = P_{g_{(l'-l), (m'-m)}}.$$

Let us stress, that the assigned power cannot take negative values, so

$$\frac{G_{l,m} + |H_{l,m}|^2 \sum_{\substack{l'=0 \\ l' \neq l}}^{L-1} \sum_{\substack{m'=0 \\ m' \neq m}}^{M-1} P_{c_{l',m'}} P_{g_{(l'-l),(m'-m)}}}{\alpha |H_{l,m}|^2} \leq K$$

One can observe that the above relations constitute the system of equations and inequalities. The solution of this system is the set of the exact values of the power that should be assigned to the certain atom satisfying the requirement of keeping the total transmit power P constant.

Our goal of the investigation is to examine the impact of the lack of orthogonality between symbols on the ACM algorithms and to propose efficient solutions of adaptive modulation for NOFDM systems. Moreover, extension of such considerations to incorporate also adaptive coding schemes is in focus of our interest.

5.3 AMC for Continuous Phase Modulation

5.3.1 Introduction to CPM

Continuous-phase modulations (CPM) [And86] are a class of modulations featuring attractive spectral properties. Moreover, due to their constant envelope property, continuous-phase modulations are almost insensible to the distortion induced by a nonlinear amplifier working close to saturation. Therefore, even though they have an intrinsically worse trade-off between bandwidth and power efficiency with respect to QAM modulations, when embedded into a nonlinear channel they could yield a higher efficiency. As a matter of fact, new interest has recently grown on the use of CPM, also thanks to the much higher processing power that is today available even in consumer-type equipment. Although CPM modulations have been studied for several years [Rim88, Moq01, Moq03, Lau86, Men95, Sha00], they have not so far found a great number of applications perhaps due to the higher complexity of CPM receivers. The GSM standard is one of few application examples, where a partial-response Gaussian MSK (GMSK) CPM modulation has been adopted.

A CPM modulator is a device that generates continuous-phase, constant-envelope modulated waveforms

$$x(t) = \sqrt{\frac{2E_s}{T_s}} e^{j\psi(t)}$$

where the phase is the time-continuous function

$$\psi(t) = 2\pi h \sum_{n=-\infty}^{\infty} a_n q(t - nT)$$

Here, $h = Q/P$ is the rational modulation index of the CPM scheme (Q and P are relatively prime integers), $q(t)$ is the phase pulse and a_n is the input symbol sequence, whose elements belong to an alphabet consisting of M symbols: $\{0, 1, \dots, M-1\}$.

The phase pulse is defined as

$$q(t) = \int_{-\infty}^t s(\tau) d\tau$$

where $s(t)$ is the frequency pulse, which has the following properties

$$\int_0^{LT_s} s(\tau) d\tau = \frac{1}{2}$$

and $s(t) = 0$ for $t < 0$ and $t > LT_s$. L is usually called *correlation length*.

CPM modulators are devices with memory. According to a well-known analysis approach [Rim88], they can be decomposed into the cascade of a convolutional encoder (the Continuous-Phase

Encoder, CPE) and a memoryless modulator (the memoryless modulator, MM). Figure 5.3.1 shows the decomposition of a CPM modulator in the cascade of the CPE and the MM.

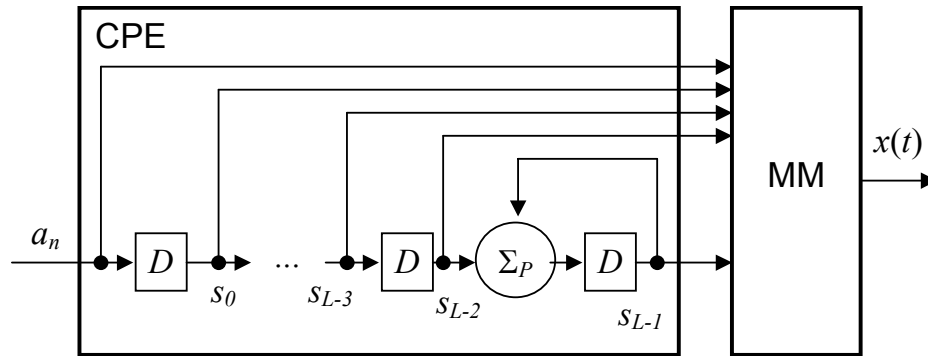


Figure 5.3.1. Decomposition of a CPM modulator.

The CPE is a time-invariant convolutional encoder operating on a ring of integers. The MM maps the CPE code words onto continuous-phase, constant envelope waveforms. This mapping is performed symbol-by-symbol: each output symbol of the CPE is mapped onto one of $P2^{mL}$ signals.

The optimal receiver consists of a log-likelihood ratio (LLR) computer, which computes the soft metrics of the received symbols, followed by a maximum-likelihood (ML) or *maximum-a-posteriori* (MAP) sequence estimator (typically, a Viterbi decoder or a SISO decoder using the BCJR algorithm [Bah74]). Figure 5.3.2 shows the block scheme of a transmission system employing a CPM modulation with decomposed modulator and optimal MAP receiver.

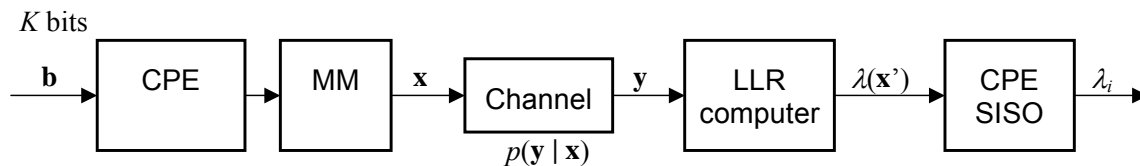


Figure 5.3.2. CPM modulation scheme.

5.3.2 Versatile CPM modulations

A major requirement of ACM systems is the availability of a set of *transmission modes* (corresponding to different QoS and channel states) that will be selected by the link adaptation algorithm to adapt to the current channel state. Each mode corresponds to a set of configuration parameters specifying the coding and modulation formats to be used. Hence, versatile modulation schemes are needed in order to implement easily-reconfigurable modulation units capable of spanning a wide range of spectral efficiencies.

The CPM spectral efficiency is considered a suitable criterion for the study of CPM modulations [Kuo04]. In order to evaluate and compare the spectral efficiency of CPM schemes, it is necessary to choose a suitable bandwidth definition. Usual bandwidth definitions consider a given percentage of the total modulated signal power. Typically, depending on the tolerable level of adjacent channel interference, values between 95% and 99.9% are chosen.

A comparison among CPM modulations can be performed according to the channel capacity of a CPM system. The CPM capacity is defined as the channel capacity (constrained mutual information) of a channel whose input consists of CPM waveforms. It is given by

$$C_{CPM} = \frac{K}{N} E_{y,x} \left\{ 1 - \frac{1}{K} \log_2 \left(\sum_{\mathbf{x}' \in \mathcal{C}} \frac{p(\mathbf{y} | \mathbf{x}')}{p(\mathbf{y} | \mathbf{x})} \right) \right\} = \frac{K}{N} E_{y,x} \left\{ 1 - \frac{1}{K} \left[\max_{\mathbf{x}' \in \mathcal{C}}^* \lambda(\mathbf{x}') - \lambda(\mathbf{x}) \right] \right\}$$

where N is the number of transmitted samples and $\lambda(\mathbf{x})$ are the log-likelihood ratios of the transmitted waveforms. This capacity can be easily computed through Monte-Carlo simulation [Ben03].

An objective of the research on versatile CPM modulations is to define classes of CPM schemes achieving good capacities and suitable for easy reconfigurable implementation. A possible approach consists in defining a suitable affinity relationship between CPM schemes. This relationship may be defined taking into account both the CPM capacity and the resources needed for their implementation (memory, processing power, etc.).

5.3.3 *Versatile and low-complexity CPM receivers*

The CPM demodulator significantly contributes to the overall receiver complexity. Both the LLR computer and the CPE SISO blocks give significant contributions. Indeed, the complexity of the CPE SISO grows exponentially with the correlation length L and it may significantly enhance the overall receiver complexity. An active research area is the study of reduced-complexity receivers for CPM modulations. The problem of reducing the complexity of the LLR computer has been addressed in [Moq03,], where an efficient solution, based on principal component analysis, has been proposed. As for the CPE SISO decoder, standard complexity reduction algorithms [Hub89, Pal93, Sim95] may be applied, but they have not so far shown a good complexity-performance trade-off. An alternative approach to complexity reduction consists in exploiting the characteristics of the Laurent decomposition [Lau86] of CPM into PAM modulations: by neglecting the PAM components with smaller energy, it is possible to significantly reduce the receiver complexity. This approach led to interesting results [Sha02].

The construction of reconfigurable CPM receivers capable of complexity-performance trade-offs over a wide range of spectral efficiencies is still an open issue. The known trellis-reduction techniques [Fra98, Sik05, Col01] can be used to reduce the complexity of CPM receivers, but a deep investigation of their performance is still unavailable. Indeed, the structure of CPM waveforms suggests alternative complexity reduction techniques, which might result in better performance. The search for versatile receiver structures is tightly connected to the search for classes of versatile CPM schemes described in Sec. 5.3.2: in fact, the evaluation of a CPM scheme must take into account both its capacity and the complexity of the receiver.

5.3.4 *Versatile coded CPM systems*

Once the problem of constructing versatile CPM modulators and receivers has been faced, suitable coding schemes must be devised in order to exploit the full capacity of the CPM channel. The scientific literature on the subject proposes two approaches: the traditional serially-concatenated CPM (SC-CPM) approach and the pragmatic CPM (P-CPM) approach. The design of efficient coded CPM systems has been addressed in [Moq01], where a scheme consisting of an outer convolutional code connected to the CPM modulator through an interleaver has been proposed (the SC-CPM approach, see Figure 5.3.4) and its properties have been investigated. At the receiver side, the soft information extracted from the received signal is passed to the inner CPM SISO demodulator [Ben98], which evaluates the extrinsic information that, after deinterleaving, is fed to the outer convolutional SISO decoder. After several iterations of this soft-decoding procedure, a hard decision is made to recover the transmitted bits. Iterating between the outer encoder and the CPE through the interleaver yields rather good performance [Nar01, Nar03]. However, the convolutional code and the CPM modulation are tightly interconnected. In general, the versatility required by AMC systems exceeds that of such coded CPM schemes. Moreover, the CPE state complexity, which is large for spectrally efficient CPM, is enhanced by the number decoding iterations, often leading to unaffordable implementation complexity.

Thanks to their high degree of flexibility, bit-interleaved coded modulation [Zeh92, Cai98] schemes have often been chosen as the modulation and coding units in physical layers with AMC capabilities. A bit-interleaved coded modulation (BICM) system consists in cascading a highly performing variable-rate binary encoder (typically, a punctured turbo or low-density parity-check code) with several modulation schemes with increasingly large signal alphabets. In [Ben07a] a first attempt to design pragmatic schemes in the model of BICM employing CPM modulation (called P-CPM in the following) has been presented. The authors showed that the pragmatic capacity of CPM schemes

heavily depends on the mapping between information bits and CPM signals, and propose a new mapping resulting in an improved pragmatic capacity.

The pragmatic approach does not require decoding iterations between the outer encoder and the CPE, since the CPM is treated exactly as a linear modulation in a bit-interleaved turbo-trellis coded modulation. A nice consequence is that the CPE state complexity is not enhanced by the number of iterations, thus permitting to increase the bandwidth efficiency through the use of a larger number of CPE states: this has not been explored in previous work on this subject [Ben07a, Ben07b]. Moreover, decoupling the CPM modulation and the outer code yields an increased flexibility. This makes pragmatic CPM a good candidate for ACM systems.

In coded CPM systems, the channel encoder is connected to the input of a binary-input soft-output channel consisting of the cascade of the CPM modulator, the “real” (AWGN) channel and the CPM soft receiver. Therefore, the capacity of such channel delimits the region of achievable rates for the considered transmission system. According to Figure 5.3.2, where we define

$$\lambda_i \triangleq \log \left(\frac{p(\mathbf{y} | b_i = 1)}{p(\mathbf{y} | b_i = 0)} \right) = \max_{\mathbf{x}: b_i(\mathbf{x})=1} \lambda(\mathbf{x}) - \max_{\mathbf{x}: b_i(\mathbf{x})=0} \lambda(\mathbf{x})$$

the *pragmatic capacity* of the CPM channel is defined as

$$C_{P-CPM} \stackrel{def}{=} \frac{1}{N} \sum_{i=1}^K E_{\mathbf{y}, \mathbf{x}} \left\{ \log_2 \frac{p(\mathbf{y} | b_i(\mathbf{x}))}{p(\mathbf{y})} \right\} \text{ bits/sample}$$

Simulation results show that the pragmatic capacity of a CPM modulation is significantly smaller than the CPM capacity for almost all considered CPM parameter values. The E_b/N_0 gap between these two curves represents the energy loss incurred by using the CPM channel as a binary-input communication channel. Such loss is often several dBs. A typical result is shown in Figure 5.3.3, where the CPM pragmatic capacity curve (labelled “Pragmatic capacity, F = [000]”) is roughly 3 dB from the CPM capacity.

This capacity loss is due both to the properties of the CPM signals and to the mapping $b_i(x)$ between CPM input bit sequences and CPM waveforms. Indeed, in order to obtain a low capacity loss, pairs of binary sequences with low Hamming distance should be mapped to pairs of waveforms with low Euclidean distance in signal space. Due to the recursive nature of the original CPE, this is not true. In fact, pairs of binary sequences with Hamming distance 1 are always mapped to pairs of waveforms with infinite Euclidean distance. Conversely, pairs of waveforms at minimum Euclidean distance correspond to pairs of input bit sequences with Hamming distance at least 2. In order to increase the pragmatic capacity, it is therefore necessary to modify such mapping. In [Ben07b] a CPM encoder resulting in an improved pragmatic capacity has been presented.

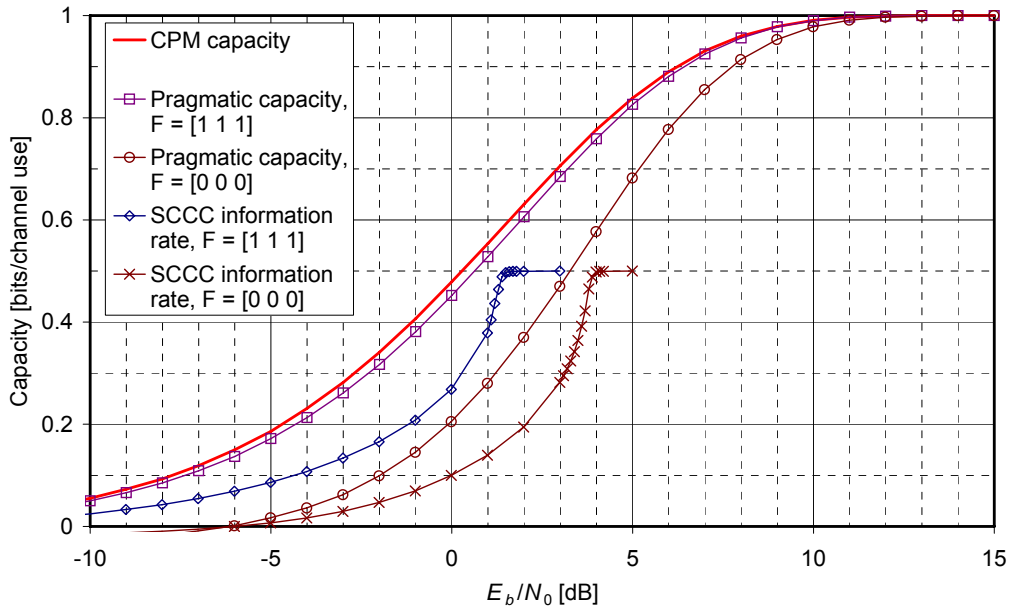


Figure 5.3.3. Capacity curves for a 3REC CPM binary scheme with $h = \frac{1}{2}$.

5.3.4.1 Serially concatenated CPM schemes

In the context of SC-CPM schemes (see Fiture 5.3.4), code versatility must be achieved by varying the parameters of the outer convolutional code and the interleaver. As for the outer convolutional code, the block length can be easily varied adopting a suitable trellis termination technique. Its rate can be varied by means of suitable puncturing algorithms. Variable outer code word lengths require the availability of interleavers with different sizes. This problem has been faced before in the context of concatenated codes (see, e.g. [Din05]). Similar techniques could be adapted to SC-CPM schemes: this topic will be the subject of further investigation.

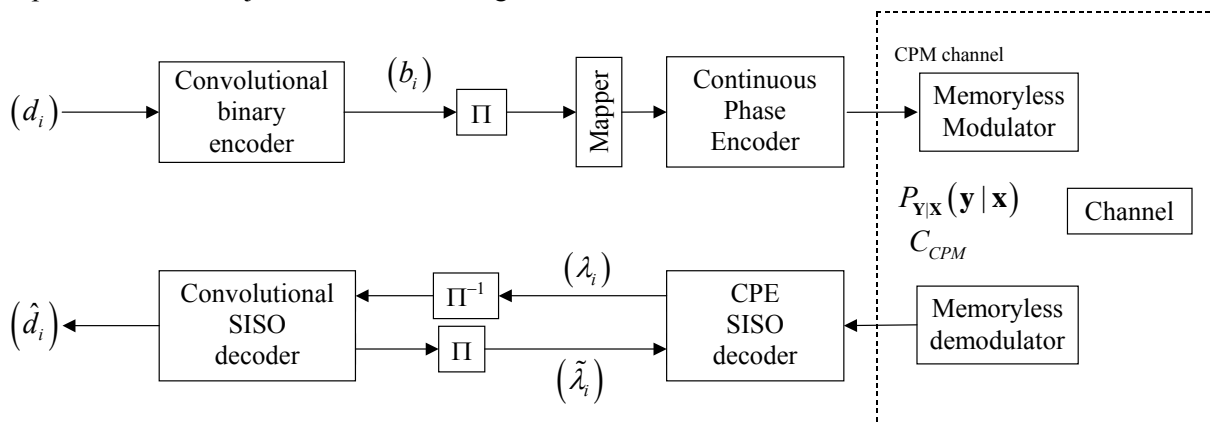


Figure 5.3.4. SC-CPM Scheme.

5.3.4.2 Pragmatic CPM schemes

In P-CPM schemes (see Figure 5.3.5), the problem of designing versatile encoders and versatile CPM modulators is decoupled. In principle, the versatile coding schemes proposed for ACM systems with linear modulations can be adopted. However, the characteristics of CPM modulations may cause significant performance degradation of binary codes with respect to their performance on linear modulations. The performance evaluation of binary codes on the CPM channel and the design of channel codes matched to the CPM modulation will be the subject of further investigation.

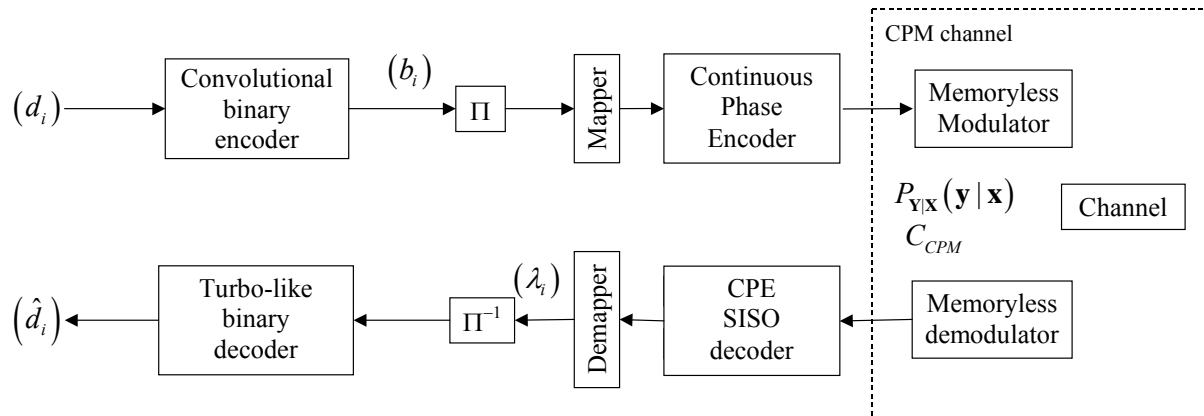


Figure 5.3.5. P-CPM scheme.

5.3.5 Objectives and future activities

The focus of the CPM research activity will be on versatile CPM and coded CPM schemes. The main tasks are the following:

- Versatile CPM modulation schemes: the scope is the study and performance estimation of classes of spectrally efficient and energy-efficient CPM modulations suitable for implementation in reconfigurable transceivers. Versatile receivers for such modulation schemes will be studied with particular attention to their reconfigurability characteristics and complexity-performance trade-offs.
- Versatile coded CPM systems: the studied versatile CPM schemes will be embedded into SC-CPM and P-CPM coded schemes. Their versatility characteristics and complexity-performance trade-offs will be studied.

As a main result, these tasks will provide elements for the design of versatile CPM coded-modulation units to be embedded in transceivers with ACM capabilities.

5.4 AMC for Coherent QAM modulation

5.4.1 Introduction

In almost every adaptive technique, the receiver must reliably and rapidly estimate the instantaneous SNR to track the fast fading evolution. Based on this estimate, the receiver must send the constellation size to be used reliably and instantaneously back to the transmitter. To reduce the channel load caused by feedback, model-based *channel prediction* algorithms can be adopted [Due00]. With this method, it is essential to predict the channel coefficients several tens-to-hundreds of symbols ahead. Moreover, in rapidly changing mobile radio environments, the vehicle speed and scattering geometry change continuously, and thus the model parameters need to be recomputed frequently [Due00]. We denote these adaptation techniques that require the knowledge of the fast fading channel conditions as FAM.

As an alternative to FAM, one may consider the SAM (Slow Adaptive Modulation) scheme in which parameters are adapted based on the performance averaged over an interval of few seconds (i.e., the BEP averaged over fast fading). With this technique, the constellation size needs only to follow the slow variation of the channel; hence it requires slower feedback rate and consequently it is simpler to implement than FAM. Indeed, the feedback rate is related essentially to the coherence time of the channel. As an example, for a mobile terminal the coherence time of the fast fading is inversely proportional to the maximum Doppler frequency: with a carrier frequency of 900 MHz the coherence time is about 72 ms and 4 ms for a mobile speed of 3 Km/h and 50 Km/h, respectively. On the other hand, the coherence time of the shadowing is proportional to the coherence distance (e.g., 100 – 200 m can be considered in suburban area and some tens of meters in urban area [Gud91]). Assuming a

coherence distance of 100 m, results in a coherence time of about 120 s and 7.2 s at 3 Km/h and 50 Km/h, respectively. Note that the coherence time of the fast fading can be an order of magnitude smaller than the coherence time of the shadowing. The importance of slower feedback rate is accentuated especially in the design of TDD systems due to the inherent delay in the feedback channel.

In [Con07], it has been proposed a SAM scheme with M-QAM using SSD. It is analyzed the SAM performance for coherent reception of M-QAM with SSD, which utilizes only a subset of the total available diversity branches [Win03]–[Mol05]. As a bench-mark, [Con07] starts by analyzing the capacity of adaptive transmission schemes with MRC in composite Rayleigh fading and shadowing channels. Then, SAM, FAM and fixed modulation schemes with SSD were compared in terms of both SE and BEO. Results show that the SAM technique can provide substantial improvement over a non-adaptive scheme in terms of both SE and BEO. Furthermore, the performance of SAM can be close to FAM in many scenarios despite the reduced feedback rate, thus the less complexity, with SAM.

5.4.2 Link level methodology for fast changing channels

In general, for a given received power, the BEO increases with the constellation size M . On the other hand, by increasing M the system becomes more efficient in spectrum utilization, that is, the SE increases. When a set of modulation levels $\{M_0, M_1, \dots, M_J\}$ is used, the minimum BEO and the maximum SE correspond to M_0 and $M_{\max} = M_J$ respectively. The system is in outage when even the smallest constellation size does not meet the target BEP. Therefore SE ranges from 0, when the system is always in outage, to $\log_2 M_J$ [bits/s/Hz], when the system never experiences outage. For a fixed target BEP, P_b , the adaptive modulation (AM) technique enables an increase in SE by increasing the constellation sizes M_j to the next available M_{j+1} when the SNR per bit exceeds the SNR required for M_{j+1} . On the other hand, it decreases the constellation size when the SNR is not sufficient to guarantee the target BEP for M_j . Hence, the transmission system using AM experiences the same BEO as a fixed (non-adaptive) modulation system with $M = M_0$, and can achieve a substantial increase in SE due to the fact that AM utilizes the largest possible constellation size (ranging from M_0 to M_J). Terms BEP and required SNR are related to the mean BEP and SNR (averaged over fast fading) when the target QoS is averaged over fast processes and AM tracks slow fading variations (SAM), while they are related to instantaneous BEP and SNR when the target QoS is in terms of instantaneous BEP and AM tracks fast fading variations (FAM). This is shown in Fig. 5.4.1, where different BEP curves are related to different system configurations among those the adaptive modulation and coding adaptation block can choose.

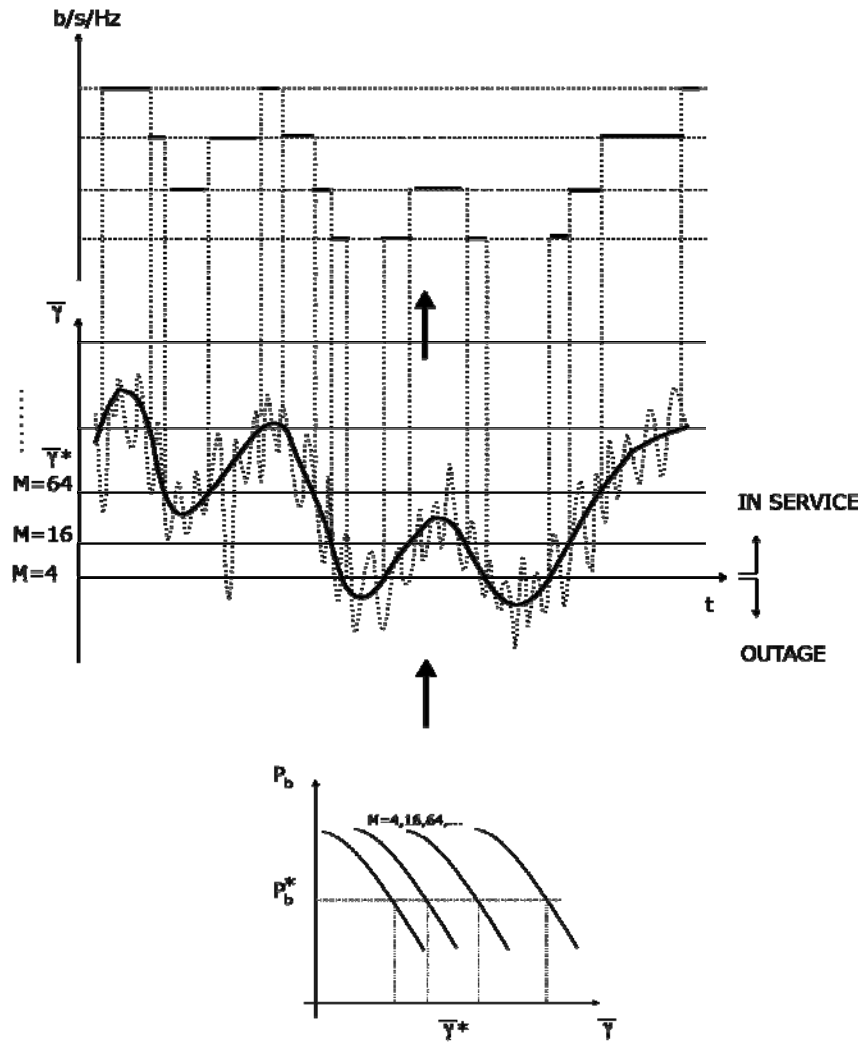


Fig. 5.4.1: The methodology for SE and BEO evaluation [Con07].

Once obtained the required SNRs for each system configuration at a target BEP (which can be evaluated analytically or by simulation depending on the type of system and channel model), one can use the analytical methodology in [Con07] to obtain both the mean spectral efficiency and the bit error outage.

In particular, the *first step* is to obtain the required SNRs (adopting notation $\bar{\gamma}_j^*$ to indicate required mean SNR for a target mean BEP P_b^* in the case of SAM and without overline for instantaneous SNR and BEP in the case of FAM, j refers to the j th operating mode) enabling each system configurations to reach the target BEP.

The *second step* is to derive the SE; as an example for SAM results to be given by

$$\begin{aligned} \eta &= \sum_{j=0}^{J-1} \tilde{M}_j \mathbb{P} \{ \bar{\gamma}_{\text{dB},j}^* < \bar{\gamma}_{\text{dB}} \leq \bar{\gamma}_{\text{dB},j+1}^* \} \\ &\quad + \tilde{M}_J \mathbb{P} \{ \bar{\gamma}_{\text{dB},J}^* < \bar{\gamma}_{\text{dB}} \} \\ &= \sum_{j=0}^{J-1} \tilde{M}_j [F_{\bar{\gamma}_{\text{dB}}}(\bar{\gamma}_{\text{dB},j+1}^*) - F_{\bar{\gamma}_{\text{dB}}}(\bar{\gamma}_{\text{dB},j}^*)] \\ &\quad + \tilde{M}_J [1 - F_{\bar{\gamma}_{\text{dB}}}(\bar{\gamma}_{\text{dB},J}^*)] , \end{aligned}$$

where $\tilde{M}_k = \log_2 M_k$ and $F_{\bar{\gamma}_{\text{dB}}}(\cdot)$

Reference DR.3.1

The BEO for adaptive schemes is given by the probability that the scheme more robust with respect to disturbances (typically the one with lower constellation size and SE) does not meet the BEP requirement.

As example, in the case of small-scale fading superimposed to log-normal shadowing, that is the mean SNR (averaged over small scale fading) in dB is a Gaussian r.v. having mean μ_{dB} (the median SNR) and standard deviation (shadowing parameter) σ_{dB} it is given by

$$P_o(P_b^*) = Q\left(\frac{\mu_{\text{dB}} - 10 \log_{10} \bar{\gamma}^*}{\sigma_{\text{dB}}}\right)$$

where $Q()$ is the Gaussian Q -function.

Note that this methodology is quite general and can be specialized to several digital wireless communication systems/standards working both in small-scale and large-scale fading. Depending on the performance index of interest for the particular system and data traffic the BEP, thus required SNRs thresholds, would be instantaneous or averaged over small-scale fading, for different diversity techniques, channel estimation techniques, adaptive modulation schemes.

In particular, in [Con07] authors derived the BEP and required SNRs for both FAM and SAM for general M-ary QAM with SSD (including maximal ratio combining, MRC, selection diversity, SD and hybrid-selection MRC, H-S/MRC); from these results mean spectral efficiency and BEO are given in Rayleigh fading and log-normal shadowing, but the methodology is general. The SE for FAM can be evaluated using methodology similar to the one for SAM except that the instantaneous SNR is compared to the thresholds obtained from the instantaneous BEP.

For N-branches H-S/MRC the mean BEP for M-QAM with coherent detection results to be given by

$$P_b(\bar{\gamma}) = \frac{2}{\sqrt{M} \log_2(\sqrt{M})} \sum_{h=1}^{\log_2(\sqrt{M})} \sum_{i=0}^{(1-2^{-h})\sqrt{M}-1} (-1)^{\lfloor \frac{i \cdot 2^{h-1}}{\sqrt{M}} \rfloor} \left(2^{h-1} - \left\lfloor \frac{i \cdot 2^{h-1}}{\sqrt{M}} + \frac{1}{2} \right\rfloor \right) \times \mathcal{I}\left(\frac{3(2i+1)^2}{2(M-1)} \tilde{b}_n\right),$$

where

$$\mathcal{I}(x) = \sum_{n=1}^{\tilde{N}} \sum_{k=1}^{\mu_n} A_{n,k} I_k(x),$$

$$I_k(x) = \frac{1}{\pi} \int_0^{\pi/2} \left(\frac{\sin^2 \theta}{\sin^2 \theta + x} \right)^k d\theta$$

and

$$\tilde{N} = N - L + 1,$$

$$\mu_n = \begin{cases} L & n = 1, \\ 1 & n = 2, \dots, \tilde{N}, \end{cases}$$

$$\tilde{b}_n = \begin{cases} \bar{\gamma} & n = 1, \\ \bar{\gamma} L / (L + n - 1) & n = 2, \dots, \tilde{N}. \end{cases}$$

In case of others kinds of system modeling the thresholds can still be obtained analytically in some cases or by simulations. Once obtained SNR thresholds the rest of the procedure is purely analytical.

To summarize, from the above methodology the system designer can obtain the minimum value of the median SNR for specified level of performance. Since the actual median SNR is tied to propagation law and location of the user, one can then determine the cellular system in terms of cell size, power levels, etc., that fulfill the specified requirements. The performance of SAM results to be close to FAM despite the need for a lower feedback rate and thus less complexity with SAM. The methodology is applicable to other modulation formats, diversity techniques and fading channels. By using the proposed methodology, one can obtain SE and BEO for various channel parameters as well as for different diversity techniques.

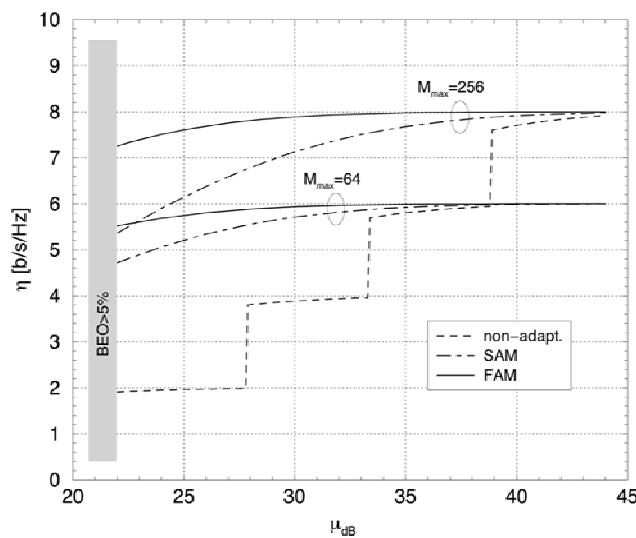


Fig. 5.4.2: Mean SE vs. median SNR for FAM, SAM and nonadaptive QAM. Maximum constellation size 64 and 256, maximum BEO of 5%, target BEP=0.01 [Con07].

In Fig. 5.4.2, comparison examples among SAM, FAM, and fixed-modulation schemes are given in terms of both SE and BEO for target BEP of 10^{-2} , typical for uncoded systems, shadowing parameter $\sigma_{dB} = 8$, median SNR $\mu_{dB} = 30$, and different values of branches N and maximum constellation-size M_{max} . As an example, for BEO of 5% with single-branch reception and M ranging from 4 to 256, non-adaptive schemes can achieve about 1.9 [bits/s/Hz], whereas SAM can achieve 5.4 [bits/s/Hz] and FAM can achieve 7.3 [bits/s/Hz].

Figure 5.4.3 shows that both BEO and SE can be enhanced by using antenna diversity, that is, when the diversity order, N , increases the performance of SAM technique approaches that of FAM despite the need for lower feedback rate. In fact, for BEO of 5% with dual-branch reception, non-adaptive schemes can achieve about 3.9 [bits/s/Hz], whereas SAM can achieve 7.1 [bits/s/Hz] and FAM can achieve 7.9 [bits/s/Hz]. For $N = 4$, the non-adaptive schemes can achieve about 5.8 [bits/s/Hz], whereas 7.6 [bits/s/Hz] and 7.9 [bits/s/Hz] are achieved by SAM and FAM, respectively. It is important to note that SE for fixed modulation scheme does not necessarily increase with the constellation size since larger constellation size can result in more frequent system outages.

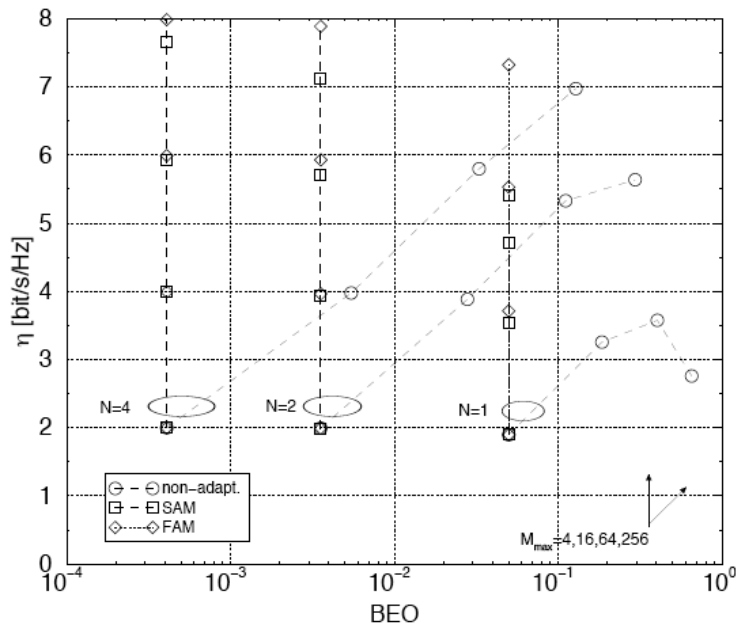


Fig. 5.4.3: Comparison between slowly and fast adapting, and non-adapting schemes in terms of BEO and mean SE for target BEP equal to 10^{-2} , $\sigma_{dB} = 8$, $\mu_{dB} = 30$, and MRC with different diversity orders, N [Con07].

5.4.3 Objectives

The framework for the derivation of required SNRs for each system configuration is, at the state-of-the-art, based on some assumptions that we intend to relax within WPR.3 to enable the design of AMC schemes for practical systems in realistic scenarios. In particular, we think that the following gaps have to be filled:

- Evaluate the performance of FAM and SAM in others channel models for different scenarios, as example by considering correlated generalized fading channel models (e.g., Nakagami-m, generalized Gamma) and various diversity scenarios (e.g., generalized selection, switched combining). Evaluate the impact of frequency selective fading, usually met in realistic scenarios, in the performance of such systems.
- Consider the presence of non-ideal channel estimation and study the impact of imperfect decision on the system configuration to be adopted. Also, the impact of imperfect decision caused by channel estimation errors on the system configuration to be adopted needs to be investigated. These errors need to be considered in design adaptive techniques either by applying a particular channel estimation scheme (e.g., pilot symbol assisted, blind or semi-blind techniques) or by modeling the estimation error (e.g., complex Gaussian random variables with proper parameters, additive noise on the actual channel gain).
- Evaluate the impact on the performance (e.g., SE and BEO) when adapting concurrently modulation, coding and, if present, diversity techniques.
- Investigate the impact of AMC schemes on upper layers protocols.
- Study the impact of adaptive transmission techniques on mutual interference when nodes are in a network. Evaluate the performance of adaptive techniques in generalized fading channels and if diversity is used in correlated ones in the context of a network.

6. CONCLUSIONS

This first WPR3 report has addressed the current status of the state of the art in AMC algorithmic design. It has highlighted the main results of the last two decades, both in the theoretical research and the actual implementation dimensions. Additionally, the report lists new promising directions and possibilities in AMC algorithmic design, mainly driven by the multi-modality of the new emerging standards. In the final Section, the report presents open research issues and proposes certain relevant research objectives to be pursued in the life of the Network of Excellence, always within the scope of the approved Technical Annex

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