**Abstract:**

This deliverable addresses the development of efficient and reliable radio interface for the broadband, QoS-guaranteed, communication link between the HAP, which offers the Internet access and broadband multimedia services, and pools of end-users, equipped with LAN/WLAN terminals connecting through a collective terminal, mounted either on the rooftop of a building (fixed scenario) or on board of a train (mobile scenario)in the mobile wireless access segment, with QoS requirements. The focus is on the design and performance analysis of adaptive coding and modulation schemes, variable-rate variable-length concatenated codes, spatial and polarization diversity, MIMO coding schemes, and on the efficient implementation and testing of adaptive channel estimation and equalization algorithms. The proposed approaches are shown to be able to either significantly increase the overall throughput, or to improve bit error rate performance with respect to standard solutions.

**Keyword list:** advanced signal processing, adaptive coding and modulation, serial and parallel concatenated convolutional codes, coded polarization, trellis-code modulation, adaptive equalization.
### DOCUMENT HISTORY

<table>
<thead>
<tr>
<th>Date</th>
<th>Revision</th>
<th>Comment</th>
<th>Author/Editor</th>
<th>Affiliation</th>
</tr>
</thead>
<tbody>
<tr>
<td>29/04/05</td>
<td>01</td>
<td>First issue</td>
<td>Emanuela Falletti, Marina Mondin</td>
<td>POLITO</td>
</tr>
</tbody>
</table>

Document Approval (CEC Deliverables only)

<table>
<thead>
<tr>
<th>Date of approval</th>
<th>Revision</th>
<th>Role of approver</th>
<th>Approver</th>
<th>Affiliation</th>
</tr>
</thead>
<tbody>
<tr>
<td>29/04/05</td>
<td>01</td>
<td>Editor (internal reviewer)</td>
<td>Alister Burr</td>
<td>UOY</td>
</tr>
<tr>
<td>29/04/05</td>
<td>01</td>
<td>On behalf of Scientific Board</td>
<td>David Grace</td>
<td>UOY</td>
</tr>
</tbody>
</table>
EXECUTIVE SUMMARY

In BWA systems, the wireless link is likely to be the bottleneck in any end-to-end mobile wireless broadband system. Therefore, the improvements of wireless link performance, in terms of spectral efficiency, channel utilization and QoS capability, directly translate to the overall improvement of the end-to-end system.

The CAPANINA project aims at the delivery of broadband services through a collective terminal able to serve all the passengers on board of a train via a HAP system. Passengers travelling on the train are equipped with WLAN terminals and access the network via a collective terminal through the HAP and a gateway located in the ground station, requesting broadband multimedia services from various content providers. In this scenario HAPs are used as a collective access network, providing BWA services to the users.

The peculiarity of the HAP propagation channel on a high-speed train, affected by the Doppler effect, shadowing from atmospheric phenomena, and blocking at low elevation angles, results in a quite critical link budget, in particular for high transmission rates. Therefore, the adoption of powerful signal processing techniques is expected to guarantee elevated service availability even for high-rate applications.

The aim of this report is to propose advanced signal processing techniques especially tailored to cope with the time variant response of the HAP-to-train / train-to-HAP wireless communication channel with reference, in particular, to the IEEE 802.16 Standard, which is suited to the HAP environment. In this report both Single Carrier and OFDM flavors of the IEEE 802.16 (IEEE 802.16a) standard are considered. Although CAPANINA WP2.1 has recently selected IEEE 802.16-SC (Single Carrier) as the reference standard for broadband access to the CAPANINA communication services, at the early stage of the work on WP2.3, neither this information, nor that related to the wireless propagation in K/Ka bands, was available, so that OFDM formats were considered as potential candidates for the physical layer. Once deliverable D09 on communication standard selection [35] and milestones [30, 36] have been issued, the focus of the work has switched to Single-Carrier modulation schemes in presence of fading channels with either strong line-of-sight component, or without LOS (in the case of low elevation angles). Nonetheless, the analysis developed on OFDM schemes, jointly with WP2.1, has helped reach the decision on choice of standard. The presence of a few, well differentiated multipath components are restricted to specific conditions, such as signal propagation in dense urban scenarios. Three reference application scenarios are identified for HAP communications to/from trains, characterized by a medium-to-high physical bit rate, high Doppler shift, and line-of-sight propagation with different Rice factors. They are intended for different kinds of services, and have therefore different QoS requirements in terms of BER (bit error rate) and maximum allowed delay.

The report firstly analyzes the performance of some of the standard coding and modulation schemes proposed in IEEE 802.16/802.16a, in both AWGN and the selected flat-fading scenarios, particularly interesting for the CAPANINA applications to trains. It is worth noticing that the standard 64- QAM modulation is able to reach the highest physical bit rate foreseen for CAPANINA applications, i.e. 120 Mbit/s or more. The information bit rate will be evidently reduced by a factor equal to the applied channel code rate. On the other hand, in the case of strong channel impairments, the system will switch its transmission format to lower level modulations (e.g. 16-QAM or QPSK) and/or more powerful channel codes, in order to reduce blocking probability.

The adaptive selection of the coding and modulation scheme on the basis of a proper metric of the received signal quality is shown to be able to increase the bandwidth efficiency (and therefore the overall throughput) with respect to fixed allocation schemes based on static traffic descriptor metrics. In this case an average information rate between 12.5 Mbit/s and 50 Mbit/s for a signal-to-noise ratio ranging in 5-15 dB is expected for line-of-sight propagation. If standard IEEE 802.16 coding formats are applied, the average information rate is expected to vary between 25 to 75 Mbit/s.

Another topic of primary importance for power-limited applications such as transmissions from HAPs is the possible reduction of the required power to achieve a target BER over fading channels, typically through the use of powerful coding schemes. If advanced coding solutions are applied, such as serial or parallel concatenated convolutional codes, 1.4 to 2 dB gain are expected with respected to the standard coding solution, given the same code length. As an alternative, serial or parallel concatenated convolutional codes with smaller frame size can be used to achieve the target BER at the same required
power of the standard code. Clearly, the choice of a code not included in the IEEE 802.16 standard would require the design of proper and proprietary transmitting and receiving devices especially tailored for the HAP applications.

In the case of significant multipath fading, enhanced performance with respect to single antenna/single polarization transmissions can be achieved by using coded polarization diversity solutions. All the coding and modulation schemes are analyzed assuming ideal channel estimation and synchronization. Furthermore, performance can be greatly improved in the presence of moderate multipath by the use of an ideal equaliser at the receiver. Therefore, the use of a Kalman-filtering approach to perform channel estimation and equalization in a channel with a few -multipath components, or flat fading channel, is suggested to reach quasi-ideal working conditions, even for high bit rates (up to 120 Mbit/s over the channel), high-level modulations and medium/high Doppler frequency shift (4.5 kHz).

Since Single Carrier modulations are inherently quite robust with respect to Doppler frequency shifts, it is expected that common synchronization algorithms are able to perform Doppler recovery. However, this topic will be object of further investigations within WP2.3, with particular attention to the performance of the selected communication standard for high-speed moving terminals.
# TABLE OF CONTENTS

1 Introduction ........................................ 15

2 QoS for Broadband Wireless Services .............. 18
   2.1 QoS in ATM Networks .......................... 18
   2.2 QoS in IEEE 802.16 Links .................... 19

3 Reference System Performance for IEEE 802.16 Standard .................. 23
   3.1 System performance for 256 OFDM transmission scheme ........... 23
      3.1.1 System performance evaluation in AWGN and multi-path environment .... 23
      3.1.2 Spectrum Spreading of the 256 OFDM Signals in Non-Linear Channel .... 30
   3.2 System Performance for Frequency Bands from 10 to 66 GHz ......... 35
      3.2.1 System Performance Evaluation of Single Carrier Modulation in AWGN Channel .... 35
      3.2.2 System Performance Evaluation of Single Carrier Modulation in Fading Channels .... 40
      3.2.3 Distortion of single carrier modulated signal in non-linear channel .... 40

4 Propagation Issues in the Stratospheric Channel .................. 45

5 Operating Scenarios and System Architecture .......... 50
   5.1 Selection of Reference Scenarios ................ 51
   5.2 An Example of Link Budget .................... 52

6 Adaptive Coding and Modulation ...................... 54
   6.1 ACM Performance Evaluation .................... 55
   6.2 Performance of ACM Combined with Spatial Diversity .............. 57
   6.3 Performance of ACM Applied to IEEE 802.16 Modulation&Coding Formats .... 59

7 Advanced Coding Techniques in OFDM-based IEEE 802.16a standard .... 60
   7.1 Design of Variable-Rate, Variable-Length Serially Concatenated Convolutional Codes .... 62
      7.1.1 Problem Formulation and Code Search Technique .............. 64
      7.1.2 Code Search Results ........................ 65

8 Advanced Coding Techniques in Single Carrier IEEE 802.16 standard .... 71
   8.1 Design of the Bit-interleaver ........................ 71
   8.2 Concatenated Channel Code Design .................... 72
   8.3 Simulation Results ............................. 74
9 Coded Polarization Diversity Aspects

9.1 Trellis-Coded Polarization Diversity .............................................. 77
   9.1.1 Decoding Strategies of 4-D TCM for Multiplexing .......................... 79
   9.1.2 Performance ............................................................................ 79
9.2 Joint Space-Time Coding and Polarization Diversity ........................... 84
   9.2.1 Simulation Results .................................................................... 86

10 Channel Estimation and Equalization .................................................. 88

10.1 Channel Models ............................................................................... 88
   10.1.1 Doppler Effect ......................................................................... 89
   10.1.2 Number of path rays to be considered ....................................... 89
10.2 Adaptive Equalisation ...................................................................... 90
   10.2.1 Equaliser Structures .................................................................. 90
   10.2.2 Complex LMS Adaptive Filter .................................................... 93
   10.2.3 Complex Adaptive Kalman Filter ............................................... 93
   10.2.4 AR Equaliser with Kalman-Updating. Analysis and Matlab Simulations . 94
   10.2.5 Programming Issues ................................................................. 94
10.3 Implementation on a TMS320C6711 Digital Signal Processor .......... 100
   10.3.1 Description of the DSP Benchmark ........................................... 100
   10.3.2 C-Program Structure ............................................................... 100
   10.3.3 Achieving Improvement by Using the Optimised DSPLIB Library .... 101
   10.3.4 DSP Performance Results ......................................................... 101

11 Conclusions ....................................................................................... 110

References ............................................................................................ 111
LIST OF FIGURES

1 OFDM sub-carrier schemes. .................................................. 23
2 Simulation model for OFDM simulation. ..................................... 24
3 BER for 256 OFDM signals in AWGN channel. ............................. 25
4 BER for 256 OFDM signals in SUI 4 channel model, with guard interval equal 1/4. ........ 27
5 BER for 256 OFDM signals in SUI 5 channel model, with guard interval equal to 1/4. ........ 27
6 BER for 256 OFDM signals in SUI 5 channel model, with guard interval equal to 1/8. .... 28
7 BER for 256 OFDM signals in SUI 5 channel model, with guard interval equal to 1/8. .... 28
8 BER for 256 OFDM signals in SUI 4 channel model, with guard interval equal to 1/32. .... 29
9 BER for 256 OFDM signals in SUI 5 channel model, with guard interval equal to 1/32. .... 29
10 AM-AM characteristic of HPA. .................................................. 30
11 AM-PM characteristic of HPA. ............................................... 31
12 Simulation model for non-linear distortion analysis. .................. 31
13 Spectrum spreading for QPSK signal in nonlinear channel. .......... 33
14 Spectrum spreading for 16-QAM signal in nonlinear channel. ........ 33
15 Spectrum spreading for 64-QAM signal in nonlinear channel. ......... 34
16 BER of QPSK signal encoded by RS encoder in AWGN channel. ....... 36
17 BER of 16-QAM signal encoded by RS encoder in AWGN channel. .... 37
18 BER of 64-QAM signal encoded by RS encoder in AWGN channel. .... 37
19 BER of QPSK signal encoded by RS and BCC encoder in AWGN channel. ........ 38
20 BER of 16-QAM signal encoded by RS and BCC encoder in AWGN channel. .......... 38
21 BER of 64-QAM signal encoded by RS and BCC encoder in AWGN channel. .......... 39
22 BER for QPSK signal encoded by RS (255,239) encoder in Rice channel. .... 40
23 BER for 16 QAM signal encoded by RS (255,239) encoder in Rice channel. .... 41
24 BER for 64 QAM signal encoded by RS (255,239) encoder in Rice channel. .... 41
25 BER for QPSK signal encoded by RS (255,239) and BCC encoder in Rice channel. ...... 42
26 BER for 16-QAM signal encoded by RS (255,239) and BCC encoder in Rice channel. ..... 42
27 BER for 64-QAM signal encoded by RS (255,239) and BCC encoder in Rice channel. ..... 43
28 Power density spectrum of QPSK signal. ................................ 44
29 Power density spectrum of 16-QAM signal. ................................ 44
30 The model of HAP mobile radio channel. Here the Rice factor $K'$ has been indicated as $c$. 46
31 Received signal power and channel state. ................................. 47
32 Rail track from Koper to Maribor showing terrain height and HAP visibility for HAP altitude 17km. ......................................................... 48
33 Channel attenuation vs. distance. ....................................... 49
Channel attenuation vs. time .................................................. 49
System architecture ............................................................. 51
BER curves of the transmitted coding modulation schemes and switching thresholds... 55
Comparison of bandwidth efficiency for FCM and ACM schemes. ......................... 56
Average BER and average bandwidth efficiency for ACM and FCM schemes without diversity. .................................................. 56
Average BER and average bandwidth efficiency for FCM scheme with diversity. .... 57
Average BER and average bandwidth efficiency for ACM scheme with diversity. .... 58
Average BER and average bandwidth efficiency for ACM scheme with diversity. .... 58
Data rate of adaptive coding and modulation systems. ..................................... 59
An example of System Architecture of the WLAN Transceiver employing variable rate, variable length ACM techniques. .................................................. 61
Block diagram of the variable-length bit-interleaved coded modulation. ................. 61
Block diagram of the SCCC encoder. The overall code rate is $R_S \approx R_o \cdot R_i$ (by neglecting the effects of code terminations), whereby $R_o$ is the rate of the outer code, while $R_i$ is the rate of the inner code. .................................................. 62
Bit error rate performance of the SCCC schemes with rate $R_s = 1/2$. Parameter $\sigma^2_f$ represents the inverse Rice factor, $K = 1/\sigma^2_f$. .................................................. 63
Bit error rate performance of the SCCC schemes with rate $R_s = 2/3$. Parameter $\sigma^2_f$ represents the inverse Rice factor, $K = 1/\sigma^2_f$. .................................................. 64
Bit error rate performance of the SCCC schemes with rate $R_s = 3/4$. Parameter $\sigma^2_f$ represents the inverse Rice factor, $K = 1/\sigma^2_f$. .................................................. 65
Block diagram of the variable-length bit-interleaved coded modulation scheme. The constituent channel codes examined are variable length SCCCs and PCCCs. ................. 71
Example of derivation of the transposition vector for a given permutation. The memory cells are depicted as square boxes. The sequence of transpositions is obtained by simply generating the desired outputs from the input and the contents of the memory cells one after another. .................................................. 72
Block diagram of the serial and parallel concatenated codes. The overall code rate is $R_S$. .................................................. 73
Bit error rate performance of QPSK modulation with variable-length PCCCs and SCCCs. The Rice factor is $K = 10$ dB. .................................................. 75
Bit error rate performance of QPSK modulation with variable-length PCCCs and SCCCs. The Rice factor is $K = 15$ dB. .................................................. 76
Bit error rate performance of QPSK modulation with variable-length PCCCs and SCCCs. The Rice factor is $K = 20$ dB. .................................................. 76
Schematic of a 4-D trellis-coded polarization multiplexing system. ......................... 77
56 BER of the 4-D trellis-coded PM and SM schemes over correlated Rayleigh fading channel \((K = 0)\). .................................................. 80
57 BER of the 4-D trellis-coded PM and SM schemes over Ricean fading channel \((K = 10)\). .................................................. 81
58 Required \(E_b/N_0\) for the BER = \(10^{-3}\) as a function of \(\alpha\), \((K = 0)\). .................................................. 82
59 Required \(E_b/N_0\) for the BER = \(10^{-3}\) as a function of \(\beta\), \((K = 10)\). .................................................. 82
60 Required \(E_b/N_0\) for the BER = \(10^{-3}\) as a function of \(K\)-factor. .................................................. 83
61 Transmitter of the hybrid spatial multiplexing and space-time block codes system based on dual-polarized antennas (PM-STBC). .................................................. 84
62 Receiver of the hybrid spatial multiplexing and space-time block codes system based on dual-polarized antennas. .................................................. 85
63 BER performance of PM-STBC compared to that of STBC over Rayleigh fading channel. .................................................. 87
64 BER performance of PM-STBC compared to that of STBC over Ricean fading channel \((K = 10)\). .................................................. 87
65 Bit error probability for some common modulations. .................................................. 91
66 Structure of a channel estimator + equalizer. .................................................. 92
67 Direct adaptive equaliser. .................................................. 92
68 Adaptive filter structure. .................................................. 93
69 BER for 64-QAM, bit rate 120 Mbit/s, \(f_d = 4500\). Rice factor \(K = 15\) dB (x–x) and \(K = 20\) dB (*–*). Training sequence length: 10000. .................................................. 95
70 Received constellation (64-QAM, bit rate 120 Mbit/s, \(f_d = 4500\). Rice factor \(K = 20\) dB, \(E_b/N_0 = 20\) dB). .................................................. 95
71 Equalised constellation (64-QAM, bit rate 120 Mbit/s, \(f_d = 4500\). Rice factor \(K = 20\) dB, \(E_b/N_0 = 20\) dB). .................................................. 96
72 BER for QPSK, bit rate 32 Mbit/s, \(f_d = 4500\), Rice factor \(K = 10\) dB (x–x) and \(K = 15\) dB (*–*). .................................................. 96
73 Received constellation (QPSK, bit rate 32 Mbit/s, \(f_d = 4500\). Rice factor \(K = 15\) dB, \(E_b/N_0 = 20\) dB). .................................................. 97
74 Equalised constellation (QPSK, bit rate 32 Mbit/s, \(f_d = 4500\). Rice factor \(K = 15\) dB, \(E_b/N_0 = 20\) dB). .................................................. 97
75 BER for QPSK, bit rate 8 Mbit/s, \(f_d = 4500\), Rice factor \(K = 10\) dB (x–x) and \(K = 15\) dB (*–*). .................................................. 98
76 Received constellation (QPSK, bit rate 8 Mbit/s, \(f_d = 4500\). Rice factor \(K = 15\) dB, \(E_b/N_0 = 20\) dB). .................................................. 98
77 Equalised constellation (QPSK, bit rate 8 Mbit/s, \(f_d = 4500\). Rice factor \(K = 15\) dB, \(E_b/N_0 = 20\) dB). .................................................. 99
78 Bench scheme. .................................................. 100
79 Bench components. .................................................. 101
80 Vector operations available in the DSPLIB library. . . . . . . . . . . . . . . . . . . . . . . . . . . . . 102
81 Matrix operations available in the DSPLIB library. . . . . . . . . . . . . . . . . . . . . . . . . . . . . 102
### LIST OF TABLES

1. General QoS requirements for different ATM service classes [27]. CER indicates Cell Error Rate. .................................................. 19
2. Summary of the QoS requirements that the PHY and MAC must provide. Note that delay in the table refers to the transmission delay from the MAC input from the upper layer at the transmit station to the MAC output to the upper layer the receiving station for information transmission. It does not include setup time, link acquisition, etc. .................. 21
3. Main parameters of 256 OFDM transmission scheme. ........................................ 24
4. SUI 4 channel model parameters. .................................................. 26
5. SUI 5 channel model parameters. .................................................. 26
6. HPA model parameters. .................................................. 30
7. Cumulative power in neighbouring frequency bands for 256 OFDM QPSK, 16-QAM and 64-QAM signals. .................................................. 32
8. Standard IEEE 802.16: baud rates and channel sizes for a roll-off factor of 0.25. ...... 35
9. IEEE 802.16 coding schemes defined for 10–66 GHz frequency bands. ................ 35
10. Parameters of the land mobile satellite channel model in Ku band. ..................... 46
11. Scenarios to be considered. .................................................. 52
12. Example of link budget for an HAP-to-train scenario. ...................................... 53
13. Main physical layer parameters for the IEEE 802.16a standard. ....................... 60
14. Theoretical Bit rate, $R_b$, as a function of the SCCC code rate $R_s$ and the system bandwidth $B_w$. .................................................. 62
15. Parameters of the designed SCCCs. The labels $S_1$-$S_6$ identify six different coding schemes whose performances are investigated in this report. .................. 62
16. Optimal puncturing patterns for the convolutional code with generator matrix $G(D) = \begin{bmatrix} 1, & 1 + D^2 \\ 1 + D^2 + D^3 \end{bmatrix}$. .................................................. 66
17. Optimal puncturing patterns for the convolutional code with generator matrix $G(D) = \begin{bmatrix} 1, & 1 + D + D^2 + D^3 \\ 1 + D + D^2 + D^3 \end{bmatrix}$. .................................................. 66
18. Optimal puncturing patterns for the convolutional code with generator matrix $G(D) = \begin{bmatrix} 1, & 1 + D^2 + D^3 \\ 1 + D + D^2 + D^3 \end{bmatrix}$. Symbol * means that the systematic puncturing pattern has a lower minimum SNR with respect to the non-systematic pattern. .................. 67
19. Optimal puncturing patterns for the convolutional code with generator matrix $G(D) = \begin{bmatrix} 1, & 1 + D + D^3 + D^4 \\ 1 + D + D^2 + D^4 \end{bmatrix}$. .................................................. 67
20. Optimal puncturing patterns for the convolutional code with generator matrix $G(D) = \begin{bmatrix} 1, & 1 + D^2 + D^3 + D^4 \\ 1 + D + D^2 + D^4 \end{bmatrix}$. .................................................. 68
21 Optimal puncturing patterns for the convolutional code with generator matrix $G(D) = \begin{bmatrix} 1, \frac{1+D^4}{1+D+D^2+D^3+D^4} \end{bmatrix}$. .................................................. 68

22 Optimal puncturing patterns for the convolutional code with generator matrix $G(D) = \begin{bmatrix} 1, \frac{1+D^2+D^3+D^4}{1+D+D^2+D^3} \end{bmatrix}$. .................................................. 69

23 Optimal puncturing patterns for the convolutional code with generator matrix $G(D) = \begin{bmatrix} 1, \frac{1+D+D^2+D^4}{1+D+D^2+D^3} \end{bmatrix}$. .................................................. 69

24 Optimal puncturing patterns for the convolutional code with generator matrix $G(D) = \begin{bmatrix} 1, \frac{1+D+D^2+D^4}{1+D+D^2+D^3} \end{bmatrix}$. .................................................. 69

25 Optimal puncturing patterns for the convolutional code with generator matrix $G(D) = \begin{bmatrix} 1, \frac{1+D+D^2+D^4}{1+D+D^2+D^3} \end{bmatrix}$. .................................................. 70

26 Puncturing Table for the 8-state recursive constituent encoder with generator matrix $G(D) = \begin{bmatrix} 1, \frac{15}{17} \end{bmatrix}$. .................................................. 74

27 Maximum number of DSP instruction cycles. .................................................. 100

28 Measured instruction cycles. .................................................. 103
LIST OF ACRONYMS

<table>
<thead>
<tr>
<th>Acronym</th>
<th>Description</th>
</tr>
</thead>
<tbody>
<tr>
<td>ABR</td>
<td>Available Bit Rate</td>
</tr>
<tr>
<td>ACM</td>
<td>Adaptive Coding and Modulation</td>
</tr>
<tr>
<td>ADC</td>
<td>Analog-to-Digital Converter</td>
</tr>
<tr>
<td>ALU</td>
<td>Arithmetical and Logical Unit</td>
</tr>
<tr>
<td>AR</td>
<td>AutoRegressive</td>
</tr>
<tr>
<td>ARQ</td>
<td>Automatic Repeat reQuest</td>
</tr>
<tr>
<td>ASDSP</td>
<td>Application Specific Digital Signal Processor</td>
</tr>
<tr>
<td>ATM</td>
<td>Asynchronous Transfer Mode</td>
</tr>
<tr>
<td>AWGN</td>
<td>Average White gaussian Noise</td>
</tr>
<tr>
<td>BE</td>
<td>Best Effort</td>
</tr>
<tr>
<td>BER</td>
<td>Bit Error Rate</td>
</tr>
<tr>
<td>BFWA</td>
<td>Broadband Fixed Wireless Access</td>
</tr>
<tr>
<td>BPSK</td>
<td>Binary Phase Shift Keying</td>
</tr>
<tr>
<td>BWA</td>
<td>Broadband Wireless Access</td>
</tr>
<tr>
<td>CBR</td>
<td>Constant Bit Rate</td>
</tr>
<tr>
<td>CDV</td>
<td>Cell Delay Variation</td>
</tr>
<tr>
<td>CC</td>
<td>Convolutional Code</td>
</tr>
<tr>
<td>CINR</td>
<td>Carrier to Interference and Noise Ratio</td>
</tr>
<tr>
<td>CIR</td>
<td>Channel Impulse Response</td>
</tr>
<tr>
<td>CLP</td>
<td>Cell Loss Probability</td>
</tr>
<tr>
<td>CSI</td>
<td>Channel State Information</td>
</tr>
<tr>
<td>CTD</td>
<td>Cell Transfer Delay</td>
</tr>
<tr>
<td>DAC</td>
<td>Digital-to-Analog Converter</td>
</tr>
<tr>
<td>DFE</td>
<td>Decision Feedback Equaliser</td>
</tr>
<tr>
<td>DSP</td>
<td>Digital Signal Processing</td>
</tr>
<tr>
<td>EGC</td>
<td>Equal Gain Combining</td>
</tr>
<tr>
<td>EMIF</td>
<td>External Memory InterFace</td>
</tr>
<tr>
<td>FCM</td>
<td>Fixed Coding Modulation</td>
</tr>
<tr>
<td>FDD</td>
<td>Frequency Division Duplex</td>
</tr>
<tr>
<td>FEC</td>
<td>Forward Error Correction</td>
</tr>
<tr>
<td>FER</td>
<td>Frame Error Rate</td>
</tr>
<tr>
<td>FIR</td>
<td>Finite Impulse Response</td>
</tr>
<tr>
<td>FPGA</td>
<td>Field Programmable Gate Array</td>
</tr>
<tr>
<td>GEO</td>
<td>GEOstationary (satellite)</td>
</tr>
<tr>
<td>GFR</td>
<td>Guaranteed Frame Rate</td>
</tr>
<tr>
<td>HAP</td>
<td>High Altitude Platform</td>
</tr>
<tr>
<td>HPI</td>
<td>Host-Port Interface</td>
</tr>
<tr>
<td>IE</td>
<td>Inner Encoder</td>
</tr>
<tr>
<td>IF</td>
<td>Intermediate Frequency</td>
</tr>
<tr>
<td>IFFT</td>
<td>Inverse Fast Fourier Transform</td>
</tr>
<tr>
<td>IIR</td>
<td>Infinite Impulse Response</td>
</tr>
<tr>
<td>IPL</td>
<td>InterPlatform Links</td>
</tr>
<tr>
<td>ITU</td>
<td>International Telecommunications Union</td>
</tr>
<tr>
<td>L1D</td>
<td>Level 1 Data cache</td>
</tr>
<tr>
<td>L1P</td>
<td>Level 1 Program cache</td>
</tr>
<tr>
<td>L2</td>
<td>Level 2 memory/cache</td>
</tr>
<tr>
<td>LDC</td>
<td>unit-Lower-triangular-Diagonal Correction</td>
</tr>
<tr>
<td>LMS</td>
<td>Least Mean Square</td>
</tr>
<tr>
<td>LNA</td>
<td>Low-Noise Amplifier</td>
</tr>
<tr>
<td>Abbreviation</td>
<td>Description</td>
</tr>
<tr>
<td>--------------</td>
<td>-------------</td>
</tr>
<tr>
<td>LO</td>
<td>Local Oscillator</td>
</tr>
<tr>
<td>LOS</td>
<td>Line Of Sight</td>
</tr>
<tr>
<td>MAC</td>
<td>Medium Access Control</td>
</tr>
<tr>
<td>MAP</td>
<td>Maximum A Posteriori</td>
</tr>
<tr>
<td>McBSP</td>
<td>Multichannel Buffered Serial Port</td>
</tr>
<tr>
<td>MFLOPS</td>
<td>Million FLOating-Point operations per Second</td>
</tr>
<tr>
<td>MIMO</td>
<td>Multiple Input Multiple Output</td>
</tr>
<tr>
<td>MIPS</td>
<td>Million Integer operation Per Second</td>
</tr>
<tr>
<td>ML</td>
<td>Maximum-Likelihood</td>
</tr>
<tr>
<td>MRC</td>
<td>Maximum Ratio Combining</td>
</tr>
<tr>
<td>MRR</td>
<td>Minimum Reserved Rate</td>
</tr>
<tr>
<td>MSE</td>
<td>Mean Square Error</td>
</tr>
<tr>
<td>MSR</td>
<td>Maximum Sustained Rate</td>
</tr>
<tr>
<td>NLOS</td>
<td>Non Line Of Sight</td>
</tr>
<tr>
<td>nrt</td>
<td>non real time</td>
</tr>
<tr>
<td>nrtPS</td>
<td>Non-Real-Time Polling Services</td>
</tr>
<tr>
<td>OE</td>
<td>Outer Encoder</td>
</tr>
<tr>
<td>PCCC</td>
<td>Parallel Concatenated Convolutional Codes</td>
</tr>
<tr>
<td>PM</td>
<td>Polarization Multiplexing</td>
</tr>
<tr>
<td>PLO</td>
<td>Phase Locked Oscillator</td>
</tr>
<tr>
<td>PRMA</td>
<td>Packet Reservation Multiple Access</td>
</tr>
<tr>
<td>QAM</td>
<td>Quadrature Amplitude Modulation</td>
</tr>
<tr>
<td>QoS</td>
<td>Quality of Service</td>
</tr>
<tr>
<td>OFDM</td>
<td>Orthogonal Frequency Division Multiplexing</td>
</tr>
<tr>
<td>QPSK</td>
<td>Quadrature Phase Shift Keying</td>
</tr>
<tr>
<td>RLS</td>
<td>Recursive Least Square</td>
</tr>
<tr>
<td>RF</td>
<td>Radio Frequency</td>
</tr>
<tr>
<td>RSSI</td>
<td>Received Signal Strength Indicator</td>
</tr>
<tr>
<td>rt</td>
<td>real time</td>
</tr>
<tr>
<td>rtPS</td>
<td>Real-Time Polling Services</td>
</tr>
<tr>
<td>SC</td>
<td>Single Carrier / Switched Combining</td>
</tr>
<tr>
<td>SCCC</td>
<td>Serially Concatenated Convolutional Codes</td>
</tr>
<tr>
<td>SISO</td>
<td>Single-Input Single-Output</td>
</tr>
<tr>
<td>SM</td>
<td>Spatial Multiplexing</td>
</tr>
<tr>
<td>SNR</td>
<td>Signal to Noise Ratio</td>
</tr>
<tr>
<td>SRF</td>
<td>Square Root Filtering</td>
</tr>
<tr>
<td>STBC</td>
<td>Space-Time Block Code</td>
</tr>
<tr>
<td>STC</td>
<td>Space-Time Coding</td>
</tr>
<tr>
<td>TCM</td>
<td>Trellis-Coded Modulation</td>
</tr>
<tr>
<td>UBR</td>
<td>Unspecified Bit Rate</td>
</tr>
<tr>
<td>UGS</td>
<td>Unsolicited Grant Services</td>
</tr>
<tr>
<td>VBR</td>
<td>Variable Bit Rate</td>
</tr>
<tr>
<td>WGS</td>
<td>Weighted Gram-Schmidt</td>
</tr>
<tr>
<td>WLAN</td>
<td>Wireless Local Area Network</td>
</tr>
<tr>
<td>WRC</td>
<td>World Radiocommunications Conference</td>
</tr>
<tr>
<td>XPD</td>
<td>Cross-Polarization Discrimination</td>
</tr>
<tr>
<td>XPIC</td>
<td>Cross-Polarization Interference Canceller</td>
</tr>
</tbody>
</table>
1 Introduction

Broadband wireless access (BWA) is an appealing way to provide flexible and easily-deployable solution for high-speed communications. BWA networks are therefore widely envisioned to have tremendous market potential, because of their distributed installation and semi-ad-hoc routing protocols, that reduce the need for heavy and costly infrastructures [6, 7]. The key requirements of a BWA system include the following [7]:

**High Bandwidth:** high bandwidth reliable links can be shared by multiple users. A control policy that allows the resources to be efficiently switched between users, on a packet-by-packet basis, is needed.

**Low Latency:** latencies of the link should be low. When users get access to the resources, they should be able to sustain high burst rates and transact packets with minimal delay. Mechanisms for achieving link reliability, such as automatic repeat requests (ARQs), should be built, such that the delays are minimized.

**Quality of Service:** QoS provides a mean for effectively partitioning the limited resources of the wireless medium. Users can be provided different levels of service, and service providers can use this to derive revenues.

**Power Savings:** in the presence of battery-operated, portable nodes, mechanisms should exist for seamlessly switching to power save modes in between data bursts while keeping latencies and control overhead in check.

In BWA systems, the wireless link will likely be the bottleneck in any end-to-end mobile wireless broadband system. Therefore, the improvements in wireless link performance, in terms of spectral efficiency, channel utilization and QoS capability, directly translate to the overall improvement of the end-to-end system.

Recently, the provision of Internet access and broadband multimedia services to passengers traveling on-board public means of transport, in particular airplanes and trains, has attracted an increasing interest of the research community, the service providers as well as the telecommunication industry, searching for new market opportunities. The CAPANINA project [8] is aiming at the delivery of broadband services through a collective terminal able to serve all the passengers on board of a train via a high altitude platform (HAP) system. A collective terminal on-board of the train interfaces through an appropriate interworking unit with on-board WLAN base stations. Passengers travelling on the train are equipped with WLAN terminals and access the network via a collective terminal through the HAP and a gateway located in the ground station, requesting broadband multimedia services from various content providers. In this scenario HAPs are used as a collective access network, providing BWA services to the users. Due to their specific characteristics [9–12] HAPs are particularly well suited for such operating scenario. Since they are typically flying in lower stratosphere at altitudes between 17 and 22 km, HAPs can serve relatively large coverage areas from a single quasi-stationary site. These characteristics provide low propagation delays with respect to satellite solutions, broadband capability, small size of antennas and terrestrial terminal equipment, and make HAPs suitable also for the provision of broadcast and multicast services.

In this operating scenario the quality of the received signal depends on the propagation channel between HAP and a terminal on the train. The dominant effects causing the deterioration of the received signal are shadowing, blocking and multipath fading. Shadowing and blocking are caused by obstacles in direct line-of-sight (LOS) propagation path between receiver and transmitter. The impairment caused by shadowing depends on the operating environment. The obstacles around the mobile terminal may reflect the radio signal resulting in multiple copies of the transmitted signal arriving to the receiver with different attenuation and delays, thus causing the multipath fading effect. The total power of reflected signals depends on antenna characteristics of the mobile terminal, the frequency of the transmitted signal and the surrounding environment. Due to the movement of the terminal, the geometry of the surrounding environment changes constantly, which reflects in changing characteristics of the propagation
channel. In order to contrast these varying propagation channel conditions, signal transmission can be optimized by employing adaptive transmission techniques.

The aim of this report is to propose advanced signal processing techniques especially tailored to cope with the time variant response of the HAP-to-train / train-to-HAP wireless communication channel with reference, in particular, to the IEEE 802.16 Standard (both in the Single Carrier and OFDM flavors), which is suited to the HAP environment [35]. In particular, adaptive modulation and coding, highly efficient variable-rate, variable-length channel codes, spatial diversity, polarization diversity coupled with trellis-code modulations and space-time codes, adaptive channel estimation and synchronization based on a Kalman filtering approach are analyzed throughout this report, as some of the most powerful techniques to provide QoS support in HAP-based broadband communications.

Several adaptive transmission techniques, based on the adaptive choice of modulation schemes and coding rates, have been proposed and studied recently to increase the system reliability and throughput in the radio communication systems with time varying channel [13–17]. All these techniques in practice require the channel state information (CSI) to be available at the transmitter to implement the adaptive transmission methods. In frequency division duplex (FDD) systems, CSI can only be estimated at the receiver, and the complete CSI or only the transmission mode, i.e. the modulation and coding scheme and transmitted power, have to be sent back via return channel to the transmitter. Thus, the performance of adaptive rate communication systems is determined by feedback and processing delay, constraints of coding modulation schemes and the rate of channel variation. Since the propagation delay in HAP systems is relatively low, i.e. comparable to delay in terrestrial communication systems, adaptive modulation and coding techniques lend themselves as an efficient way of improving the system performance.

The 47–48 GHz and 28 GHz bands assigned for HAP-based communications, and the consequent high free-space attenuation, suggest inserting high-gain channel codes, coupled with medium-gain adaptive antennas, to replace the use of directive high-gain fixed ground antennas, which are not feasible in inherently moving scenarios. The use of concatenated channel codes (possibly coupled with smart antennas) seem therefore a natural solution, since it can improve the QoS performance at the physical layer, enhancing the received signal level with respect to the received noise and interference. One of the most promising candidates for high-gain coding is represented by the family of the variable-rate, variable-length Serially Concatenated Convolutional Codes (SCCCs). Indeed, they can provide variable rates — which is required in time variant scenarios to adaptively match the propagation conditions, by offering different degrees of protection to the data —, and/or high rate coding for spectral efficiency. They can be optimized, in the sense of minimizing the required SNR for target BER and/or target FER values. For applications requiring high spectral efficiency, high rate codes must satisfy the system requirements in terms of the required BER or FER at a target Signal to Noise Ratio (SNR). High rate punctured Convolutional Codes (CCs) or a suitable concatenation of such codes are among the most commonly used codes for Forward Error Correction (FEC). Puncturing is the most commonly used technique to obtain high-rate CC, since the trellis complexity of the overall code is the same as the lower rate mother code whose output is punctured. Using puncturing, changing the code rate is equivalent to only changing the transmitted symbols and has no impact on the trellis structure of the code used at the receiver for decoding. Since the main complexity of a channel codec resides at the decoder, this solution offers a very flexible and cost effective method of supporting variable rate coding.

If multiple antennas, spatially separated by a few tens of wavelengths, are employed at both transmitter and receiver, a possible alternative is the use of MIMO (Multiple Input Multiple Output) techniques. In fact, digital communication using MIMO systems has recently emerged as one of the most promising technologies for future wireless communication systems. A key characteristic of MIMO systems is the ability to turn multipath propagation, conventionally a limitation of wireless communications, into a benefit for the users. Although recent research has concluded that HAP radio links in K/Ka bands are in LOS for most of the time [30, 36], in particularly dense urban environments multipath can be expected also for the HAP case, with a few hundred ns delay, thus leaving room for the use of space diversity-based signal processing techniques. MIMO based spatial multiplexing (SM) [18] and space-time coding (STC) [19, 20] techniques effectively exploit the multipath, and thus, achieve significant capacity gain and more reliable communication, respectively. However, the performance of these signal processing strategies is strongly dependent on the whole MIMO channel, which is in turn related to the transmit and receive antenna features such as height and spacing and the scattering environment. In practice,
large antenna spacings are normally required at both base stations and receivers, which is undesired and possibly unavailable. The use of co-located dual-polarized antennas is not only a cost-and space-effective substitute, but has the potential to improve system performance for certain environments.

The use of dual-polarized antennas in MIMO systems usually involves exploiting polarization diversity which arises from depolarization, i.e. from the loss of polarization orthogonality of the transmitted signal [21], due to both the anisotropy of the transmission medium and the imperfections of the transmit and receive antennas. One typical application of dual-polarized antennas is in conjunction with spatial multiplexing to achieve higher data rate. It has been argued [22, 23] that for Rayleigh fading channels and in the presence of high transmit correlation, the use of dual-polarized antennas (Polarization Multiplexing, PM) can bring significant performance improvement, but when it comes to the less correlated fading channel, it often leads to a performance loss. Furthermore, in Ricean fading channels with high Rice factor, polarization multiplexing was found to be generally highly superior to spatial multiplexing.

Another application of dual-polarized antennas in MIMO systems is with space-time coding. The combination of spatial multiplexing and space-time coding is getting more attention since it may maximize the average data rate over the MIMO channel and guarantee a minimum order of diversity. In [24], space-time block codes (STBC) [20] have been utilized in such system to provide transmit diversity so that the process of detecting and decoding successive streams or layers is a completely linear process. Furthermore, as shown in [24, 26], the structure of STBC can be further exploited to perfectly suppress the interference from the co-channel terminals while decoding signals from the desired one.

In order to apply all the previously described performance improvement techniques, the receiver must be properly synchronized; moreover, for many of these techniques channel state information (CSI) must be provided, and equalization must be performed. One of the most serious issues in mobile radio links analysis and design is in fact channel estimation and equalization. A very promising approach to these topics is represented by Kalman filter based algorithms, because of their recognized capability to solve the compromise between time-efficiency and performance. Besides, Kalman based equalizers may jointly operate with other digital filtering algorithms, thus allowing simultaneous channel estimation & equalization. Therefore, Kalman filter based algorithms have been investigated in the framework of analyzing implementation aspects of signal processing algorithms developed for software receivers suitable for use on an aerial platform.

Kalman based equalizers are well-known solutions and a lot of work has already been disseminated regarding both theory and practical implementation. However, most of these works address bandwidths lower than those required in HAPs applications. On the other hand, large bandwidths impose new restrictions to Kalman filter programming, mainly related with the possible number of instructions executable by DSP devices between consecutive samples of the received data. There is a compromise between processing speed and algorithm stability (i.e. filter parameter stability), or, in other words, among algorithm optimization, MIPS in the DSP and selection of the most suitable DSP arithmetics (i.e., fixed/floating point algebra). Besides, other implementation issues such as quantization noise, modularization, adaptability, computational efficiency or consumed power are also of capital importance. All these issues affect the cost of HAP user’s terminals and their operative lifetime (in the battery operated case).

The rest of the document is organized as follows. Section 2 discusses the main QoS issues in broadband wireless systems, Section 3 reports a performance evaluation of the standard IEEE 802.16 modulation and coding schemes in different scenarios, to represent the benchmark to compare the performance of different modulation and coding strategies not considered in the Standard. Section 4 recalls the main characteristics of the stratospheric propagation channel, while Section 5 describes the reference scenario and global system architecture addressed throughout the report, and discusses an example of link budget. Sections 6, 7, 8 and 9 are devoted to the detailed analysis of three advanced signal processing approaches suitable to the use in the CAPANINA context: adaptive modulation and coding with and without spatial diversity, adaptive channel coding applied to OFDM and Single Carrier (SC) modulations, coded polarization diversity realized with trellis-coded modulations and space-time block codes. Finally, Section 10 presents a laboratory setup currently in development to test the performance of channel estimation and equalization algorithms in the presence of flat fading effects on the received signal. Finally, Section 11 draws the conclusions.
2 QoS for Broadband Wireless Services

In common wired networks, QoS issues are handled at the network level. The network level QoS requirements (e.g., bandwidth to be reserved, tolerable end-to-end delay and tolerable data loss rate) are directly derived from the transport characterization (e.g., average data rate, sensitivity to real-time skew in data, intermedia/intramedia synchronization) of the multimedia flow [27].

Among the communication standards addressing BWA, nowadays the IEEE 802.16 specification is breaking the QoS "gridlock" that has traditionally plagued broadband wireless equipment designs: 802.16 working group has defined a number of unique QoS parameter that guarantee levels for throughput, latency, and jitter. This enables service providers to offer flexible and enforceable QoS guarantees – a benefit that has never been available with other fixed broadband wireless standards [28].

802.16 working group recognized the value of the ATM service model and incorporated four analogous classes of service, each of which can be tailored to distinct application flows. Then, the IEEE 802.16 standard defined a polling-based MAC layer that is more deterministic than the contention-based MAC used by 802.11. 802.16's MAC layer enables classification of QoS and non-QoS dependant application flows and maps them to connections with distinct scheduling services, enabling both guaranteed handling and traffic enforcement.

In this section we first review the basic QoS mechanisms of an ATM network (Paragraph 2.1), then we see their extension to the IEEE 802.16 MAC layer (Paragraph 2.2). It is worth noticing that, despite the analogy of the ATM's and 802.16's models, the 802.16 MAC layer is not restricted to an ATM transmission mode, but can be applied also to IP.

2.1 QoS in ATM Networks

In ATM (Asynchronous Transfer Mode) networks, the network level QoS requirements are specified by parameters such as cell transfer delay (CTD), cell delay variation (CDV), and cell loss probability (CLP). Indeed, much of ATM's success is directly attributable to its QoS guarantees [28]. In fact, the QoS specifiable Virtual Channels (VCs) provide the ability to meaningfully distinguish the data cells sent over the air and not treat all of them according to some generic policy. Therefore, in a multimedia flow the different traffic streams with different QoS requirements can be transported over different VCs [27].

ATM's attempt to deliver QoS is defined by different classes of service based on constant and variable bit rates. These classes are:

**Constant bit rate (CBR)** In CBR, traffic is characterized by a continuous stream of bits at a steady rate. This class is for low-bandwidth traffic that is highly sensitive to delay and intolerant to cell loss.

**Variable bit rate (VBR)** VBR applies to voice or video applications that use compression. Within this class are real-time VBR (rt-VBR) where real-time end-to-end delivery is critical and non-real time VBR (nrt-VBR), where delay is less critical.

**Best effort services (BE)** These include available bit rate (ABR) and unspecified bit rate (UBR) and apply to LAN traffic that is more tolerant of delays and cell loss.

Since UBR is subject to increased cell and packet loss and does not specify bit rate or traffic parameters, it has no QoS guarantees. ABR is a managed service based on minimum cell rate (MCR) and has a low cell loss. Neither ABR or UBR service classes offer delay variation guarantees. When it came to delivery of real-time services over wireless, however, the QoS challenge was considerable.

For nrt-VBR, ABR, UBR, and guaranteed frame rate (GFR) traffic, CLP is the most important QoS parameter to be guaranteed. If it is assumed that the CDV can be compensated by using buffering and output control at the base station/access point, it gets merged within the CTD issue [27].

For wireless multimedia networks, the general QoS (e.g., loss and delay) requirements for the different components in a multimedia traffic stream can be described as in Table 1, in terms of the different ATM service classes. Nonetheless, general QoS metrics (loss, throughput, delay and delay variation—or jitter—, sequencing, and total errors) have been generally adopted by industry groups for testing QoS over not only ATM networks, but also Frame Relay and IP, for example.
Table 1: General QoS requirements for different ATM service classes [27]. CER indicates Cell Error Rate.

<table>
<thead>
<tr>
<th>Service Class</th>
<th>Applications</th>
<th>Bandwidth</th>
<th>Max. Delay</th>
<th>Target CER</th>
</tr>
</thead>
<tbody>
<tr>
<td>CBR</td>
<td>voice telephony, digital TV</td>
<td>32 kbit/s – 2 Mbit/s</td>
<td>30–40 ms</td>
<td>$&lt; 10^{-2}$</td>
</tr>
<tr>
<td>rt-VBR</td>
<td>real-time video-conferencing</td>
<td>32 kbit/s–2 Mbit/s (avg.)</td>
<td>40–90 ms</td>
<td>$&lt; 10^{-3}$</td>
</tr>
<tr>
<td>nrt-VBR</td>
<td>Multimedia email, digital video</td>
<td>1–10 Mbit/s</td>
<td>unbounded</td>
<td>$&lt; 10^{-6}$</td>
</tr>
<tr>
<td>ABR, GFR</td>
<td>web browsing</td>
<td>1–10 Mbit/s</td>
<td>unbounded</td>
<td>$&lt; 10^{-8}$</td>
</tr>
<tr>
<td>UBR</td>
<td>file transfer</td>
<td>1–10 Mbit/s</td>
<td>unbounded</td>
<td>$&lt; 10^{-8}$</td>
</tr>
</tbody>
</table>

However, the network level end-to-end QoS for multimedia services is dependent on the achieved link level QoS. Guaranteeing QoS means achieving a specified data transmission rate and bit error rate for each connected user in the system.

The issue of QoS, therefore, has become a critical area of concern for suppliers of broadband wireless access (BWA) equipment and their customers. Enforceable QoS is an essential foundation for widespread acceptance of broadband wireless, since it allows for more efficient sharing of the operator’s infrastructure as demand for capacity increases with subscriber take-up [28].

2.2 QoS in IEEE 802.16 Links

In IEEE 802.16 wireless links each connection is associated with a single scheduling data service and each data service is associated with a set of QoS parameters that quantify aspects of its behavior. The 802.16 MAC provides QoS differentiation for different types of applications that might operate over 802.16 networks. The 802.16 standard defines the following types of scheduling services [29]:

**Unsolicited Grant Services (UGS)** UGS is designed to support CBR services, such as T1/E1 emulation, and Voice Over IP (VoIP) without silence suppression.

**Real-Time Polling Services (rtPS)** rtPS is designed to support real-time services that generate variable size data packets on a periodic basis, such as MPEG video or VoIP with silence suppression.

**Non-Real-Time Polling Services (nrtPS)** nrtPS is designed to support non-real-time services that require variable size data grant burst types on a regular basis.

**Best Effort (BE) Services** BE services are typically provided by the Internet today for Web surfing.

Each subscriber station-to-base station connection is assigned a service class as part of the creation of the connection. When packets are classified in the convergence sublayer, the connection into which they are placed is chosen based on the type of QoS guarantees that are required by the application [29].

Furthermore, there are two types of polling mechanisms:

**Unicast** When a subscriber station (SS) is polled individually, it is allocated bandwidth to send bandwidth request messages.

**Contention-based** Contention-based bandwidth request is used when insufficient bandwidth is available to individually poll many inactive SS’s. The allocation is multicast or broadcast to a group of SS’s that have to contend for the opportunity to send bandwidth requests.

The key QoS metrics associated with each of the above four scheduling services are:

- the maximum sustained rate (MSR),
- the minimum reserved rate (MRR),
- the maximum latency,
• the maximum jitter, and
• the priority.

The MRR is associated with different scheduling services, acting as the "guarantee", while the MSR serves to rate limit a connection or "enforcement" of a maximum sustained rate. MSR, as defined by the 802.16 standard, is the peak information rate of the service measured in bits per second. The service must be policed to conform to this parameter on the average over time on the wireless link. No additional policing is required at the base station in the downlink direction. MRR specifies the minimum rate reserved for service flow in bits per second and specifies the minimum amount of data to be transported on behalf of the service flow when averaged over time. In other words, the MRR is guaranteed, while MSR is provided on a best-effort basis if the network has the resources. Service providers will typically plan their networks such that the sum of all MRRs does not exceed the total capacity of the networks. Best effort services and MSRs are therefore accommodated from the excess capacity of the network beyond the sum of all MRR guarantees.

Specifications set out in 802.16 outline a number of key components that cooperate to guarantee QoS over the wireless interface. These range from MAC layer control protocols for authentication and service flows establishment, to PHY layer mechanisms that maintain a low bit error rates (BER) of $10^{-6}$ through dynamic modulation and coding techniques [28].

**PHY Mechanisms**

A key attribute of 802.16-based systems utilizing time division duplex (TDD) radios is their more efficient use of spectrum. The 802.16a-OFDM, for example, achieves its high data rate and efficiency by using multiple orthogonal (overlapping) carrier signals (OFDM). This parallel carrier ability is called multi-carrier modulation (MCM) or discrete multi-tone (DMT), and is ideal for addressing errors that may arise in indoor and outdoor wireless environments, in particular up to the X band [30].

*Spectral efficiency* is a measure of the number of bits that can be carried by the channel. It is specified in bits per Hz either "over the air" or as a net useable Ethernet rate. A system spectral efficiency is therefore directly related to total available bandwidth. A system that has high efficiency will deliver higher throughputs and therefore support a greater number of total subscribers.

Finally, *dynamic modulation and coding* is a particularly important feature of the 802.16 standard in allowing providers to maintain prescribed QoS levels. Unlike copper circuits, which tend to maintain a constant bit error rate, wireless links have BERs that are fluid due to environmental conditions (multipath, fast fade, refraction etc.). The 802.16 standard employs dynamic control of modulation and coding to mitigate the environmental effects and maintain a constant BER. Dynamic modulation and coding parameters are defined by the carrier to interference and noise ratio (CINR). This represents a measure of quality for wireless signals. The higher the value of the CINR, the more throughput a link can maintain. When a signal degrades, the CINR value decreases. The 802.16 standard adapts to signal degradation by dropping to a lower modulation, thereby trading BER for throughput. The first step is to drop the coding from 3/4 to 1/2. Then, modulation can continue to drop down to maintain a constant BER of $10^{-6}$ which represents a common baseline.

While *retransmitting traffic* for data services does not represent a significant problem (data can accommodate high latency), it is critical that real-time applications avoid the need for retransmission. From a real-time perspective, it is far preferable to sustain errors or loss for data, since this does not effect delay and jitter. Real-time services, on the other hand, cannot tolerate these problems and therefore a sustained BER is critical to providing acceptable QoS.

Document [65], by IEEE 802.16.1 System Requirements Task Group, provides functional requirements that are guidelines for developing an interoperable 802.16.1 air interface and contains the QoS requirements for an 802.16 radio link in 10-66 GHz frequency bands. This document summarizes the main requirements at PHY/MAC layer in Table 2. This has been the basis for selecting the reference scenarios that will be discussed in Section 5.

In a context such as that addressed to CAPANINA applications, including fast mobility, high data rate and multimedia applications over bi-directional links, it is evident that providing QoS-guaranteed services is a critical issues. Therefore, highly efficient schemes of adaptive modulation and coding, possibly coupled with diversity solutions able to increase the overall system reliability and spectral efficiency, have to be identified and tuned to the specific CAPANINA applications. Not only, reliable,
<table>
<thead>
<tr>
<th>Bearer Service</th>
<th>MAC Payload Rate</th>
<th>Max Rate</th>
<th>Max Delay (One way)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Circuit-Based</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>High Quality Narrowband/Voice Frequency Telephony (Vocoder MOS $\geq 4.0$)</td>
<td>32 Kbps - 64 Kbps</td>
<td>BER $10^{-5}$</td>
<td>5 ms</td>
</tr>
<tr>
<td>High Quality Narrowband/Voice Frequency Telephony (Vocoder MOS $&lt; 4.0$)</td>
<td>6 Kbps - 16 Kbps</td>
<td>BER $10^{-4}$</td>
<td>10 ms</td>
</tr>
<tr>
<td>Trunking</td>
<td>$\leq 155$ Mbps</td>
<td>BER $10^{-4}$</td>
<td>5 ms</td>
</tr>
<tr>
<td>Variable Packet</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>Time Critical Packet Services</td>
<td>4 - 13 Kbps (voice) and 32 - 1.5 Mbps (video)</td>
<td>BER $10^{-4}$</td>
<td>10 ms</td>
</tr>
<tr>
<td>Non-Time Critical Services: IP, IPX, FR... Audio/video streaming, Bulk data transfer, etc...</td>
<td>$\leq 155$ Mbps</td>
<td>$\leq 155$ Mbps</td>
<td>N/A</td>
</tr>
<tr>
<td>MPEG video</td>
<td>$\leq 8$ Mbps</td>
<td>BER $10^{-11}$</td>
<td>TBD</td>
</tr>
<tr>
<td>Fixed-length Cell/Packet</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>ATM Cell Relay - CBR</td>
<td>16 Kbps - 155 Mbps</td>
<td>CLR $3 \cdot 10^{-7}$</td>
<td>10 ms</td>
</tr>
<tr>
<td></td>
<td></td>
<td>CER $4 \cdot 10^{-6}$</td>
<td></td>
</tr>
<tr>
<td></td>
<td></td>
<td>CMR $1$/day</td>
<td></td>
</tr>
<tr>
<td></td>
<td></td>
<td>SECBR $10^{-4}$</td>
<td></td>
</tr>
<tr>
<td>ATM Cell Relay - rt - VBR</td>
<td>Same as CBR above</td>
<td>CLR $10^{-5}$</td>
<td>10 ms</td>
</tr>
<tr>
<td></td>
<td></td>
<td>CER $4 \cdot 10^{-6}$</td>
<td></td>
</tr>
<tr>
<td></td>
<td></td>
<td>CMR $1$/day</td>
<td></td>
</tr>
<tr>
<td></td>
<td></td>
<td>SECBR $10^{-4}$</td>
<td></td>
</tr>
<tr>
<td>ATM Cell Relay - other</td>
<td>$\leq 155$ Mbps</td>
<td>CLR $10^{-5}$</td>
<td>N.A.</td>
</tr>
<tr>
<td></td>
<td></td>
<td>CER $4 \cdot 10^{-6}$</td>
<td></td>
</tr>
<tr>
<td></td>
<td></td>
<td>CMR $1$/day</td>
<td></td>
</tr>
<tr>
<td></td>
<td></td>
<td>SECBR $10^{-4}$</td>
<td></td>
</tr>
</tbody>
</table>

Table 2: Summary of the QoS requirements that the PHY and MAC must provide. Note that delay in the table refers to the transmission delay from the MAC input from the upper layer at the transmit station to the MAC output to the upper layer the receiving station for information transmission. It does not include setup time, link acquisition, etc.
fast, adaptive channel estimation and synchronization algorithms are needed, in order to cope with the fast mobility and variability of the addressed scenarios. This is specifically the goal of the work reported in this document.
3 Reference System Performance for IEEE 802.16 Standard

Work-package 2.1 selected IEEE 802.16-SC (Single Carrier) as the reference standard for broadband access to the CAPANINA communications services, due to the fact that no significant multipath effect is expected in the 28/31 GHz band, and SC modulations seem to be slightly more robust to non-linear distortions produced by power amplifiers [35]. However, at the early stage of the work on WP2.3, this kind of information was not yet available, so that also OFDM formats were considered as potential candidates for the physical layer. For this reason, we present in this report an analysis of advanced coding and modulation techniques applied to both OFDM and SC schemes.

In this section in particular we analyze the performance of some modulation and coding formats proposed in the selected standard, in order to provide a common benchmark for the advanced signal processing techniques investigated in the next sections.

Two sets of frequency bands, namely 2-11 GHz and 10-66 GHz are distinguished in IEEE 802.16 standard. Lower frequency bands are selected for providing wireless access in urban, suburban and rural environment with no line of sight conditions, thus for harsh multipath environment. For that reason the IEEE standard proposes the robust modulation scheme to cope with the channel impairments: trellis coded single carrier modulation, multicarrier modulation based on 256 OFDM signal and orthogonal frequency division multiple access system with 1024 subcarriers. While, in upper frequency bands the line of sight is expected without multipath distortions, with possible time dependant attenuation due to meteorological effects like severe rain and with possible shadowing and blocking of the signal in the case of mobile communications. Therefore, only single carrier communication is proposed in upper frequency band. On the other hand, 256 OFDM is the standard selected in WiMAX for lower frequency bands, then we focus our analysis to standard OFDM transmission scheme in this case. An additive white Gaussian noise channel and various multipath communication channels are used as a reference.

3.1 System performance for 256 OFDM transmission scheme

Two aspects of 256 OFDM signal performance are analysed in the next section, the system performance (BER) in AWGN channel and multi-path environment, and OFDM signal distortion due to non-linear high power amplifier.

3.1.1 System performance evaluation in AWGN and multi-path environment

The 256 OFDM transmission scheme consists of 256 subcarriers. Among them 192 subcarriers carry information, 27 upper carriers and 28 lower carriers are used as a frequency guard band, 8 subcarriers transmit the pilot tones, while DC carrier is unused. The structure of 256 OFDM is shown in Figure 1. The main parameters of 256 OFDM transmission scheme are shown in Table 3. The three types of encoding are proposed in standard:

- code type 1: concatenated coding with Reed Solomon (RS) outer encoder and convolutional encoder as the inner code,
- code type 2: concatenated coding with Reed Solomon outer encoder and parity bit check code as the inner code
### Table 3: Main parameters of 256 OFDM transmission scheme.

<table>
<thead>
<tr>
<th>PARAMETER</th>
<th>VALUE</th>
</tr>
</thead>
<tbody>
<tr>
<td>$N_{FFT}$ – number of carriers</td>
<td>256</td>
</tr>
<tr>
<td>$N_{used}$ – number of used carriers</td>
<td>200</td>
</tr>
<tr>
<td>$f_s/BW$ – sampling frequency ratio</td>
<td>For bandwidths n*1.75 MHz: 8/7 For other bandwidths: 7/6</td>
</tr>
<tr>
<td>$T_g/T_b$ – guard time ratio</td>
<td>1/4, 1/8, 1/16, 1/32</td>
</tr>
<tr>
<td>Guard region subcarriers</td>
<td>-128, -127, ..., -101, 101, 102, ..., 127</td>
</tr>
<tr>
<td>Pilot tones</td>
<td>-84, -60, -36, -12, 12, 36, 60, 84</td>
</tr>
</tbody>
</table>

The concatenated RS with convolutional inner encoder is mandatory, while other two codes are optional. For that reason we perform simulations only for code type 1.

The simulation model is shown in Figure 2. The information data are encoded with inner and outer encoder and modulated by OFDM modulator. The channel model developed at Stanford University, also known as SUI channel, introduces the multipath distortion to the signal [66, 67]. In the absence—at the early stage of this work—of suitable multipath models for the HAP channel, the SUI models, intended for terrestrial systems, provided an estimate of performance on the worst-case HAP channel, where the elevation of the platform is very low, and hence the radio path is nearly horizontal at the terminal. After that the Gaussian noise is added. FEC decoding, using hard decoding algorithm, is followed by the OFDM demodulator. The BER is calculated comparing transmitted and received bit stream. The outer encoder is RS (255,239) and the inner encoder is convolutional encoder with coding rate 1/2 as specified in IEEE 802.16 standard.

First the results are obtained for AWGN channel. The results for QPSK, 16-QAM and 64-QAM are plotted in Figure 3, for encoder rate 1/2 and guard interval 1/4. Less than 4dB additional energy per transmitted bit is required to achieve the same BER for 16-QAM signal comparing to QPSK signal. As expected, 64-QAM signal requires the highest energy per bit to achieve the same BER. For example, to achieve BER=0.00001 the required Eb/No ratios are 6.2dB, 9.7dB and 13.5dB for QPSK, 16-QAM and 64-QAM respectively.

Various channel models have been developed, which attempt to characterise the RF multipath environment for frequency bands between 2 and 11 GHz. Among them the Stanford University Interim (SUI) channel models, which are the extension of AT&T channel models, are widely used in simulating the performance of IEEE 802.16 system in multipath environment. Three basic terrain types are proposed in SUI models:

- Category A: Hilly and moderate to heavy tree density terrain,
- Category B: Hilly and light tree density or flat and moderate to heavy tree density and
- Category C: Flat and light tree density.

The SUI channel models have been selected for design, development and testing of WiMAX technology in six different scenarios, (SUI 1 to SUI 6). For the reference simulation, we have selected two models,
describing the harsh multipath environment, denoted by SUI 4 and SUI 5. The parameters of SUI 4 and SUI 5 channel models are shown in Table 4 and Table 5.

The system performance of the 256 OFDM signal is plotted in Figures 4–9, for two channel models, SUI 4 and SUI 5, three modulation schemes (QPSK, 16 QAM and 64 QAM) and four guard intervals (1/4, 1/8, 1/32). There is no system performance degradation, when the guard interval is comparable to the delay of the reflected signal. However, when the reflected signal is delayed more than the guard interval, the significant system performance degradation is observed. In SUI 5 channel model, the signal exhibits stronger distortion and larger multipath delays, and consequently the BER curves for low guard intervals namely 1/32 and 1/8, do not follow the waterfall curve. Instead an error floor occurs depending on modulation scheme and guard interval. The error floor can be extremely high, for some cases above $10^{-3}$, which makes the system unusable. The perfect channel state information has been used to detect the OFDM signal. The results would be slightly different, if an error in channel state information were introduced in simulations.
### Channel SUI - 4

<table>
<thead>
<tr>
<th></th>
<th>Tap 1</th>
<th>Tap 2</th>
<th>Tap 3</th>
<th>Units</th>
</tr>
</thead>
<tbody>
<tr>
<td>Delay</td>
<td>0</td>
<td>1.5</td>
<td>4</td>
<td>μs</td>
</tr>
<tr>
<td>Power (omni.ant.)</td>
<td>0</td>
<td>0</td>
<td>0</td>
<td>dB</td>
</tr>
<tr>
<td>90% K-factor (omni)</td>
<td>0</td>
<td>0</td>
<td>0</td>
<td>dB</td>
</tr>
<tr>
<td>75% K-factor (omni)</td>
<td>0</td>
<td>0</td>
<td>0</td>
<td>dB</td>
</tr>
<tr>
<td>Power (30° ant.)</td>
<td>0</td>
<td>-10</td>
<td>-20</td>
<td>dB</td>
</tr>
<tr>
<td>90% K-factor (30°)</td>
<td>1</td>
<td>0</td>
<td>0</td>
<td>dB</td>
</tr>
<tr>
<td>75% K-factor (30°)</td>
<td>5</td>
<td>0</td>
<td>0</td>
<td>dB</td>
</tr>
<tr>
<td>Doppler</td>
<td>0,2</td>
<td>0,15</td>
<td>0,25</td>
<td>Hz</td>
</tr>
</tbody>
</table>

**Antenna Correlation:** \( \rho_{\text{ENV}} = 0.3 \)

**Gain Reduction Factor:** \( \text{GRF} = 4\text{dB} \)

**Normalization Factor:** \( F_{\text{OMNI}} = -1,9218\text{dB} \)

\( F_{\text{30°}} = -0,4532\text{dB} \)

**Terrain Type:** B

**Omni antenna:** \( \tau_{\text{RMS}} = 1,257\mu\text{s} \)
overall K: \( K = 0,2 \) (90%); \( K = 0,6 \) (75%)

**30° antenna:** \( \tau_{\text{RMS}} = 0,563\mu\text{s} \)
overall K: \( K = 1,0 \) (90%); \( K = 3,2 \) (75%)

<table>
<thead>
<tr>
<th></th>
<th>Tap 1</th>
<th>Tap 2</th>
<th>Tap 3</th>
<th>Units</th>
</tr>
</thead>
<tbody>
<tr>
<td>Delay</td>
<td>0</td>
<td>4</td>
<td>10</td>
<td>μs</td>
</tr>
<tr>
<td>Power (omni.ant.)</td>
<td>0</td>
<td>-5</td>
<td>-10</td>
<td>dB</td>
</tr>
<tr>
<td>90% K-factor (omni)</td>
<td>0</td>
<td>0</td>
<td>0</td>
<td>dB</td>
</tr>
<tr>
<td>75% K-factor (omni)</td>
<td>0</td>
<td>0</td>
<td>0</td>
<td>dB</td>
</tr>
<tr>
<td>Power (30° ant.)</td>
<td>0</td>
<td>-11</td>
<td>-22</td>
<td>dB</td>
</tr>
<tr>
<td>90% K-factor (30°)</td>
<td>1</td>
<td>0</td>
<td>0</td>
<td>dB</td>
</tr>
<tr>
<td>75% K-factor (30°)</td>
<td>5</td>
<td>0</td>
<td>0</td>
<td>dB</td>
</tr>
<tr>
<td>Doppler</td>
<td>2</td>
<td>1,5</td>
<td>2,5</td>
<td>Hz</td>
</tr>
</tbody>
</table>

**Antenna Correlation:** \( \rho_{\text{ENV}} = 0.3 \)

**Gain Reduction Factor:** \( \text{GRF} = 4\text{dB} \)

**Normalization Factor:** \( F_{\text{OMNI}} = -1,5113\text{dB} \)

\( F_{\text{30°}} = -0,3573\text{dB} \)

**Terrain Type:** A

**Omni antenna:** \( \tau_{\text{RMS}} = 2,842\mu\text{s} \)
overall K: \( K = 0,1 \) (90%); \( K = 0,3 \) (75%); \( K = 1,0 \) (50%)

**30° antenna:** \( \tau_{\text{RMS}} = 1,276\mu\text{s} \)
overall K: \( K = 0,4 \) (90%); \( K = 1,3 \) (75%); \( K = 4,2 \) (50%)

Table 4: SUI 4 channel model parameters.

Table 5: SUI 5 channel model parameters.
Figure 4: BER for 256 OFDM signals in SUI 4 channel model, with guard interval equal 1/4.

Figure 5: BER for 256 OFDM signals in SUI 5 channel model, with guard interval equal to 1/4.
Figure 6: BER for 256 OFDM signals in SUI 5 channel model, with guard interval equal to 1/8.

Figure 7: BER for 256 OFDM signals in SUI 5 channel model, with guard interval equal to 1/8.
Figure 8: BER for 256 OFDM signals in SUI 4 channel model, with guard interval equal to 1/32.

Figure 9: BER for 256 OFDM signals in SUI 5 channel model, with guard interval equal to 1/32.
3.1.2 Spectrum Spreading of the 256 OFDM Signals in Non-Linear Channel

The OFDM signals are sensitive to non-linear distortion mainly introduced by high power amplifier in communication circuits. In the following subsection we calculate the spectrum spreading in neighbouring frequency bands for different operating point of nonlinear amplifier.

The model of non-linear amplifier proposed in [63] is used to simulate the signal properties. The AM/AM distortion of nonlinear amplifier can be described as:

\[ A(r) = a \left( 1 - \exp (-br) \right) + cr \exp (-dr^2), \]  

(1)

where \( r \) is the magnitude of the signal at the input of the amplifier and \( a, b, c \) are the coefficients of the model. The AM-PM characteristic of the amplifier can be represented by the following exponential function:

\[ \Phi(r) = f \left( 1 - e^{-g(r-h)} \right) \]  

(2)

where \( \Phi(r) \) is the phase shift of the output signal in radians and the parameters define the following: \( f \) is magnification factor, \( g \) is steepness of the curve and \( h \) is shift of the curve in \( r \)-axis.

Different sets of coefficients for AM-AM and AM-PM characteristics are found for different batches of amplifiers. Two sets, describing representative samples, are presented in [63] and recapitulated in Table 6. The AM-AM and AM-PM characteristics of the HP A are plotted in Figures 10 and 11, respectively. The markers represent the measured data, while the lines describe the model of HP A. HP A amplifies higher magnitudes less than smaller ones, which causes squeezing of the scattering diagram. The higher signal magnitudes are phase shifted more than smaller, which is why the scattering diagram is twisted.

The simulation model is plotted in Figure 12. The random data source is modulated in OFDM modulator. The signal spectrum is calculated before and after non-linear distortion in order to estimate the level of distortion caused by high power amplifier. The results are plotted in Figures 13–15. The black solid or dashed lines in denote the linear amplification of OFDM signal. Other curves are obtained for...
different operating points of HPA. Even with high back-off from saturation (-8dB) the OFDM signal exhibits high spectrum spreading in neighbouring frequency bands due their high peak to average power ratio (PAPR). Backing off the operating point of HPA causes the communication system energetically inefficient, thus inappropriate for applying in the systems with limited power resources, which is particularly the case in the high altitude platforms. No major differences in different modulation schemes (QPSK, 16-QAM and 64-QAM) are observed from simulation results. In order to determine the co-channel interference in the first, the second and the third neighbouring frequency band, the cumulative power in neighbouring bands due to non-linear distortion is calculated. The results are collected in Table 5. Again, no major differences are observed between different modulation schemes (QPSK, 16-QAM and 64-QAM). Moving the operating point closer to the saturation causes an increase in the amount of power spreading to neighbouring frequency bands, as expected. Frequency bands located far from the band where the signal is transmitted are less affected by signal power spreading.

The simulation results show significant power spreading to neighbouring frequency bands when non-linear distortion caused by the high power amplifier appears in the system. A straightforward solution, backing off the operating point of high power amplifier significantly, is not suitable due to the limited power resources at the HAP. Techniques such as feedforward and feedback linearisation, synthesis techniques (LINC and CALLUM), predistortion and envelope elimination and restoration will be required to linearize nonlinear characteristics of the high power amplifiers. In addition the techniques developed to minimize peak to average power ration of the OFDM signals such as signal coding, which exclude sub-channel data combinations giving rise to large peaks, using only a sub-set of the sub-channels, truncating the peaks in signals or adding a bandlimited signal to cancel the amplitude peaks, might be considered. These techniques will reduce the PAPR on the one hand, but on other they may also reduce the system throughput and increase the system overhead and in-band self interference. However the results given here show that such techniques would be essential if OFDM were used in a HAP system.
<table>
<thead>
<tr>
<th></th>
<th></th>
<th></th>
<th></th>
<th></th>
<th></th>
</tr>
</thead>
<tbody>
<tr>
<td>-8</td>
<td>-35.3</td>
<td>-48.9</td>
<td>-59.5</td>
<td>8.6</td>
<td>0.2</td>
</tr>
<tr>
<td>-6</td>
<td>-35.0</td>
<td>-48.4</td>
<td>-57.9</td>
<td>6.4</td>
<td>0.3</td>
</tr>
<tr>
<td>-4</td>
<td>-33.1</td>
<td>-48.7</td>
<td>-56.7</td>
<td>4.5</td>
<td>0.4</td>
</tr>
<tr>
<td>-2</td>
<td>-30.1</td>
<td>-48.9</td>
<td>-55.8</td>
<td>3.2</td>
<td>0.6</td>
</tr>
<tr>
<td>0</td>
<td>-27.1</td>
<td>-47.7</td>
<td>-55.5</td>
<td>2.3</td>
<td>0.8</td>
</tr>
<tr>
<td>2</td>
<td>-24.4</td>
<td>-45.7</td>
<td>-54.8</td>
<td>1.8</td>
<td>1.0</td>
</tr>
<tr>
<td>4</td>
<td>-22.1</td>
<td>-43.2</td>
<td>-52.9</td>
<td>1.5</td>
<td>1.2</td>
</tr>
<tr>
<td>Linear</td>
<td>-47.2</td>
<td>-63.7</td>
<td>-68.2</td>
<td>11.5</td>
<td>1.0</td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th></th>
<th></th>
<th></th>
<th></th>
<th></th>
<th></th>
</tr>
</thead>
<tbody>
<tr>
<td>-8</td>
<td>-35.3</td>
<td>-48.7</td>
<td>-57.4</td>
<td>8.4</td>
<td>0.2</td>
</tr>
<tr>
<td>-6</td>
<td>-35.2</td>
<td>-48.2</td>
<td>-56.6</td>
<td>6.3</td>
<td>0.3</td>
</tr>
<tr>
<td>-4</td>
<td>-33.4</td>
<td>-48.3</td>
<td>-55.7</td>
<td>4.5</td>
<td>0.4</td>
</tr>
<tr>
<td>-2</td>
<td>-30.3</td>
<td>-48.3</td>
<td>-54.9</td>
<td>3.2</td>
<td>0.6</td>
</tr>
<tr>
<td>0</td>
<td>-27.2</td>
<td>-47.4</td>
<td>-54.3</td>
<td>2.3</td>
<td>0.8</td>
</tr>
<tr>
<td>2</td>
<td>-24.5</td>
<td>-45.4</td>
<td>-53.5</td>
<td>1.8</td>
<td>1.0</td>
</tr>
<tr>
<td>4</td>
<td>-22.2</td>
<td>-42.8</td>
<td>-51.9</td>
<td>1.5</td>
<td>1.2</td>
</tr>
<tr>
<td>Linear</td>
<td>-46.6</td>
<td>-57.5</td>
<td>-60.1</td>
<td>10.7</td>
<td>1.0</td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th></th>
<th></th>
<th></th>
<th></th>
<th></th>
<th></th>
</tr>
</thead>
<tbody>
<tr>
<td>-8</td>
<td>-34.5</td>
<td>-47.8</td>
<td>-55.4</td>
<td>8.6</td>
<td>0.2</td>
</tr>
<tr>
<td>-6</td>
<td>-34.3</td>
<td>-47.4</td>
<td>-54.8</td>
<td>6.4</td>
<td>0.3</td>
</tr>
<tr>
<td>-4</td>
<td>-32.6</td>
<td>-47.5</td>
<td>-54.1</td>
<td>4.5</td>
<td>0.4</td>
</tr>
<tr>
<td>-2</td>
<td>-29.7</td>
<td>-47.4</td>
<td>-53.4</td>
<td>3.2</td>
<td>0.6</td>
</tr>
<tr>
<td>0</td>
<td>-26.8</td>
<td>-46.4</td>
<td>-52.8</td>
<td>2.3</td>
<td>0.8</td>
</tr>
<tr>
<td>2</td>
<td>-24.1</td>
<td>-44.7</td>
<td>-52.1</td>
<td>1.8</td>
<td>1.0</td>
</tr>
<tr>
<td>4</td>
<td>-21.8</td>
<td>-42.5</td>
<td>-50.8</td>
<td>1.5</td>
<td>1.2</td>
</tr>
<tr>
<td>Linear</td>
<td>-43.2</td>
<td>-54.0</td>
<td>-56.7</td>
<td>11.3</td>
<td>1.0</td>
</tr>
</tbody>
</table>

Table 7: Cumulative power in neighbouring frequency bands for 256 OFDM QPSK, 16-QAM and 64-QAM signals.
Figure 13: Spectrum spreading for QPSK signal in nonlinear channel.

Figure 14: Spectrum spreading for 16-QAM signal in nonlinear channel.
Figure 15: Spectrum spreading for 64-QAM signal in nonlinear channel.
### Table 8: Standard IEEE 802.16: baud rates and channel sizes for a roll-off factor of 0.25.

<table>
<thead>
<tr>
<th>Channel size (MHz)</th>
<th>Symbol rate (MBd)</th>
<th>Bit rate (Mb/s) QPSK</th>
<th>Bit rate (Mb/s) 16-QAM</th>
<th>Bit rate (Mb/s) 64-QAM</th>
<th>Recommended Frame duration (ms)</th>
<th>Number of PSs/frame</th>
</tr>
</thead>
<tbody>
<tr>
<td>20</td>
<td>16</td>
<td>32</td>
<td>64</td>
<td>96</td>
<td>1</td>
<td>4000</td>
</tr>
<tr>
<td>25</td>
<td>20</td>
<td>40</td>
<td>80</td>
<td>120</td>
<td>1</td>
<td>5000</td>
</tr>
<tr>
<td>28</td>
<td>22.4</td>
<td>44.8</td>
<td>89.6</td>
<td>134.4</td>
<td>1</td>
<td>5600</td>
</tr>
</tbody>
</table>

**Table 9: IEEE 802.16 coding schemes defined for 10–66 GHz frequency bands.**

<table>
<thead>
<tr>
<th>Code type</th>
<th>Outer code</th>
<th>Inner code</th>
<th>Usage</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>RS over GF(256)</td>
<td>–</td>
<td>Able to correct up to 16 bytes of errors per block. Suitable for large data blocks, or when high data rate is required.</td>
</tr>
<tr>
<td>2</td>
<td>RS over GF(256) (same as type 1)</td>
<td>BCC(24,16)</td>
<td>Useful for low to moderate coding rates.</td>
</tr>
<tr>
<td>3</td>
<td>RS over GF(256) (same as type 1)</td>
<td>parity check code</td>
<td>Optional.</td>
</tr>
<tr>
<td>4</td>
<td>Turbo code</td>
<td>–</td>
<td>Optional.</td>
</tr>
</tbody>
</table>

### 3.2 System Performance for Frequency Bands from 10 to 66 GHz

The single carrier modulations are proposed in IEEE 802.16 standard for the frequency bands from 10–66 GHz. Such a system use Nyquist square-root raised cosine pulse shaping with a roll-off factor of 0.25 and operate on the default RF channel arrangement shown in Table 8.

The standard defines four coding schemes, summarized in Table 9. In code type 1, the outer code is a systematic Reed Solomon code over GF(256) and no inner code is used. The block length is 255 bytes, and the information block length varies from 6 to 255 bytes. The code is able to correct up to 16 bytes of errors in a block. The code is suitable for large data blocks, or when high data rate is required. In code type 2 the outer code is the same RS code as in code type 1. The inner code is a (24,16) block convolutional code (BCC) and it can be considered as an equivalent of an unsystematic block code. Code type 2 is useful for low to moderate coding rates. In code type 3 the block convolutional encoder from code type 2 is replaced by simple parity check code. In code type 4 the block turbo code with no inner code is used. The codes type 1 and type 2 are mandatory for implementation, while the codes type 3 and type 4 are optional. In the following, we limit our analysis only to the mandatory code types.

#### 3.2.1 System Performance Evaluation of Single Carrier Modulation in AWGN Channel

The system performance for the code type 1 is simulated at first for QPSK, 16-QAM and 64-QAM modulated signals. The following Reed Solomon codes are chosen: RS (255,239), RS (255,247), RS (255,251) and RS (255,253). The results are plotted in Figure 16, Figure 17 and Figure 18. The solid line in figures denotes the modulation performance without encoding. An increase of system performance by approximately 1dB is observed, when 1, 2, 4 and 8 bytes can be corrected by RS encoder.

The code type 2, using block convolutional encoder as inner and RS encoder as outer encoder, is tested in the second set of simulations. The output of convolutional encoder is punctured as defined in the standard by the ratio 3/4 thus the coding rate of the convolutional encoder is increased to 2/3. The results are plotted in Figure 19, Figure 20 and Figure 21. The convolutional encoder significantly decreases the BER. Approximately 3dB less energy per transmitted bit is required for the same BER in the case when the convolutional encoder is added to the system. When the coding rate of RS encoder...
Figure 16: BER of QPSK signal encoded by RS encoder in AWGN channel.

is decreased from (255,253) to (255,239) approximately 2 dB less power is required to obtain the same BER for all tested modulation schemes.
Figure 17: BER of 16-QAM signal encoded by RS encoder in AWGN channel.

Figure 18: BER of 64-QAM signal encoded by RS encoder in AWGN channel.
Figure 19: BER of QPSK signal encoded by RS and BCC encoder in AWGN channel.

Figure 20: BER of 16-QAM signal encoded by RS and BCC encoder in AWGN channel.
Figure 21: BER of 64-QAM signal encoded by RS and BCC encoder in AWGN channel.
3.2.2 System Performance Evaluation of Single Carrier Modulation in Fading Channels

In Section 5 a few reference characteristics of possible HAP fading channels will be discussed and selected. It will be clear that LOS (line-of-sight) channels with Rice factor $K = 10$ dB and $K = 20$ dB are suitable models for the single carrier HAP to train link. Figures 22 to 27 show the performance of single carrier modulation schemes in the reference channels, and the comparison with the AWGN and Rayleigh fading channel. No significant channel difference is observed in signal performance for AWGN and Rice channel with Rice factor $K = 20$ dB independent of coding and modulation scheme, while $2 - 3$ dB higher transmit power compared to the AWGN channel is required when the Rice factor is decreased to $K = 10$ dB. Similar results are expected for other combinations of coding and modulation schemes.

3.2.3 Distortion of single carrier modulated signal in non-linear channel

The distortion of single carrier modulation due to non-linear amplifier were well analysed in the Helinet project [64]. In this report we are going only to summarize results obtained for QPSK and 16-QAM modulation schemes during the research in the Helinet project.

Spectral spreading was evaluated by simulation using computer generated signals passed through filters and HPA. Power spectral density of the transmitted signal was determined using fast Fourier transform (FFT). The spectrum spreading of the square M-QAM signals into neighboring frequency bands, which are amplified by non-linear amplifier, depends on the peak to average power ratio (PAPR) of the signal. The PAPR of unfiltered 4-QAM signal is 0dB and is 2.5dB for 16-QAM signal. The filtering of the signals increases the PAPR. The enlargement strongly depends on filters in use, but generally the enlargement is higher for lower level QAM signals. When the QAM signal is filtered by FIR raised cosine filter with roll-off 0.3 the PAPR obtained by computer simulations are 3.6dB and 4.5dB for 4-QAM and 16-QAM signals, respectively. The average power for PAPR ratio is obtained calculating the average power of computer generated signal, while the peak power is defined as a power, which is not exceeded by 1% of signal samples. The 4-QAM signal has the lowest PAPR and consequently its spectrum spreading due to non-linear amplification is lower than for 16-QAM signals.

A significant difference in spectrum spreading was observed for different batches of the same amplifier type. When boundary characteristics of HPAs, plotted in Figure 22 and Figure 23, are used in computer simulations, 5dB difference is observed in normalised frequency range from 1 to 2.

The results for 4-QAM and 16-QAM signals amplified by the HPA with the highest degradation

![Figure 22: BER for QPSK signal encoded by RS (255,239) encoder in Rice channel.](image-url)
are plotted in Figure 28 and Figure 29, respectively. A 10dB higher power density spectrum in the normalised frequency range between 1 and 2 is observed when comparing the linearly and non-linearly (operating point op=0.0dB relative to the 1dB compression point) amplified 16-QAM signals, while the 4-QAM signal exhibits a 1dB better result. By increasing the input signal power, the input signal is getting closer to the amplifier saturation, and the degradation is greater. This results in higher adjacent channel interference. An increase of signal power by 1dB enlarges the spectrum emission in neighboring bands by approximately 1dB, while a decrease of the signal power by 1dB reduces the spectrum spreading by 1dB.

Figure 23: BER for 16 QAM signal encoded by RS (255,239) encoder in Rice channel.
Figure 24: BER for 64 QAM signal encoded by RS (255,239) encoder in Rice channel.

Figure 25: BER for QPSK signal encoded by RS (255,239) and BCC encoder in Rice channel.
Figure 26: BER for 16-QAM signal encoded by RS (255,239) and BCC encoder in Rice channel.

Figure 27: BER for 64-QAM signal encoded by RS (255,239) and BCC encoder in Rice channel.
Figure 28: Power density spectrum of QPSK signal.

Figure 29: Power density spectrum of 16-QAM signal.
4 Propagation Issues in the Stratospheric Channel

The model of radio channel is required to analyse, design and optimise fading countermeasures in communication systems, such as adaptive coding and modulation, diversity techniques, data rate switching, etc. Unfortunately, there is no universally accepted mobile channel model describing the HAP radio channel for any carrier frequency and a suitable model especially addressing the CAP ANINA scenarios is currently under investigation [36]. However, there are several satellite mobile channel models for different frequency bands [33, 37, 38], which can be adjusted to the HAP channel model, for instance the three state channel model used to model the HAP radio channel in L-band [39]. Among models of satellite land mobile radio channels, the Lutz model [33, 40] based on Markov chain is widely applied in satellite community. The model consists of a generic part, which is a two state finite state machine, and the set of parameters, which allow the adjustment of the channel model to the different propagation scenarios, carrier frequency, and elevation angles. The extension of Lutz satellite land mobile channel model to high speed train propagation environment is proposed in [38], where the three state channel model is introduced for Ku band. The first state describes the LOS condition, the second state represents the shadowed channel, while the third state denotes the channel conditions, when the radio propagation is completely blocked, such as experienced in railway tunnels and under bridges. In addition to these effects the periodic deep fades due to the power supply arches are also modelled.

Due to the lack of suitable HAP channel models based on propagation measurement we apply in this study the channel model for Ku band proposed in [38], although this model does not perfectly fit to the HAP propagation environment. In particular, the elevation angle and the distance in GEO satellite system is nearly constant while in HAP system the elevation angle and the distance vary according to relative position of HAP and mobile. The mobile HAP channel measurements are expected to give different statistical parameters of the channel, however this will not significantly change the properties of adaptive transmission system simulated in this section. The multiplicative noise (fading) in [38] was modelled with three state channel model shown in Figure 30. The signal reflection from the surrounding objects is modelled as a complex Gaussian process. The first state corresponds to LOS conditions, where the received signal is a superposition of strong direct radio ray and weak reflections from surrounding objects. The amplitude $S$ of fading coefficients obeys the Rice distribution [37]

\[ p(S) = K \exp(S + 1) I_0 \left(2K\sqrt{S}\right) \]  

where $I_0$ is the modified Bessel function of order zero. The coefficient $K$ is called Rice factor and determines the ratio between the power of direct and reflected signals. In Figure 30 the LOS condition is represented by the left branch of the model.

In the shadowed conditions, i.e. in the absence of LOS signal, the channel is in the second state represented in Figure 30 by the middle branch of the HAP channel model. The amplitude of fading coefficients obeys Rayleigh distribution. Slow channel variation can be described by multiplying the Rayleigh distributed coefficients by multipliers which obey the lognormal distribution with parameters $\sigma$ and $\mu$ [38]

\[ p_{sh}(S) = \frac{1}{S_0} \exp \left(-\frac{10}{\sqrt{2\pi}\ln 10} \frac{S}{S_0} \right) \exp \left(-\frac{(10\log S_0 - \mu)^2}{2\sigma^2}\right). \]  

The $\mu$ is mean power level decrease and $\sigma^2$ is the variance of power level due to shadowing. The resulting distribution is a Rayleigh/lognormal distribution.

When the received signal is completely blocked due to tunnels and bridges, the received signal level is below the noise margin of the receiver, and the channel is in the third state. The signal is masked by the receiver noise and no information can be retrieved from the received signal.

The finite state machine with three states, depicted in Figure 30, controls the transition from one state to another. The transition probabilities depend on the propagation environment, the speed of the train and carrier frequency. The parameters considered in [39] for the satellite land mobile channel model in Ku band, obtained for a low gain $10 \times 10$ cm flat receiver antenna with 19 dBi gain and the elevation angle of 34 degrees, are summarised for different propagation environments in Table 10. The second, the third and the fourth columns in Table 10 denote the probability of the channel model in LOS, shadowed and blocked state respectively.
Figure 30: The model of HAP mobile radio channel. Here the Rice factor $K$ has been indicated as $c$.

<table>
<thead>
<tr>
<th>Environment</th>
<th>LOS</th>
<th>Shadowed</th>
<th>Blocked</th>
<th>$\mu$ [dB]</th>
<th>$\sigma$ [dB]</th>
<th>$K$ [dB]</th>
</tr>
</thead>
<tbody>
<tr>
<td>Highway</td>
<td>0.90</td>
<td>0.07</td>
<td>0.03</td>
<td>-8</td>
<td>1.5</td>
<td>17</td>
</tr>
<tr>
<td>Rural</td>
<td>0.78</td>
<td>0.16</td>
<td>0.06</td>
<td>-7</td>
<td>2.0</td>
<td>17</td>
</tr>
<tr>
<td>Suburban</td>
<td>0.80</td>
<td>0.17</td>
<td>0.03</td>
<td>-7</td>
<td>2.0</td>
<td>18</td>
</tr>
<tr>
<td>Urban</td>
<td>0.60</td>
<td>0.10</td>
<td>0.30</td>
<td>-7</td>
<td>2.0</td>
<td>17</td>
</tr>
</tbody>
</table>

Table 10: Parameters of the land mobile satellite channel model in Ku band.

The received power of the signal generated by the model of HAP mobile radio channel using parameters for suburban environment is depicted in Figure 31, where different channel states are also indicated.

In the case of wireless access systems for high speed trains, the railway tracks determine the path of the train. In order to overcome the problem of the elevation angle and the distance, we designed the empirical channel model for HAP to train communications, using the representative digital relief model of Slovenia. The railway track, covering all kinds of propagation environments from Koper to Maribor in the length more than 300km is selected as a study case in order to test the HAP performance in real environment. The railway track is shown in Figure 32. On the left lower corner of the map on the coast the port Koper is located. From the coast the train climbs to Karst plateau, then the train descents towards Ljubljana, located approximately in the centre of Slovenia. After that the railway track follows the valleys and gorges of Sava and Savinja rivers towards north/east and through hilly terrain reaches the city of Maribor. If the HAP is located 17km above Ljubljana the yellow line on the map shows the visibility between the platform and the train, while the red line denotes the blocking case. The poor HAP visibility is observed, when the train climbs from the cost to the Karst plateau, because the elevation angle of HAP is low and hills located north/east from port Koper shadow the HAP. The second region of high probability of shadowing occurs in gorge of river Sava, where in spite of high elevation angle, the HAP is blocking due to steep and high slopes of surrounding mountains.

The channel attenuation along the path and time for the selected reference railway track are shown in Figure 33 and Figure 34, respectively. When no direct visibility from platform to train exists, the red marker is plotted on the curve. The channel attenuation is calculated assuming free space propagation.
According to the estimated visibility from the digital relief model, a two-state model can be used to
test the adaptive coding and modulation schemes for HAP communications. The first state represent
the line of sight conditions with Gaussian channel, while in the second state the signal is completely
blocked due to obstacles between HAP and train antennas. At the moment no movement of platform is
considered in the model. In order to take into account also cluttering due to lower obstacles along the
railway track, the two state model can be replaced by three state model, adding a state for shadowed
conditions represented by Rice fading.

It is demonstrated that large scale multipath effect, due to terrain, hills, buildings, other vehicles, is
rarely an issue for rural and sub-urban environments, since obstacle surfaces are “rough” with respect
to the signal wavelength (millimetric in K/Ka bands), thus dispersing the impinging energy instead of
reflecting it back [41]. Measurements taken in the EHF band (40 GHz) from airplanes confirm these
considerations [42, 43]. In this scenario, the structure of the channel model in Figure 30 describes a
single propagation path, more or less shadowed, along the ground-user-to-HAP link direction.

On the other hand, in dense urban scenarios specular reflections are possible, due to the presence
of building surfaces as glass and marble, smooth enough to generate significant multipath up to 40 GHz
at least, with propagation delays from 30 to 850 ns. In this case, a more elaborate channel simulator
should be used, taking into account the multiple copies of the transmitted signal arriving at the receiver
with non-negligible delays [36].

As a consequence of the discussion above, different kind of environments can be considered in
order to fit the channel model parameters to the characteristics of the propagation environment. The
CAPANINA consortium indicated the following (internal document CAP-0233-WP22-POL-CON-P01):
Rural (clear sky and rain), Urban, (clear sky and rain) and Tunnels (model to be investigated). Extrap-
olating measurements done in COST-252, the different multipath conditions (mean number of “active”
macroscopic scatters) can be expected as follows:

- Rural environment (clear sky-LOS, dense trees-NLOs and precipitations): 0 @ 25° elevation angle
  and directive antenna, 1 @ lower elevation angles (scattering from train).
- Urban environment (clear sky – LOS shadowed 15 – 25 dB): 2 to 4 scatters with short delay profile

Figure 31: Received signal power and channel state.
Figure 32: Rail track from Koper to Maribor showing terrain height and HAP visibility for HAP altitude 17km.

• and 2 with long delay. Mean power delay profiles: Attenuations: 20 to 30 dB; Long delays: 350 to 850 ns; Short delays: 30 to 100 ns.

• Urban environment (precipitations – LOS shadowed): 2 to 4 scatters with short delay profile and 2 with long delay. Mean power delay profiles: Attenuations $>30$ dB; Long delays: 350 to 850 ns; Short delays: 30 to 100 ns.
Figure 33: Channel attenuation vs. distance.

Figure 34: Channel attenuation vs. time
5 Operating Scenarios and System Architecture

A fully developed CAPANINA system must provide for bi-directional high-capacity links from the aerial platform to a number of ground stations, which can be either still (customer premises) or in fast movement (trains). Over those links, a variety of broadband traffic is expected, from video-on-demand and video streaming to community TV broadcasting and broadband Internet. In particular, broadband Intranet and video streaming seem to be the most viable applications for the future CAPANINA trials [31]. The above applications require to define a communication framework able to handle high-capacity links which connect pools of end-users (e.g., the passengers of a train, the WLAN users in an enterprise campus) with an “intermediate flying gateway” to the Internet, through suitable wired or wireless access points in the vicinity of the users’ equipment. These links from the “intermediate flying gateway” to the users’ access point can be intended as a kind of “last mile” section of the wireless communication network, which must be carefully designed taking into account the peculiarities of the stratospheric infrastructure and the technological and regulatory constraints it is submitted to.

From the system architecture point of view HAP communication system can be used in different configurations [12,32]. In the simplest configuration HAPs are used as stand-alone platforms, providing broadband wireless access only to terminals located in their coverage area. In the case of multi-platform constellation HAPs can be interconnected via ground stations or by interplatform links (IPL) forming a network of HAPs, thus arbitrarily extending the system coverage. Furthermore, a HAP system can be deployed as a stand-alone network or it can be connected to external networks via gateways providing suitable internetworking functionality.

It is of interest to investigate also the potential gain offered by exploiting space and platform diversity. The principle of diversity is to provide two (or more) statistically independent channels for transmission of the same information. In the case of space diversity, independent channels are provided by receiving the same signal using multiple antennas suitably separated in space. In the scenario of wireless access from train via HAP the implementation of space diversity on train by mounting more than one antenna does not present any problem. On the other hand, due to limited availability of space, weight and power for the communication payload on the platform, space diversity in the sky segment can only be achieved by transmitting the same signal from multiple platforms. Corresponding system architecture alternatives are graphically depicted in Figure 35.

From the perspective of space and platform diversity the above system architecture enables four operating scenarios:

1. Single HAP – single train antenna: This scenario does not provide any diversity hence the transmitted signal is very sensitive to all kinds of fading, including fading due to electrical bridges, trellises and posts [30].

2. Single AP – multiple train antennas: Train antennas are mounted on the roof of the train at appropriate distance from each other to exploit space diversity and thus counteract fading due to electrical trellises, bridges, posts, i.e. obstacles with dimensions smaller than the train length. Such system may exploit the receive diversity in downlink and the transmit diversity in uplink. However, in spite of using space diversity on train the radio signal may be shadowed or even completely blocked if the propagation channels are correlated, i.e. obstacle is larger than the distance between train antennas (deep canyons, localised heavy rain, etc.).

3. Multiple HAPs – single train antenna: In a multiple HAP system there may exist another platform providing LOS link to the train, hence the link reliability can be significantly improved by using platform diversity, where the same signal is transmitted from multiple platforms and received by a single antenna on the train. Such system may exploit the transmit diversity in downlink and the receive diversity in uplink. Platform diversity is particularly attractive in the case where the network of platforms exists, which may be interconnected via interplatform links, via a ground station suitably located in the overlapping part of HAP coverage areas, or even via two separate ground stations in their respective coverage areas, which are connected via terrestrial network.

4. Multiple HAPs – multiple train antennas: In this scenario the same signal is transmitted from multiple platforms and received by multiple antennas on the train. The system can exploit full
space diversity in transmit and receive directions, on uplink and downlink. In this scenario the HAP communication system can be considered as multi input multi output (MIMO) system [33]. The network of HAPs as a MIMO system requires close cooperation between physical, MAC and network layers.

In scenarios 2, 3 and 4, which are exploiting space and/or platform diversity, the received signals may be combined into one signal using the following methods:

- **Maximum Ratio Combining (MRC)**, where coefficients of diversity branches are optimized in order to minimize the noise added from faded branches.
- **Equal Gain Combining (EGC)**, where the signals from all faded branches are equally weighted and superimposed.
- **Selection / Switched Combining (SC)**, where only the strongest signal is used for data detection.

The MRC is the optimum combining method, while EGC and SC incur a penalty of some dB compared to MRC. In Section 6 all three combining methods will be analyzed in combination with adaptive coding and modulation schemes.

### 5.1 Selection of Reference Scenarios

In order to produce consistent comparison with different transmission systems (i.e. modulations and coding schemes, algorithms for the adaptability of the transmission schemes, bit rates, channelizations, synchronization and channel estimation algorithms etc...) we need to identify a few reference operating scenarios. In particular, the consortium agreed to select three high mobility environments, whose parameters are summarized in Table 11.

Before presenting results of the research activity developed in the fields of advanced signal processing techniques for CAPANINA scenarios, it is interesting to observe a link budget, evaluated for one of the most critical applications, which is the high-rate transmission from HAP to high-speed trains.
Table 11: Scenarios to be considered.

### 5.2 An Example of Link Budget

Table 12 reports an example of link budget for the HAP-to-train downlink. The HAP is assumed to fly at an altitude of 20 km and to provide broadband services over a region of about 35 km radius on the ground. The region covered by the HAP is divided into cells and frequency reuse is adopted to avoid inter-cell interference. A uniform cellular structure is obtained by illuminating the covered area with a high-gain multi-beam on-board antenna which generates multiple elliptic beams, to give circular footprints on the ground. Each HAP antenna is supposed to provide 22 dB gain. Within each cell a communication link compliant with the IEEE 802.16a-OFDM standard is established. For the link-budget considered here, a 20 MHz system bandwidth has been assumed, along with a medium-length guard interval (1/8 of the total symbol period), resulting in a coded bit-rate of 61.44 Mbps, provided via a 16-QAM modulation scheme. The channel code is assumed to provide a coding gain of 6 dB.

In order to guarantee a bit error rate of $10^{-6}$ over the whole covered area, the link budget has been evaluated by considering a receiver on board of a train located at the edge of the region. The power received at the ground terminal is affected by atmospheric attenuation effects, evaluated as 4.5 dB, essentially due to oxygen and water vapors, and by Ricean fading, with 20 dB Rice factor as in a rural environment [43]. To ensure an outage probability of $10^{-3}$, the fading margin has been set to 10 dB. Finally, an implementation loss of 2 dB has been included, accounting for both physical and algorithmic loss effects.

The train receiver is equipped with an array of sensors, controlled by an adaptive algorithm whose...
Table 12: Example of link budget for an HAP-to-train scenario.

The main role is to keep the receiver antenna steered toward the direction of the HAP, despite the possible rapid relative movement of both end-points of the link. By adopting a 16 sensor smart antenna, the theoretical receiver antenna gain is 17 dB.

With these assumptions, the platform is required to transmit a power of 15.2 dBW per spot beam, resulting in a total power required for broadband communications of 30.9 dBW.

As it can be observed, such a critical scenario requires the adoption of powerful signal processing techniques at both the transmitter and the receiver, to guarantee elevated service availability, even for high-rate applications.
6 Adaptive Coding and Modulation

We distinguish two types of adaptive modulation schemes proposed to improve the system capacity:

- In pre-estimated adaptive modulation scheme the modulation type, burst structure and data transmission rate are assigned at call setup and the transmission parameters do not adapt to the variation of the channel characteristics during the connection.

- In dynamic adaptive modulation scheme the modulation parameters are controlled slot-by-slot and can be changed adaptively during the connection.

The dynamic adaptive modulation scheme, which is considered in this section, is usually combined with packet reservation multiple access (PRMA) [16]. Suppose that two different signals are combined in the adaptive communication system with spectral efficiency of 1 and 2 bits per symbol respectively. In the case of a good channel two bits are transmitted in one symbol interval, while in the case of a bad channel only one bit per symbol interval is transmitted. In order to obtain the constant data throughput required for some telecommunication services, only one time slot in the frame is reserved when the signal with two bits per symbol is transmitted. Conversely, two time slots are required in the frame when signals with only one bit per symbol are transmitted.

In ACM systems a set of coding modulation schemes have to be chosen for the transmission. According to the HAP channel characteristics modelled as a three state model, as described in Section 4, we propose two groups of coding modulation schemes:

- Bandwidth efficient coding modulation schemes for LOS conditions.
- Power efficient coding modulation schemes for shadowed channel.

For illustrative purposes we chose similar modulation schemes as proposed in IEEE 802.16 single carrier broadband wireless access standard, in particular 16-QAM for LOS conditions and uncoded and encoded QPSK for the shadowed HAP mobile radio channel. The elevation angle variation due to train motion is less than 1% in the observation interval shown in Figure 31, and thus in the following simulations we assume it fixed at 34 degrees. The coding rates of convolutional encoder considered for QPSK signal are $r = 1/2$ and $r = 1/4$. The BER curves of selected signals in Gaussian channel are shown in Figure 36.

Various criteria were proposed in the literature to select the appropriate coding modulation scheme, such as: received signal strength indicator (RSSI) [13], eye closure [17], bit error rate calculated from BCH block codes [13], Euclidean distance, delay spread [44], etc... The RSSI approach is widely used because of its simplicity and quite accurate channel characterisation for flat fading quasi-static radio channels. In flat fading quasi-static radio channel, the channel phase and attenuation are constant during the transmission of one frame. For illustrative purposes we restrict our analysis to quasi-static flat fading radio channel, and therefore the RSSI approach is applied. Assuming quasi-static Gaussian channel and constant amount of noise generated at the receiver, the BER of the received signal is proportional to the received signal strength. The thresholds, which determine the point of switching between different coding modulation schemes, as shown in Figure 36, are calculated from BER dependency on RSSI, as follows.

In order to comply with the constant BER criterion the threshold for switching between two coding modulation schemes is obtained by drawing horizontal line at the target BER. The line intersects the BER curves. The projections of the intersection points on the abscissa define the thresholds for switching between coding modulation schemes. The most robust coding modulation scheme, i.e. $r = 1/4$ convolutionally encoded QPSK modulation, is used when the SNR is between the projection of the first and the second intersection point. The information is carried by the second coded modulation scheme, i.e. $r = 1/2$ convolutionally encoded QPSK, when SNR is between the second and the third projection of intersection points. The coding modulation scheme with the highest spectral efficiency, i.e. 16-QAM, is used for SNR higher than the projection of the last intersection point on abscissa. When the SNR ratio is lower than the first intersection point, the target BER cannot be achieved and no information is transmitted. The BER curves are generated by computer simulations.
6.1 ACM Performance Evaluation

In order to study the increase of the system performance due to adaptive coding modulation we assume a communication system with a perfect carrier and symbol timing recovery, an optimum maximum likelihood receiver and a perfect knowledge of the channel state information at the transmitter. The transmitter selects transmission scheme in accordance to the target BER. Four transmission schemes with different bandwidth efficiency are foreseen for the transmission in the system with adaptive coding and modulation, i.e. 0.5 bits/Hz/s, 1 bits/Hz/s, 2 bits/Hz/s and 4 bits/Hz/s. No signal is transmitted in poor radio channel conditions. Only the most power efficient scheme is transmitted in fixed coding modulation scheme, when the channel allows a BER below the target value.

The instantaneous bandwidth efficiency of the communication system with the fixed coding modulation (FCM) scheme and adaptive coding modulation (ACM) scheme are shown in Figure 37 obtained for the time-series of the channel gain shown in Figure 31. The received power noise is 15 dB below the average received signal strength. The channel attenuation is known at the transmitter side at the moment of the transmission. The target BER is $10^{-3}$. When the FCM scheme is applied, the 4-QPSK signal encoded by 1/4 convolutional encoder is transmitted in the case of channel attenuation when the target BER can be achieved, and no data is transmitted otherwise. The ACM scheme performs similar to FCM scheme in the poor channel conditions, while the coding modulation scheme with higher bandwidth efficiency is used in good and medium channel conditions. Consequently, higher average system throughput is achieved when ACM scheme is applied.

Simulation results confirmed that the FCM modulation scheme has lower BER and much lower system throughput in comparison to the ACM modulation scheme for any simulated environment. The corresponding results for suburban propagation environment are plotted in Figure 38, where the notation $<\text{SNR}>$ denotes average SNR over many frames while the notation $<\text{Bandwidth efficiency}>$ denotes average bandwidth efficiency for each value of SNR. At high SNR the system throughput is limited by the most bandwidth efficient modulation scheme. In the case of ACM scenario the modulation scheme is 16-QAM with bandwidth efficiency $R_b/B = 4 \text{ bits/s/Hz}$, while in the case of FCM we use only $r = 1/4$ convolutionally encoded QPSK modulation with bandwidth efficiency $R_b/B = 0.5 \text{ bits/s/Hz}$. At low SNR both ACM and FCM schemes have the same performance. Since in poor radio channel conditions the modulation scheme with the highest power efficiency is transmitted in both scenarios,
the dots in Figure 37 overlay the squares. When the received signal power occasionally permits the transmission of the modulation scheme with higher bandwidth efficiency, the system throughput of ACM scheme becomes higher in comparison to FCM scheme. However, due to the usage of less robust coding modulation scheme in the ACM scenario, the BER of FCM scheme is lower than the BER of ACM scheme. Nonetheless, the BER is lower than the target value in both scenarios. The decrease of BER at low SNR, observed in Figure 38 for FCM and ACM schemes, is due to two reasons. First, at low average SNR there is a high probability that no data transmission mode is selected, which decreases the average BER, and second, when data is transmitted usually the most robust modulation scheme is selected with the most steep BER curve (see Figure 36). This causes the significant decrease of average BER in the case where the instantaneous channel attenuation is lower than the modulation scheme switching threshold.

Figure 38: Average BER and average bandwidth efficiency for ACM and FCM schemes without diversity.
6.2 Performance of ACM Combined with Spatial Diversity

When the channel is in the blocked state, the required quality of service cannot be achieved even by applying the adaptive coding and modulation scheme. In such a case the communication system reliability can be improved by a suitable diversity technique. Considering the dimensions of the train, the most suitable diversity technique for the scenario addressed in this section is the space diversity, where uncorrelated radio channels can be simply obtained by two antennas mounted at the beginning and at the end of the train. Thus, in the case of Gaussian noise, two uncorrelated channels are assumed, while the state of the channel for the second antenna is calculated taking into account the channel state of the first antenna and the train speed. Combining space diversity with described fixed coding modulation scheme slightly increases the bandwidth efficiency of the system at low signal to noise ratios, as shown in Figure 39. The MRC method outperforms other combining methods independent of signal to noise ratio, because it weights received signals according to their reliability before combining them in a single signal. At low SNR the SC method provides better performance than EGC method. At low SNR the probability is high that one channel is in a deep fading while the other is not. In such a case the SC method detects information only from a good channel, while EGC method superimposes both signals and detects data from the combined signal, which has lower SNR than the SNR of non-faded signal. As shown in Figure 39 the BER values of analysed combining methods are below the target value for the entire observed range of SNR.

Applying space diversity to adaptive coding and modulation significantly increases the system throughput. The corresponding results for suburban environment are depicted in Figure 40. The analysis of the bandwidth efficiency of the ACM scheme with and without diversity in the average SNR range between 2 and 15 dB can provide a rough estimation of an increase in average bandwidth efficiency using different combining methods. The SC method increases the bandwidth efficiency of the system approximately by 20%, while additional 30% of the bandwidth efficiency is gained with EGC method. The MRC method offers 250% higher bandwidth efficiency than ACM scheme without diversity. While in the system without diversity FCM and ACM schemes show similar performance at low BER, in the system with space diversity the ACM scheme outperforms the FCM scheme even at low signal to noise ratios. The average BER is always below target value for all simulated approaches, as shown in Figure 40. Similar results have been obtained also for other propagation environments.

Finally, we analyse also the system performance with platform diversity, which can be efficiently implemented in a network of HAPs. The efficiency of the technique drastically depends on the propagation environment. When blocking and shadowing probability from a single platform is low (rural and suburban environment) no significant improvement of the system performance is observed. However, when the signal from one platform is blocked due to high obstacles, like in an urban area or in rural area when train travels through a deep canyon, the platform diversity can generally provide an alternative propagation channel with LOS conditions. The problem with blocked signal is even more severe in an urban area, where all platforms can be blocked with high buildings most of the time. In situations such as tunnels and street canyons repeater stations might be added on the ground close to the railway to
Figure 40: Average BER and average bandwidth efficiency for ACM scheme with diversity.

Figure 41: Average BER and average bandwidth efficiency for ACM scheme with diversity.

provide an additional diverse signal and hence improve availability.

Figure 41 compares simulation results obtained with the platform diversity to those obtained only with space diversity on trains in suburban and urban environments. In both cases the MRC method was applied for signal combining. Considering the altitude of platforms at approximately 20 km we assume a comparable distance to both platforms, thus on average both radio links exhibit the same signal to noise ratio. In spite of comparable distance to both platforms transmitting the same signal, the received signals exhibit different delays, which causes a similar effect as multipath propagation on the downlink. The problem can be solved by time advancing of the signal in the platform, from which we are expecting longer delays. In the uplink the signal can be delayed according to precalculated propagation delay before combing. Simulation results plotted in Figure 41 assume perfect phase and timing synchronisation of received signals.\(^1\)

While in suburban environment no significant differences are observed in the system performance, the platform diversity technique at high signal to noise ratios in urban environment provides notably better performance than only space diversity technique. The bandwidth efficiency is on average higher for 0.5 bit/Hz/s for high signal to noise ratios. At low signal to noise ratios the dominant impairment on radio link is additive Gaussian noise, hence the difference between the space and platform diversity is not observed. However, at high signal to noise ratios, signal blocking is the main source of signal jamming, and the system performance is significantly improved when the system may take the advantage of link from an alternative platform.

\(^1\)It is evident that, in practice, an efficient synchronization mechanism is needed at the receiver, which can properly delay and synchronise the replica of both signals; a kind of equalisation is required at the receiver, and a synchronisation on the burst level on both transmitted platforms. TJ-28/04
6.3 Performance of ACM Applied to IEEE 802.16 Modulation & Coding Formats

In this section we simulate an ACM system applied to a set of coding modulation schemes chosen from the available coding modulation schemes defined in the single carrier IEEE 802.16. The switching thresholds are determined as described above. The following set of coding modulation schemes are chosen:

- QPSK RS(255, 239) with BCC,
- 16-QAM RS(255, 239) with BCC, and
- 64-QAM RS(255, 239) with BCC.

The target BER in initial analysis is set to $10^{-3}$. The transmit power is set so as to allow the signal reception below target BER. The data rate for channel attenuation depicted in Figure 34 using adaptive coding and modulation is given in Figure 42. The adaptive coding and modulation significantly increases the system throughput. The average system throughput for ACM scheme is 1.98, while the average system throughput for FCM scheme is 1.08 bits per symbol. The coding modulation scheme for fixed coded modulation scheme is chosen so as to guarantee reliable communications. The perfect channel state information is assumed to illustrate the superior performance of ACM scheme over FCM scheme. Errors in channel state information will affect both FCM and ACM scheme, although the degradation ACM schemes performance is expected to be higher. The channel state can be accurately predicted from the train speed and position, while the attenuation to the train is determined by the platform position and the weather condition, so in the analysis we have initially assumed that the channel state is perfectly known at the transmitter. The effect of errors in CSI, and the feasibility of combining schemes that require co-phasing at the receiver of signals from different transmitters, can be considered in further work.
Table 13: Main physical layer parameters for the IEEE 802.16a standard.

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>No. of total OFDM carriers, (N_u)</td>
<td>256</td>
</tr>
<tr>
<td>No. of unused lower carriers, (N_{u,l})</td>
<td>28</td>
</tr>
<tr>
<td>No. of unused higher carriers, (N_{u,h})</td>
<td>27</td>
</tr>
<tr>
<td>Guard interval ratio, (G)</td>
<td>1/4, 1/8, 1/16, 1/32</td>
</tr>
<tr>
<td>Bandwidths, (B_w)</td>
<td>20, 25, 28 MHz</td>
</tr>
<tr>
<td>Modulation schemes</td>
<td>BPSK, QPSK, 16QAM, 64QAM</td>
</tr>
<tr>
<td>OFDM symbol time</td>
<td>12.5 (\mu)s @20MHz, 10 (\mu)s @25MHz, 9 (\mu)s @28MHz</td>
</tr>
</tbody>
</table>

7 Advanced Coding Techniques in OFDM-based IEEE 802.16a standard

As previously discussed, HAP-based broadband communication applications require for adaptive high-gain coding and modulation schemes. When the selection of the CAPANINA reference standard was not finalized yet, initial investigations identified variable-rate, variable-length Serially Concatenated Convolutional Codes (SCCCs) coupled with the IEEE 802.16a-OFDM physical layer as a potential candidate for the CAPANINA modulation and coding scheme. The main parameters of IEEE 802.16a-OFDM have been summarized in Table 13 [45].

The overall system architecture, containing the proposed channel encoding scheme, is depicted in Figure 43, while the architecture of the considered variable-rate, variable-length bit-interleaved modulation and channel coding scheme is shown in Figure 44. Data bits are channel encoded through a variable-rate, variable-length SCCC with overall rate \(R_s\). The encoded bits are first interleaved through a random permutation \(\Pi_M\) of length \(M\), and then modulated through a Gray-mapped \(2^{2n}\)-QAM, \(n = 1, \ldots, 4\) modulator. In this way, all the IEEE 802.16a modulation formats are considered, plus the 256-QAM modulation, which is not included in the standard but is required to guarantee the maximum bit rate specification of the CAPANINA project (120 Mbit/s). The size \(M\) of the random permutation is selected in order to counteract the effects of multipath fading, specially in the downlink transmission. Modulated symbols are fed to an OFDM modulator with \(N_u = 256\) carriers, according to the IEEE 802.16a standard. The theoretical bit rates \(R_b\) achievable with three different code rates \(R_s\) are shown in Table 14 for different values of the system bandwidth \(B_w\).

The modulated data are fed into a IFFT block which generates the OFDM signal. The OFDM-modulated data constitutes the signal:

\[
s(t) = \sum_{m=-\infty}^{+\infty} \sum_{k=0}^{N-1} d_k(m) e^{j2\pi f_k(t-mT_s)} r(t-mT_s) \tag{5}
\]

where \(d_k(m)\) is the \(m\)-th modulated data of the \(k\)-th carrier, and \(r(t)\) is a pulse waveform associated with any symbol (\(r(t) = 1\) for \(t \in [0, T_s]\) and \(r(t) = 0\) elsewhere). In the case of a complex modulation, each \(d_k\) is a complex number \(d_k = d_{kI} + jd_{kQ}\) and the previous equation can be written as follows:

\[
s(t) = \sum_{m=-\infty}^{+\infty} \sum_{k=0}^{N-1} (d_{kI}(m) + jd_{kQ}(m)) e^{j2\pi f_k(t-mT_s)} r(t-mT_s) \tag{6}
\]

After some straightforward mathematical calculations, it is possible to obtain the following OFDM-
modulated signal:

\[ s(t) = \sum_{m=-\infty}^{t=\infty} \sum_{k=0}^{N-1} \left[ d_{Ik} (m) \cos(2\pi f_k (t - m T_s)) - d_{Qk} (m) \sin(2\pi f_k (t - m T_s)) \right] r(t - m T_s) \\
+ j \sum_{m=-\infty}^{t=\infty} \sum_{k=0}^{N-1} \left[ d_{Ik} (m) \sin(2\pi f_k (t - m T_s)) + d_{Qk} (m) \cos(2\pi f_k (t - m T_s)) \right] r(t - m T_s) \]

(7)

where \( T_s \) is the symbol period of the OFDM signal, and \( f_k = f_0 + \frac{k}{N-1} \) for \( k = 0, \ldots, N-1 \) is the frequency of the \( k \)-th OFDM carrier. Because of the oversampling by \( N_s \), simulation time is such that the symbol period of the OFDM signal is \( T_s' = N_s \Delta \), where \( \Delta \) is the simulation time. After OFDM modulation, a guard interval can be inserted into the signal in order to preserve it from channel distortion.

At the receiver side, synchronization and carrier recovery jointly estimate the timing and the carrier frequency offset in order to recover the frequency error and the time delay introduced by the stratospheric channels.

The considered SCCC scheme is shown in Figure 45 along with the variable-length interleaver \( \Pi_M \) used in the serial concatenation. The interleaver length \( N \) can be selected in order to guarantee the required QoS, given possible constraints on the amount of delay introduced by the serial concatenation. The interleaver has been optimized according to the technique discussed in the next section, a design method for variable-length, prunable interleavers that guarantees optimal BER/FER performance for any interleaver size between 1 and the target maximal length \( N \). The convolutional encoders, along with the Puncturing Patterns (PP) applied, have been chosen in order to maximize the interleaver gain [46] produced by the serial concatenation [47]. In particular, the Outer Encoder (OE) polynomial generators are \( G_o (D) = [1, 35/23] \), while the Inner Encoder (IE) polynomial generators are \( G_i (D) = [1, 7/5] \).

Preliminary BER performance are shown in Figures 46–48, for QPSK and 16-QAM modulations. The system and simulation parameters are summarized in Tables 13, 14 and 15. In particular, Table 15 shows the OE code rate \( R_o \), the PP of the OE in octal notation \( PP_o \), the IE code rate \( R_i \), the PP of the IE in octal notation \( PP_i \), the SCCC code rate \( R_c \), the SCCC interleaver size \( N \) and the number or decoding iterations \( N_{it} \). As far as the channel model is concerned, Rice fading with Rice factor equal
$B_w, R_s = \frac{1}{2}$ | BPSK | QPSK | 16QAM | 64QAM | 256QAM |
<table>
<thead>
<tr>
<th></th>
<th></th>
<th></th>
<th></th>
<th></th>
<th></th>
</tr>
</thead>
<tbody>
<tr>
<td>20MHz</td>
<td>7.68</td>
<td>15.36</td>
<td>30.72</td>
<td>46.08</td>
<td>61.44</td>
</tr>
<tr>
<td>25MHz</td>
<td>9.60</td>
<td>19.20</td>
<td>38.40</td>
<td>57.60</td>
<td>76.8</td>
</tr>
<tr>
<td>28MHz</td>
<td>10.66</td>
<td>21.33</td>
<td>42.66</td>
<td>64.00</td>
<td>85.28</td>
</tr>
</tbody>
</table>

$B_w, R_s = \frac{1}{3}$ | BPSK | QPSK | 16QAM | 64QAM | 256QAM |
<table>
<thead>
<tr>
<th></th>
<th></th>
<th></th>
<th></th>
<th></th>
<th></th>
</tr>
</thead>
<tbody>
<tr>
<td>20MHz</td>
<td>10.24</td>
<td>20.48</td>
<td>40.96</td>
<td>61.44</td>
<td>81.92</td>
</tr>
<tr>
<td>25MHz</td>
<td>12.80</td>
<td>25.60</td>
<td>51.20</td>
<td>76.80</td>
<td>102.4</td>
</tr>
<tr>
<td>28MHz</td>
<td>14.22</td>
<td>28.44</td>
<td>56.88</td>
<td>85.33</td>
<td>113.76</td>
</tr>
</tbody>
</table>

$B_w, R_s = \frac{1}{4}$ | BPSK | QPSK | 16QAM | 64QAM | 256QAM |
<table>
<thead>
<tr>
<th></th>
<th></th>
<th></th>
<th></th>
<th></th>
<th></th>
</tr>
</thead>
<tbody>
<tr>
<td>20MHz</td>
<td>11.52</td>
<td>23.04</td>
<td>46.08</td>
<td>69.12</td>
<td>92.16</td>
</tr>
<tr>
<td>25MHz</td>
<td>14.40</td>
<td>28.80</td>
<td>57.60</td>
<td>86.40</td>
<td>115.2</td>
</tr>
<tr>
<td>28MHz</td>
<td>15.99</td>
<td>32.00</td>
<td>63.99</td>
<td>96.00</td>
<td>127.92</td>
</tr>
</tbody>
</table>

Table 14: Theoretical Bit rate, $R_b$, as a function of the SCCC code rate $R_s$ and the system bandwidth $B_w$.

Figure 45: Block diagram of the SCCC encoder. The overall code rate is $R_s \approx R_o \cdot R_i$ (by neglecting the effects of code terminations), whereby $R_o$ is the rate of the outer code, while $R_i$ is the rate of the inner code.

to 10 dB and 20 dB has been considered [42]. Because of the extremely reduced delay spread shown by the stratospheric channel, the guard interval has not been considered in these simulations.

### 7.1 Design of Variable-Rate, Variable-Length Serially Concatenated Convolutional Codes

This section presents the results of extensive search for optimal puncturing patterns for recursive convolutional codes leading to systematic and nonsystematic codes of rate $\frac{k}{k+1}$ ($k$ is an integer) to be used in SCCCs. The code optimization is in the sense of minimizing the required SNR for two target BER and two target FER values. In order to keep the trellis complexity of the code constant and to permit the use of a simplified decoder that can accommodate multiple rates, a mother CC is punctured, to obtain codes with a variety of rates. The focus of this section is on punctured CCs of rate $\frac{k}{k+1}$.

<table>
<thead>
<tr>
<th>$R_o$</th>
<th>$PP_o$</th>
<th>$R_i$</th>
<th>$PP_i$</th>
<th>$R_s$</th>
<th>$N$</th>
<th>$N_M$</th>
</tr>
</thead>
<tbody>
<tr>
<td>$S_1$</td>
<td>2/3</td>
<td>15</td>
<td>3/4</td>
<td>35</td>
<td>1/2</td>
<td>1198</td>
</tr>
<tr>
<td>$S_2$</td>
<td>2/3</td>
<td>15</td>
<td>3/4</td>
<td>35</td>
<td>1/2</td>
<td>3598</td>
</tr>
<tr>
<td>$S_3$</td>
<td>4/5</td>
<td>266</td>
<td>5/6</td>
<td>527</td>
<td>2/3</td>
<td>1331</td>
</tr>
<tr>
<td>$S_4$</td>
<td>4/5</td>
<td>266</td>
<td>5/6</td>
<td>527</td>
<td>2/3</td>
<td>5331</td>
</tr>
<tr>
<td>$S_5$</td>
<td>7/8</td>
<td>15532</td>
<td>6/7</td>
<td>2627</td>
<td>3/4</td>
<td>1369</td>
</tr>
<tr>
<td>$S_6$</td>
<td>7/8</td>
<td>15532</td>
<td>6/7</td>
<td>2627</td>
<td>3/4</td>
<td>4798</td>
</tr>
</tbody>
</table>

Table 15: Parameters of the designed SCCCs. The labels $S_1$-$S_6$ identify six different coding schemes whose performances are investigated in this report.
It is known that for soft-decision Viterbi decoding, the BER of a convolutional code of rate $R_c = \frac{k}{n}$ with BPSK or QPSK modulation in Additive White Gaussian Noise, or even HAP channel with strong LOS component, can be well upper-bounded by the following expression:

$$P_b \leq \frac{1}{k} \sum_{d=d_{free}}^{\infty} w_d Q\left(\sqrt{\frac{2E_b}{N_o}R_c d}\right),$$

(8)

in which $d_{free}$ is the minimum non-zero Hamming distance of the CC, $w_d$ is the cumulative Hamming weight associated with all the paths that diverge from the correct path in the trellis of the code, and re-emerge with it later and are at Hamming distance $d$ from the correct path, and finally $Q(.)$ is the Gaussian integral function, defined as

$$Q(t_o) = \frac{1}{\sqrt{2\pi}} \int_{t_o}^{\infty} e^{-\frac{t^2}{2}} dt.$$

Similarly, it is possible to obtain an upper-bound on the FER of the code as follows:

$$P_f \leq \sum_{d=d_{free}}^{\infty} m_d Q\left(\sqrt{\frac{2E_b}{N_o}R_c d}\right),$$

(9)

where $m_d$ is the multiplicity of all the the paths that diverge from the correct path in the trellis of the code and re-emerge with it later and are at Hamming distance $d$ from the correct path. A classical approach for the design of good punctured codes consists in finding the puncturing pattern that yields a code whose distance spectrum has the property of having the maximum minimum distance $d_{free}$. A better approach is to obtain the distance spectra of the punctured codes and to select the one which minimizes the BER upper-bound based on the first few terms of the distance spectra. The optimum code which leads to the best distance spectra, may be selected as the best punctured code, provided that it is not catastrophic. We recall that the code is catastrophic if a finite number of channel errors can cause an infinite number of decoding errors. In terms of the state transition graph of a CC, this condition requires that this graph should not possess zero-weight cycles other than the self loop associated with the zero state. We note in passing that systematic codes are always non catastrophic. Another aspect which has to be taken into account is related to the recursive nature of the designed punctured CCs. In fact, when an Infinite Impulse Response (IIR) convolutional code is punctured, the resultant encoder is not necessarily recursive. If the punctured CC has to be used in a parallel concatenated scheme, or as an inner code in a serially concatenated code, it must be recursive in order for the interleaver to yield a coding gain.
7.1.1 Problem Formulation and Code Search Technique

Let us review some mathematical notation for punctured CCs. Puncturing is obtained by regularly deleting some output bits of a mother code with rate $1/N$. As a result of puncturing, the trellis of the punctured code becomes periodically time-varying.

Consider a rate $1/2$ systematic mother code to be punctured. Such a code is specified by a $1 \times 2$ generator matrix $G(D) = [1, g_1(D)]$ defined by two polynomials $g_1(D)$ and $g_2(D)$ specifying the connections of the finite state encoder. In this formulation, $g_i(D) = g_{i0} + g_{i1}D + \ldots + g_{i\nu}D^\nu$, where $i = 1, 2$, $g_{il} \in \{0, 1\}$, $l = 0, \ldots, \nu$ and $\nu$ is the code memory (the constraint length of the code is $\nu + 1$). Generator matrix $G(D)$ expresses the fact that the considered encoders are recursive (i.e., we have the ratio of two polynomials in the indeterminate variable $D$). The code symbols which are punctured are specified through zeros in a suitable puncturing pattern matrix, usually indicated as a $2 \times k$ matrix (for a rate $1/2$ mother encoder) or as a sequence of length $2k$ as $<x_1, y_1, x_2, y_2, \ldots, x_k, y_k>$, whereby $x_i$ is the $i$-th systematic bit and $y_i$ the corresponding parity bit output from the encoder. For a rate $k/(k+1)$ code, the input bits are grouped in blocks of $k$ elements, and from the output of the encoder a total number of $k - 1$ bits are deleted.

The adopted code search strategy consists in finding a puncturing pattern which minimizes the required SNR for a total of four different given BER and FER target values (two target BER and two target FER values). Notice that the criteria of maximizing the minimum distance of the code may be valuable if the target BER is extremely small. When on the other hand, the target BER is in the moderate range, it is insufficient to only consider the minimum distance of the code. Indeed, the first few lowest distance terms of the distance spectra of the code should be considered in the optimization process. Furthermore, minimization of the BER and/or FER based on the first few terms of the distance spectra would require knowledge of the CC operating SNR, which is not always given or known specially if the CC is embedded in a PCCC or SCCC scheme. Given these considerations, it would be better to minimize the required SNR, having identified an operating target BER or FER.

We searched for the best puncturing patterns using the inverse of the upper-bound to the BER and FER of the code specified by (8) and (9) as the effective cost function. To limit the search complexity, we used the first four dominant terms of the distance spectra for the evaluation of the inverse functions. The target values of BER and FER were set to the values $BER_1 = 10^{-3}$, $BER_2 = 10^{-6}$, $FER_1 = 10^{-1}$ and $FER_2 = 10^{-3}$.
Furthermore, we extended the search to both the best systematic puncturing patterns, i.e., those puncturing patterns which leave the punctured codes systematic, and the best overall puncturing patterns under the more general condition with no constraint on whether the resulting punctured codes are systematic or not. The reason for this strategy is that at low signal-to-noise ratios, the iterative MAP decoding algorithm used in the decoding of PCCC and SCCC yields better results if the constituent codes are systematic, even though the systematic codes may have worse $d_{\text{free}}$ values in comparison to the non-systematic codes. Obviously, this behavior depends on how many systematic bits the puncturing pattern eliminates.

Mother codes selected for puncturing are the best recursive rate $1/2$, $4$-, $8$- and $16$-state convolutional codes proposed in the literature for the construction of both PCCCs and SCCCs. Matrix generators of the considered codes are shown in the extensive puncturing pattern tables presented in the document.

To the best of our knowledge, all the mother convolutional codes we have used are among the best recursive convolutional codes obtained by using primitive feedback polynomials for the code generator. We have used these codes as mother codes by following the generally accepted rule that “good mother codes” lead to “good punctured codes”.

### 7.1.2 Code Search Results

The results of our code search are presented in Tables 16–25. The tables are organized as follows. Each row contains the best puncturing patterns resulting in punctured codes of rate $\frac{k}{k+1}$ for four different BER and FER optimization targets. In any given row, the upper puncturing patterns are the best global patterns obtained irrespective of whether the resulting punctured codes are systematic or not. The lower patterns are the best puncturing patterns resulting in systematic codes only. Next to each puncturing pattern, we report the following triplets: the punctured code’s free distance, its multiplicity and the cumulative input weight of the error patterns leading to the free distance at the output of the code.

As expected, better results in terms of free distance are obtained when we do not limit the search to only systematic puncturing patterns. However, it is known that the sub-optimal iterative MAP decoding algorithm used for concatenated codes, can yield better results in terms of BER and FER at low SNR values if systematic constituent punctured codes are used. The puncturing patterns which yield better results in terms of the CC’s minimum distance give better results for the SCCC at high SNR values, i.e.,
in the error-floor region.

The puncturing patterns are represented in Octal form. A given puncturing pattern should be read from right to left by collecting \( k \)-pairs of systematic-parity bits. As an example, the puncturing pattern yielding a code with rate \( \frac{2}{3} \) for the 4-state code, should be interpreted as follows: \( p = 133_8 = 1011_2 \Rightarrow < x_1, y_1, x_2, y_2 > \) (the subscript denotes the base of the numbers). In this case the puncturing pattern leaves the encoder systematic and deletes the first parity bit associated with every two input bits. For rate \( \frac{7}{8} \), the non-systematic puncturing pattern is \( p = 13253_8 = 0101101010111_2 \Rightarrow < x_1, y_1, \ldots, x_6, y_6, x_7, y_7 > \).

<table>
<thead>
<tr>
<th>Target FER1</th>
<th>Puncturing Pattern</th>
<th>Target FER2</th>
<th>Puncturing Pattern</th>
<th>Target BER1</th>
<th>Puncturing Pattern</th>
<th>Target BER2</th>
<th>Puncturing Pattern</th>
</tr>
</thead>
<tbody>
<tr>
<td>2</td>
<td>( 13, (3,1,3) )</td>
<td>15, (3,1,3)</td>
<td>13, (3,1,3)</td>
<td>13, (3,1,3)</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>3</td>
<td>( 33, (4,4,7) )</td>
<td>33, (4,4,7)</td>
<td>33, (4,4,7)</td>
<td>33, (4,4,7)</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>4</td>
<td>( 263, (2,4,13) )</td>
<td>263, (2,4,13)</td>
<td>263, (2,4,13)</td>
<td>263, (2,4,13)</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>5</td>
<td>( 1253, (2,4,9) )</td>
<td>1253, (2,4,9)</td>
<td>1253, (2,4,9)</td>
<td>1253, (2,4,9)</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>6</td>
<td>( 1253, (2,4,9) )</td>
<td>1253, (2,4,9)</td>
<td>1253, (2,4,9)</td>
<td>1253, (2,4,9)</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>7</td>
<td>( 1253, (2,4,9) )</td>
<td>1253, (2,4,9)</td>
<td>1253, (2,4,9)</td>
<td>1253, (2,4,9)</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>8</td>
<td>( 1253, (2,4,9) )</td>
<td>1253, (2,4,9)</td>
<td>1253, (2,4,9)</td>
<td>1253, (2,4,9)</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>9</td>
<td>( 1253, (2,4,9) )</td>
<td>1253, (2,4,9)</td>
<td>1253, (2,4,9)</td>
<td>1253, (2,4,9)</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>10</td>
<td>( 1253, (2,4,9) )</td>
<td>1253, (2,4,9)</td>
<td>1253, (2,4,9)</td>
<td>1253, (2,4,9)</td>
<td></td>
<td></td>
<td></td>
</tr>
</tbody>
</table>

**Table 16: Optimal puncturing patterns for the convolutional code with generator matrix**

\[
G(D) = \begin{bmatrix} 1, \frac{1+D^2}{1+D+D^2} \end{bmatrix}.
\]

<table>
<thead>
<tr>
<th>Target FER1</th>
<th>Puncturing Pattern</th>
<th>Target FER2</th>
<th>Puncturing Pattern</th>
<th>Target BER1</th>
<th>Puncturing Pattern</th>
<th>Target BER2</th>
<th>Puncturing Pattern</th>
</tr>
</thead>
<tbody>
<tr>
<td>2</td>
<td>( 13, (3,1,3) )</td>
<td>13, (3,1,3)</td>
<td>13, (3,1,3)</td>
<td>13, (3,1,3)</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>3</td>
<td>( 33, (4,4,7) )</td>
<td>33, (4,4,7)</td>
<td>33, (4,4,7)</td>
<td>33, (4,4,7)</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>4</td>
<td>( 263, (2,4,13) )</td>
<td>263, (2,4,13)</td>
<td>263, (2,4,13)</td>
<td>263, (2,4,13)</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>5</td>
<td>( 1253, (2,4,9) )</td>
<td>1253, (2,4,9)</td>
<td>1253, (2,4,9)</td>
<td>1253, (2,4,9)</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>6</td>
<td>( 1253, (2,4,9) )</td>
<td>1253, (2,4,9)</td>
<td>1253, (2,4,9)</td>
<td>1253, (2,4,9)</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>7</td>
<td>( 1253, (2,4,9) )</td>
<td>1253, (2,4,9)</td>
<td>1253, (2,4,9)</td>
<td>1253, (2,4,9)</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>8</td>
<td>( 1253, (2,4,9) )</td>
<td>1253, (2,4,9)</td>
<td>1253, (2,4,9)</td>
<td>1253, (2,4,9)</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>9</td>
<td>( 1253, (2,4,9) )</td>
<td>1253, (2,4,9)</td>
<td>1253, (2,4,9)</td>
<td>1253, (2,4,9)</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>10</td>
<td>( 1253, (2,4,9) )</td>
<td>1253, (2,4,9)</td>
<td>1253, (2,4,9)</td>
<td>1253, (2,4,9)</td>
<td></td>
<td></td>
<td></td>
</tr>
</tbody>
</table>

**Table 17: Optimal puncturing patterns for the convolutional code with generator matrix**

\[
G(D) = \begin{bmatrix} 1, \frac{1+D+D^2}{1+D+D^2} \end{bmatrix}.
\]
Table 18: Optimal puncturing patterns for the convolutional code with generator matrix $G(D) = \begin{bmatrix} 1 & 1+D+D^3 \\ 1+D+D^3 & 1 \end{bmatrix}$. Symbol $\ast$ means that the systematic puncturing pattern has a lower minimum SNR with respect to the non-systematic pattern.

<table>
<thead>
<tr>
<th>Pattern</th>
<th>$d_{free}$</th>
<th>$m_{free}$</th>
<th>$w_{free}$</th>
</tr>
</thead>
<tbody>
<tr>
<td>3</td>
<td>53, (3, 2, 6)</td>
<td>53, (3, 2, 6)</td>
<td>53, (3, 2, 6)</td>
</tr>
<tr>
<td>4</td>
<td>1253, (2, 1, 2)</td>
<td>1253, (2, 1, 2)</td>
<td>1253, (2, 1, 2)</td>
</tr>
<tr>
<td>5</td>
<td>1253, (2, 1, 2)</td>
<td>1253, (2, 1, 2)</td>
<td>1253, (2, 1, 2)</td>
</tr>
<tr>
<td>6</td>
<td>1253, (2, 1, 2)</td>
<td>1253, (2, 1, 2)</td>
<td>1253, (2, 1, 2)</td>
</tr>
<tr>
<td>7</td>
<td>1253, (2, 1, 2)</td>
<td>1253, (2, 1, 2)</td>
<td>1253, (2, 1, 2)</td>
</tr>
<tr>
<td>8</td>
<td>1253, (2, 1, 2)</td>
<td>1253, (2, 1, 2)</td>
<td>1253, (2, 1, 2)</td>
</tr>
<tr>
<td>9</td>
<td>1253, (2, 1, 2)</td>
<td>1253, (2, 1, 2)</td>
<td>1253, (2, 1, 2)</td>
</tr>
<tr>
<td>10</td>
<td>1253, (2, 1, 2)</td>
<td>1253, (2, 1, 2)</td>
<td>1253, (2, 1, 2)</td>
</tr>
</tbody>
</table>

Table 19: Optimal puncturing patterns for the convolutional code with generator matrix $G(D) = \begin{bmatrix} 1 & 1+D+D^3 \\ 1+D+D^3 & 1 \end{bmatrix}$. 

<table>
<thead>
<tr>
<th>Pattern</th>
<th>$d_{free}$</th>
<th>$m_{free}$</th>
<th>$w_{free}$</th>
</tr>
</thead>
<tbody>
<tr>
<td>3</td>
<td>53, (3, 2, 6)</td>
<td>53, (3, 2, 6)</td>
<td>53, (3, 2, 6)</td>
</tr>
<tr>
<td>4</td>
<td>1253, (2, 1, 2)</td>
<td>1253, (2, 1, 2)</td>
<td>1253, (2, 1, 2)</td>
</tr>
<tr>
<td>5</td>
<td>1253, (2, 1, 2)</td>
<td>1253, (2, 1, 2)</td>
<td>1253, (2, 1, 2)</td>
</tr>
<tr>
<td>6</td>
<td>1253, (2, 1, 2)</td>
<td>1253, (2, 1, 2)</td>
<td>1253, (2, 1, 2)</td>
</tr>
<tr>
<td>7</td>
<td>1253, (2, 1, 2)</td>
<td>1253, (2, 1, 2)</td>
<td>1253, (2, 1, 2)</td>
</tr>
<tr>
<td>8</td>
<td>1253, (2, 1, 2)</td>
<td>1253, (2, 1, 2)</td>
<td>1253, (2, 1, 2)</td>
</tr>
<tr>
<td>9</td>
<td>1253, (2, 1, 2)</td>
<td>1253, (2, 1, 2)</td>
<td>1253, (2, 1, 2)</td>
</tr>
<tr>
<td>10</td>
<td>1253, (2, 1, 2)</td>
<td>1253, (2, 1, 2)</td>
<td>1253, (2, 1, 2)</td>
</tr>
</tbody>
</table>
Table 20: Optimal puncturing patterns for the convolutional code with generator matrix \( G(D) = \left[ \begin{array}{c} 1 + D^2 + D^3 \\ 1 + D + D^2 + D^3 \end{array} \right] \).

<table>
<thead>
<tr>
<th>Target BER1 Pattern</th>
<th>Target BER2 Pattern</th>
<th>Target FER1 Pattern</th>
<th>Target FER2 Pattern</th>
</tr>
</thead>
<tbody>
<tr>
<td>((\text{dfree}, \text{mfree}, \text{wfree}))</td>
<td>((\text{dfree}, \text{mfree}, \text{wfree}))</td>
<td>((\text{dfree}, \text{mfree}, \text{wfree}))</td>
<td>((\text{dfree}, \text{mfree}, \text{wfree}))</td>
</tr>
<tr>
<td>(4, 1, 5)</td>
<td>(4, 1, 5)</td>
<td>(4, 1, 5)</td>
<td>(4, 1, 5)</td>
</tr>
<tr>
<td>27, (3, 1, 5)</td>
<td>27, (3, 1, 5)</td>
<td>27, (3, 1, 5)</td>
<td>27, (3, 1, 5)</td>
</tr>
<tr>
<td>53, (2, 1, 2)</td>
<td>53, (2, 1, 2)</td>
<td>53, (2, 1, 2)</td>
<td>53, (2, 1, 2)</td>
</tr>
<tr>
<td>166, (3, 5, 19)</td>
<td>166, (3, 5, 19)</td>
<td>166, (3, 5, 19)</td>
<td>166, (3, 5, 19)</td>
</tr>
<tr>
<td>5253, (2, 3, 6)</td>
<td>5253, (2, 3, 6)</td>
<td>5253, (2, 3, 6)</td>
<td>5253, (2, 3, 6)</td>
</tr>
<tr>
<td>125253, (2, 38, 76)</td>
<td>125253, (2, 38, 76)</td>
<td>125253, (2, 38, 76)</td>
<td>125253, (2, 38, 76)</td>
</tr>
<tr>
<td>25253, (2, 10, 20)</td>
<td>25253, (2, 10, 20)</td>
<td>25253, (2, 10, 20)</td>
<td>25253, (2, 10, 20)</td>
</tr>
<tr>
<td>2527, (3, 3, 15)</td>
<td>2527, (3, 3, 15)</td>
<td>2527, (3, 3, 15)</td>
<td>2527, (3, 3, 15)</td>
</tr>
<tr>
<td>25253, (2, 4, 8)</td>
<td>25253, (2, 4, 8)</td>
<td>25253, (2, 4, 8)</td>
<td>25253, (2, 4, 8)</td>
</tr>
<tr>
<td>2725252, (2, 83, 166)</td>
<td>2725252, (2, 83, 166)</td>
<td>2725252, (2, 83, 166)</td>
<td>2725252, (2, 83, 166)</td>
</tr>
</tbody>
</table>

Table 21: Optimal puncturing patterns for the convolutional code with generator matrix \( G(D) = \left[ \begin{array}{c} 1 + D^2 + D^3 \end{array} \right] \).

<table>
<thead>
<tr>
<th>Target BER1 Pattern</th>
<th>Target BER2 Pattern</th>
<th>Target FER1 Pattern</th>
<th>Target FER2 Pattern</th>
</tr>
</thead>
<tbody>
<tr>
<td>((\text{dfree}, \text{mfree}, \text{wfree}))</td>
<td>((\text{dfree}, \text{mfree}, \text{wfree}))</td>
<td>((\text{dfree}, \text{mfree}, \text{wfree}))</td>
<td>((\text{dfree}, \text{mfree}, \text{wfree}))</td>
</tr>
<tr>
<td>(4, 1, 5)</td>
<td>(4, 1, 5)</td>
<td>(4, 1, 5)</td>
<td>(4, 1, 5)</td>
</tr>
<tr>
<td>27, (3, 1, 5)</td>
<td>27, (3, 1, 5)</td>
<td>27, (3, 1, 5)</td>
<td>27, (3, 1, 5)</td>
</tr>
<tr>
<td>53, (2, 1, 2)</td>
<td>53, (2, 1, 2)</td>
<td>53, (2, 1, 2)</td>
<td>53, (2, 1, 2)</td>
</tr>
<tr>
<td>166, (3, 5, 19)</td>
<td>166, (3, 5, 19)</td>
<td>166, (3, 5, 19)</td>
<td>166, (3, 5, 19)</td>
</tr>
<tr>
<td>5253, (2, 3, 6)</td>
<td>5253, (2, 3, 6)</td>
<td>5253, (2, 3, 6)</td>
<td>5253, (2, 3, 6)</td>
</tr>
<tr>
<td>125253, (2, 38, 76)</td>
<td>125253, (2, 38, 76)</td>
<td>125253, (2, 38, 76)</td>
<td>125253, (2, 38, 76)</td>
</tr>
<tr>
<td>25253, (2, 10, 20)</td>
<td>25253, (2, 10, 20)</td>
<td>25253, (2, 10, 20)</td>
<td>25253, (2, 10, 20)</td>
</tr>
<tr>
<td>2527, (3, 3, 15)</td>
<td>2527, (3, 3, 15)</td>
<td>2527, (3, 3, 15)</td>
<td>2527, (3, 3, 15)</td>
</tr>
<tr>
<td>25253, (2, 4, 8)</td>
<td>25253, (2, 4, 8)</td>
<td>25253, (2, 4, 8)</td>
<td>25253, (2, 4, 8)</td>
</tr>
<tr>
<td>2725252, (2, 83, 166)</td>
<td>2725252, (2, 83, 166)</td>
<td>2725252, (2, 83, 166)</td>
<td>2725252, (2, 83, 166)</td>
</tr>
</tbody>
</table>
Table 22: Optimal puncturing patterns for the convolutional code with generator matrix $G(D) = \frac{1}{1+D^2+D^3+D^4}$.

<table>
<thead>
<tr>
<th>Target BER1</th>
<th>Pattern</th>
<th>Optimal puncturing pattern</th>
<th>Target FER1</th>
<th>Pattern</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>13, (2,6)</td>
<td>13, (2,6)</td>
<td>13, (2,6)</td>
<td>13, (2,6)</td>
</tr>
<tr>
<td>2</td>
<td>3, (1,3)</td>
<td>3, (1,3)</td>
<td>3, (1,3)</td>
<td>3, (1,3)</td>
</tr>
<tr>
<td>3</td>
<td>253, (3,29)</td>
<td>253, (3,29)</td>
<td>253, (3,29)</td>
<td>253, (3,29)</td>
</tr>
<tr>
<td>4</td>
<td>2, (4,6)</td>
<td>2, (4,6)</td>
<td>2, (4,6)</td>
<td>2, (4,6)</td>
</tr>
<tr>
<td>5</td>
<td>253, (3,12,33)</td>
<td>253, (3,12,33)</td>
<td>253, (3,12,33)</td>
<td>253, (3,12,33)</td>
</tr>
</tbody>
</table>

Table 23: Optimal puncturing patterns for the convolutional code with generator matrix $G(D) = \frac{1}{1+D^2+D^3+D^4}$.

<table>
<thead>
<tr>
<th>Target BER1</th>
<th>Pattern</th>
<th>Optimal puncturing pattern</th>
<th>Target FER1</th>
<th>Pattern</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>13, (2,6)</td>
<td>13, (2,6)</td>
<td>13, (2,6)</td>
<td>13, (2,6)</td>
</tr>
<tr>
<td>2</td>
<td>3, (1,3)</td>
<td>3, (1,3)</td>
<td>3, (1,3)</td>
<td>3, (1,3)</td>
</tr>
<tr>
<td>3</td>
<td>253, (3,12,33)</td>
<td>253, (3,12,33)</td>
<td>253, (3,12,33)</td>
<td>253, (3,12,33)</td>
</tr>
</tbody>
</table>

Table 24: Optimal puncturing patterns for the convolutional code with generator matrix $G(D) = \frac{1}{1+D^2+D^3+D^4}$.

<table>
<thead>
<tr>
<th>Target BER1</th>
<th>Pattern</th>
<th>Optimal puncturing pattern</th>
<th>Target FER1</th>
<th>Pattern</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>13, (2,6)</td>
<td>13, (2,6)</td>
<td>13, (2,6)</td>
<td>13, (2,6)</td>
</tr>
<tr>
<td>2</td>
<td>3, (1,3)</td>
<td>3, (1,3)</td>
<td>3, (1,3)</td>
<td>3, (1,3)</td>
</tr>
<tr>
<td>3</td>
<td>253, (3,12,33)</td>
<td>253, (3,12,33)</td>
<td>253, (3,12,33)</td>
<td>253, (3,12,33)</td>
</tr>
<tr>
<td>Target BER1</td>
<td>Target BER2</td>
<td>Target FER1</td>
<td>Target FER2</td>
<td></td>
</tr>
<tr>
<td>------------</td>
<td>------------</td>
<td>------------</td>
<td>------------</td>
<td></td>
</tr>
<tr>
<td>(dfree, mfree, wfree)</td>
<td>(dfree, mfree, wfree)</td>
<td>(dfree, mfree, wfree)</td>
<td>(dfree, mfree, wfree)</td>
<td></td>
</tr>
<tr>
<td>1</td>
<td>16, (4,1,3)</td>
<td>16, (4,1,3)</td>
<td>16, (4,1,3)</td>
<td></td>
</tr>
<tr>
<td>2</td>
<td>16, (4,1,3)</td>
<td>16, (4,1,3)</td>
<td>16, (4,1,3)</td>
<td></td>
</tr>
<tr>
<td>3</td>
<td>16, (4,1,3)</td>
<td>16, (4,1,3)</td>
<td>16, (4,1,3)</td>
<td></td>
</tr>
<tr>
<td>4</td>
<td>352, (3,2,6)</td>
<td>352, (3,2,6)</td>
<td>352, (3,2,6)</td>
<td></td>
</tr>
<tr>
<td>5</td>
<td>352, (3,2,6)</td>
<td>352, (3,2,6)</td>
<td>352, (3,2,6)</td>
<td></td>
</tr>
<tr>
<td>6</td>
<td>352, (3,2,6)</td>
<td>352, (3,2,6)</td>
<td>352, (3,2,6)</td>
<td></td>
</tr>
<tr>
<td>7</td>
<td>352, (3,2,6)</td>
<td>352, (3,2,6)</td>
<td>352, (3,2,6)</td>
<td></td>
</tr>
<tr>
<td>8</td>
<td>352, (3,2,6)</td>
<td>352, (3,2,6)</td>
<td>352, (3,2,6)</td>
<td></td>
</tr>
<tr>
<td>9</td>
<td>352, (3,2,6)</td>
<td>352, (3,2,6)</td>
<td>352, (3,2,6)</td>
<td></td>
</tr>
<tr>
<td>10</td>
<td>352, (3,2,6)</td>
<td>352, (3,2,6)</td>
<td>352, (3,2,6)</td>
<td></td>
</tr>
</tbody>
</table>

Table 25: Optimal puncturing patterns for the convolutional code with generator matrix \( G(D) = \left[ 1, \frac{1+D+D^2+D^4}{1+D+D^2+D^4} \right] \).
8 Advanced Coding Techniques in Single Carrier IEEE 802.16 standard

This section focuses on the design of advanced coding techniques to be employed in Single Carrier (SC) IEEE802.16-based physical layer. As for the OFDM case, we identified variable-rate, variable-length Serially Concatenated Convolutional Codes and also Parallel Concatenated Convolutional Codes (PCCCs) as potential candidate coding schemes capable of guaranteeing variable degrees of QoS as manifested in the BER performance. Unlike the previous section, however, hereafter we will discuss codes with significantly smaller size with respect to those previously analyzed, due to the fact that in this section we aim also to compare the performance of the proposed schemes with that of some selected codes included in the standard 802.16, whose size is smaller than those analyzed in the previous section.

The architecture of the considered variable-rate, variable-length bit-interleaved modulation and channel coding scheme is shown in Figure 49, similar to Figure 44. In this section, all the IEEE802.16 modulation formats are considered in addition to the 256-QAM modulation format which is not included in the standard but is required to guarantee the maximum bit rate specification of the CAP ANINA project (120 Mbit/s).

With this setup, in the next section we introduce the idea behind variable-length interleavers for concatenated channel code design and for bit-interleaving.

8.1 Design of the Bit-interleaver

This section focuses on interleaver design for the physical layer of HAPs. The design technique described in the following is needed for the design of the interleavers adopted in the code concatenation and for the design of bit-interleaved coded modulations [54].

Consider an indexed set of elements $x_1, x_2, x_3, ..., x_N$. A given interleaver performs a particular permutation of this set of elements. The permutation $\pi$ acts on the indices of the elements. Henceforth, the notation $\pi(i) = j$ is used to mean that the $j$-th input symbol is carried to the $i$-th position at the output. It is a basic result in group theory [55] that any permutation $\pi$ on a set of elements $S$ can be written as a product of disjoint cycles, and $S$ may be divided into disjoint subsets such that each cycle operates on a different subset. A cycle of length two is called a transposition. It is easy to verify that any finite cycle can be written as a product of transpositions. Hence, we conclude that transpositions represent the elementary constituents of any permutation.

The Finite State Permuter (FSP) introduced in [56], is a realization of an interleaver in the form of a sliding window transposition box of fixed length equal to the delay of the permutation it effectuates on its input sequence, and with the property that the transposition performed at a given time slot is responsible for the generation of the output at the same time slot.

Any permutation on a finite set of elements can be represented using a unique transposition vector associated with its FSP realization. As an example consider the permutation

$$\pi = \begin{pmatrix} 1 & 2 & 3 & 4 & 5 \\ 4 & 3 & 1 & 2 & 5 \end{pmatrix}. \tag{10}$$

Consider the queue model of the FSP and assume that data enters from left to right as depicted in
Figure 50: Example of derivation of the transposition vector for a given permutation. The memory cells are depicted as square boxes. The sequence of transpositions is obtained by simply generating the desired outputs from the input and the contents of the memory cells one after another.

Figure 50. Let us label the transpositions to be performed sequentially with the head of the queue to generate the desired outputs, using positive integers. Let the integer 1 denote the case whereby no transposition is performed with the head of the queue and the element at the head of the queue is simply ejected. Then the transposition vector (to be read from left to right) that fully defines permutation $\pi$ is $T_\pi = (4, 2, 2, 1, 1)$. The delay of this permutation is 3, and the FSP can be implemented using three memory cells. Any permutation on $N_\pi$ elements uniquely defines a transposition vector of size $N_\pi$. Conversely, any transposition vector of size $N_\pi$, defines a unique permutation on $N_\pi$ elements. Note that when synthesizing a permutation using the transposition vector $T$, the $k$-th element of vector $T$, when scanned from right to left, can only assume values in the set $\{1, 2, ..., k\}$.

The description of a permutation using the transposition vector of its associated FSP realization turns out to be quite useful for implementing interleavers of variable sizes. This is because of a prefix symbol substitution property that is best described through an example. Consider the example transposition vector $T_\pi = (4, 2, 2, 1, 1)$ associated with permutation $\pi$ above. Take the binary sequence 10110 labelled from left to right. Permutation $\pi$ maps this sequence to 00111. Consider a new transposition vector $T_{\pi_2} = (3, T_\pi) = (3, 4, 2, 1, 1)$ associated with the permutation

$$\pi_2 = \begin{pmatrix} 1 & 2 & 3 & 4 & 5 & 6 \\ 3 & 5 & 4 & 2 & 1 & 6 \end{pmatrix}.$$  \hspace{1cm} (11)

Note that $\pi_2$ looks quite different from $\pi$, yet its output corresponding to the shifted sequence 010110 is 001110 which is a shifted version of the output generated by $\pi$. In essence, the descriptions of $\pi$ and $\pi_2$ using the transposition vectors preserves information about the prefix symbol substitution property of these two permutations on error patterns whereby the first transposition exchanges a zero with a zero, or a one with a one.

8.2 Concatenated Channel Code Design

This section focuses on the design of variable-length concatenated channel codes for improving BER performance as compared to the channel code proposed in the IEEE802.16 standard. We will first...
Recursive, rate-1/2 CC 
Puncturing Matrix 
Variable-length Interleaver 
\( \pi_N \) 
Recursive, rate-1/2 CC 
Puncturing Matrix 

Figure 51: Block diagram of the serial and parallel concatenated codes. The overall code rate is \( R_S \).

present the proposed code architectures, and then we will compare their performance with a code fragment considered in the IEEE802.16 standard.

The considered concatenated channel codes based on both PCCC and SCCC schemes are shown in Figure 51 along with the variable-length interleaver \( \Pi_N \) used in the concatenation. The interleaver length \( N \) can be selected in order to guarantee the required QoS, given possible constraints on the amount of delay introduced by the parallel and the serial concatenations. The interleavers have been optimized using the methods proposed in [56, 58] which present design techniques for variable-length prunable interleavers that guarantees optimal BER performance for any interleaver size between 1 and a target maximal length \( N \). The convolutional encoders, along with the Puncturing Patterns (PPs) applied have been chosen in order to maximize the interleaver gains [46] produced by the code concatenation [47, 57]. For conciseness, we shall omit the details about the interleaver design techniques, the techniques related to the design of the constituent encoders and the choice of the PPs. We invite the interested reader to consult papers [56, 58] for what concerns interleaver design for PCCCs and SCCCs, and the papers [47, 57] for what regards the constituent convolutional encoders and PPs.

For the SCCC scheme whose architecture is depicted in Figure 51, the constituent convolutional encoders are as follows:

- **Outer Encoder (OE) polynomial generators** is \( G_o(D) = [1, 15/17] \),

- **Inner Encoder (IE) polynomial generators** is \( G_i(D) = [2/3] \).

Note that polynomial generators are given in octal notation. As an example, \( 15_8 = 1101 \) corresponds to the polynomial \( 1 + D + D^3 \).

This specific SCCC code architecture presents a simple 2-state IE with rate-1 since the systematic part of the codeword is punctured. There are two basic reasons for which this code architecture is to be preferred to others. First of all, the decoder of the IE is very computationally efficient and the overall serial concatenation converges very fast despite the simple IE decoder. Furthermore, the rate-1 IE allows the design of variable rate SCCCs by simply changing the PP of the OE in the SCCC.

With this architecture in mind, we found a series of PPs for various rates to be applied to the OE following the design criteria proposed in [47, 57]. Note that with this setup, the SCCC rate \( R_s \) corresponds to the OE rate \( R_o \).

In Table 26, we report the best PPs specified in the appropriate column yielding a punctured convolutional encoder with rate \( \frac{k}{k+s} \) (with \( k \in \{2, \ldots , 8\} \) and \( s \in \{1, \ldots , k - 1\} \)) satisfying the design criteria discussed in [47, 57]. For serial concatenation of convolutional codes the asymptotic BER of an SCCC [46] for very large interleaver sizes \( N \), both for LOS transmission and multipath transmission with strong LOS component [36] and for high signal-to-noise ratios, \( E_b/\!\!/N_o \), can be approximated as:

\[
P_b \approx C_o N^{-\frac{d_f^2}{2}} Q \left( \sqrt{d_f^2 d_2 R_s \frac{E_b}{N_o}} \right)
\]

for even values of \( d_f \), and

\[
P_b \approx C_o N^{-\frac{d_f^2+1}{2}} Q \left( \sqrt{\left( d_f^2 - 3 \right) d_2^2 + 2d_{min}^{(3)} R_s \frac{E_b}{N_o}} \right)
\]

for odd values of \( d_f \). In both equations, the terms \( C_o \) and \( C_o \) do not depend on the interleaver length \( N \), \( d_f \) is the free distance of the OE, \( d_2 \) is the effective distance of the IE (i.e., the minimum weight
of the inner code codewords generated by weight-2 input sequences), $R_s$ is the rate of the SCCC, and $d_{min}^{(3)}$ is the minimum weight of the inner code codewords generated by weight-3 input sequences. Equations (12) and (13) suggest that the interleaving gain is given by $N^{-\frac{d_{min}^{(3)}+1}{2}}$ for odd values of $d_{min}^{(3)}$, and $N^{-\frac{d_{min}^{(3)}}{2}}$ for even values of $d_{min}^{(3)}$. In particular, maximizing the free distance $d_{min}^{(3)}$ of the OE can increase the interleaver gain of the serial concatenation. Indeed, this is the basic criterion adopted in the search for optimal PPs to be applied to the OE in the SCCC.

Next to each PP listed in Table 26, we show the minimum distance $d_m$, the number of nearest neighbors $M_m$ yielding the minimum distance $d_m$, and the total weight $W_m$ of these input patterns as the triplet $(d_m, M_m, W_m)$. As an example of how to read the table entries, consider the rate 2/3 PP 7 shown in Table 26. This PP leads to a code whose minimum distance 4 is due to three input patterns with cumulative weight 12.

The PPs are represented in octal notation. A given PP should be read from right to left by collecting $k$-pairs of systematic-parity bits. As an example, the PP in Table 26 which yields a code with rate 2/3 for the 8-state code, should be interpreted as follows: $p = 78 = 011112 = < x_1, y_1, x_2, y_2 >$ (the subscript denotes the base of the numbers). In this case the PP deletes the first systematic bit associated with every two input bits.

Concerning to the PCCC scheme whose architecture is depicted in Figure 51, the constituent convolutional encoders are as follows:

- Upper Encoder (UE) polynomial generators is $G_u(D) = [1, 15/17]$,
- Lower Encoder (LE) polynomial generators is $G_l(D) = [1, 15/17]$.

### 8.3 Simulation Results

This section focuses on BER simulation results of the proposed variable-length, bit-interleaved PCCC and SCCC where we make comparisons with one code fragment as proposed in the IEEE802.16 standard. As a reference example, we considered the concatenation of a systematic Reed-Solomon (RS) outer block code with a nonsystematic convolutional IE, and BPSK/QPSK modulation. Here we focus on the shortest length code proposed in IEEE802.16 composed of a shortened RS (24,18,3) code (i.e., a code whose input frame size corresponds to 18 bytes, and which is capable of correcting 3 erroneous bytes). The standard foresees the serial concatenation of such a RS code with a punctured CE (Convolutional Encoder) of rate 2/3 so that the overall concatenation yields a rate-1/2 code with frame sizes $k = 144, n = 288^2$.

For comparison, we designed two equivalent concatenated codes having the same rate and the same frame sizes. In particular, we considered a SCCC of size $k = 144, n = 295$, and a PCCC of size $k = 144, n = 295$. Slight differences in codeword sizes are due to code termination. In addition, the interleaver designed for both PCCC and SCCC are variable-length prunable interleavers guaranteeing a variable degree of QoS.

Preliminary BER performance curves are shown in Figures (52)-(54) for QPSK modulation. As far as the channel model is concerned, Rician fading with Rice factor $K$ equal to 10dB, 15dB, and 20dB has been considered [42].

\footnote{Note that $k$ is the information frame length, while $n$ is the codeword length.}
Figure 52: Bit error rate performance of QPSK modulation with variable-length PCCCs and SCCCs. The Rice factor is $K = 10$ dB.

In all the figures, the label "RS(24,18)+CC-2/3" refers to the BER performance of the IEEE802.16 serial code discussed above.

Consider as a reference example, curves shown in Figure 52. At a target BER of $10^{-6}$, the proposed SCCC(295, 144) yields a coding gain of about 1.4dB, while the PCCC(295, 144) yields a coding gain of about 2dB compared to the code proposed in the IEEE802.16 standard. For guaranteeing the target BER $10^{-6}$ at the same $E_b/N_o$, we can use the proposed SCCC with a smaller frame size (with performance curve labelled SCCC(151, 72)), while the proposed PCCC(151, 72), despite its smaller size, yields a coding gain of about 1dB.

Similar conclusions can be drawn from Figures 53 and 54, whereby the only difference is the different Fading parameters considered.

Higher coding gains can be achieved by simply employing longer interleavers as shown in Figures 52–54 for the concatenated codes SCCC(432, 216) and PCCC(432, 216). From a practical point of view, coding gains allow increase of the HAP coverage area, or alternatively for the same $E_b/N_o$, allow a reduction of the transmitted power from the HAP while guaranteeing the same target performance.

Finally, note that for applications requiring target BERs greater than say, $10^{-7}$, generally PCCCs present higher coding gains with respect to an equivalent SCCC, so this class of codes are preferred. For applications requiring very low target BER, SCCCs perform better than PCCCs, as is evident from Figures 52–54.
Figure 53: Bit error rate performance of QPSK modulation with variable-length PCCCs and SCCCs. The Rice factor is $K = 15$ dB.

Figure 54: Bit error rate performance of QPSK modulation with variable-length PCCCs and SCCCs. The Rice factor is $K = 20$ dB.
9 Coded Polarization Diversity Aspects

Traditionally, dual-polarized data transmission is an effective frequency-reuse scheme that is widely utilized to exploit the limited bandwidth available for radio communication systems. In such systems, two signals sharing the same bandwidth and carrier frequency are transmitted simultaneously on the two orthogonal polarizations of the antenna. Theoretically speaking, the system capacity of a dual-polarized communication system could be doubled, compared with a single antenna, single polarization system. Nevertheless, the loss of polarization orthogonality, a change in the polarization state of the transmitted signal \[21\], due to both the anisotropy of the transmission medium and the imperfections of the transmit and receive antennas can induce crosstalk and significantly degrade the system performance. Such depolarization is generally mitigated by a cross-polarization interference canceller (XPIC).

9.1 Trellis-Coded Polarization Diversity

In the literature, system performance is typically examined on the basis of uncoded schemes. In fact, pure multiplexing of this sort allows for full independent usage of the antennas; however, it is usually associated with a poor diversity gain and rarely the best transmission scheme for a given BER target. Jointly encoding the multiple data streams can result in additional coding and diversity gain which can help improve the performance, even though the data rate is kept at the same level \[49\]. Trellis-coded modulation (TCM) is a combined coding and modulation technique which enables significant coding gain over uncoded transmission without increasing the bandwidth or the transmitted power \[48\]. Hence, it is widely utilized in terrestrial and satellite mobile communications, where spectrum efficiency is at a premium and transmitted power is limited. Compared with two-dimensional (2-D) schemes, 4-D TCM offers joint encoding between the two orthogonal 2-D constellations, which makes it suitable as a channel code for multiplexing. In this section, we study the performance of 4-D trellis coded polarization multiplexing under varying channel conditions and compare it with spatial multiplexing (SM). A statistical channel model introduced in \[50\] is adopted, applicable to both terrestrial and satellite channels and to a variety of environments. We propose a computationally feasible optimal decoding strategy for 4-D TCM based on such systems and identify the channel states where the use of 4-D TCM is beneficial from an error-probability point of view. Our consideration is restricted to a link with one dual-polarized transmit and one dual-polarized receive antenna.

Figure 55 shows schematically a 4-D trellis-coded polarization multiplexing (PM) system with one dual-polarized antenna at each end of the link. In fact, the underlying channel is a two-input two-output channel since each polarization branch can be regarded as an isolated physical channel.

The 4-D trellis coded data symbol is divided into two constituent 2-D symbols, which are then launched simultaneously from the two orthogonal polarizations sharing the same bandwidth and carrier frequency. After being received on the corresponding polarizations, the 2-D symbols are jointly decoded by a 4-D trellis decoder. In a frequency-flat channel, the received signals can be written as

\[
r = H d + n
\]

where \(d = \begin{bmatrix} d_1 & d_2 \end{bmatrix}^T\) denotes the 2×1 transmit signal vector whose elements are drawn from a finite
(complex) constellation in which the average energy of the elements is 1, \( r = [ r_1 \ r_2 ]^T \) is the 2×1 received signal vector, \( n \) is the 2×1 temporally i.i.d. zero-mean complex Gaussian noise vector and \( H \) is the 2×2 channel transfer matrix or polarization matrix given by

\[
H = \begin{bmatrix}
    h_{11} & h_{12} \\
    h_{21} & h_{22}
\end{bmatrix}
\]  

(15)

where the coefficients \( h_{11} \) and \( h_{22} \) are the gains of the co-polarized antennas, while \( h_{12} \) and \( h_{21} \) are the cross coupling energies from one polarization to the other, respectively. In practice, two polarization schemes are generally used: horizontal/vertical or slanted (±45°). Hereafter, we assume that both transmitter and receiver employ the slanted scheme. This is due to the symmetry in reflections off the earth’s surface for both polarizations, which is not the case for other polarization configurations [23]. Therefore, the channel matrix is taken to be circularly symmetric. Without loss of generality, \( H \) can be decomposed into the sum of a fixed component \( \bar{H} \) representing the line of sight (LOS) propagation and a variable Rayleigh fading component \( \tilde{H} \)

\[
H = \sqrt{\frac{K}{1+K}} \begin{bmatrix}
    h_{11} & h_{12} \\
    h_{21} & h_{22}
\end{bmatrix} + \sqrt{\frac{1}{1+K}} \begin{bmatrix}
    \tilde{h}_{11} & \tilde{h}_{12} \\
    \tilde{h}_{21} & \tilde{h}_{22}
\end{bmatrix}
\]  

(16)

where \( \bar{H} \) and \( \tilde{H} \) are weighted by the Ricean K-factor. The elements of matrix \( \tilde{H} \) do not change and are complex values satisfying

\[
|h_{11}|^2 = |h_{22}|^2 = 1, \quad |h_{12}|^2 = |h_{21}|^2 = \beta
\]  

(17)

where \( 0 \leq \beta \leq 1 \) represents the cross-polarization discrimination (XPD) of the fixed channel component. Good XPD corresponds to small values of \( \beta \) and vice versa. It is important to note that, for pure LOS conditions, \( \beta \) is solely a function of an antenna’s ability to separate the orthogonal polarizations [51].

The elements of the matrix \( \tilde{H} \) are modelled as zero-mean complex Gaussian distributed random variables whose variances depend on the propagation states and the antenna characteristics. In general, we set

\[
E\left\{ \tilde{h}_{11}^2 \right\} = E\left\{ \tilde{h}_{22}^2 \right\} = 1
\]  

(18)

\[
E\left\{ \tilde{h}_{12}^2 \right\} = E\left\{ \tilde{h}_{21}^2 \right\} = \alpha
\]  

(19)

where \( E\{\cdot\} \) indicates statistical expectation and \( 0 < \alpha \leq 1 \) is directly related to the XPD for the scattered component of the channel.

Recent literature simply assumes that the elements of the MIMO channel are i.i.d., or uncorrelated. However, this assumption may be overly optimistic since they are, in general, correlated because of closely spaced antennas and correlated cross coupling. The correlation between the four elements of \( \tilde{H} \) can be described by the normalized correlation coefficients

\[
\rho_t = \frac{E\left\{ \tilde{h}_{11} \tilde{h}_{11}^* \right\}}{\sqrt{\alpha}} = \frac{E\left\{ \tilde{h}_{21} \tilde{h}_{22}^* \right\}}{\sqrt{\alpha}}
\]  

(20)

\[
\rho_r = \frac{E\left\{ \tilde{h}_{11} \tilde{h}_{21}^* \right\}}{\sqrt{\alpha}} = \frac{E\left\{ \tilde{h}_{12} \tilde{h}_{22}^* \right\}}{\sqrt{\alpha}}
\]  

(21)

where \( \rho_t \) and \( \rho_r \) are transmit and receive correlation coefficients, respectively.

Note that, \( \alpha = \beta = 1 \) amounts to having two spatially separated antennas of the same polarization at both ends of MIMO systems. Therefore, it provides an approach to evaluate the performance of polarization and spatial multiplexing strategies for the same channel conditions.

A tilt of either the transmit or receive antenna would clearly change the channel parameters in (16). Any misalignment of the antennas can be described by multiplying \( H \) by a unitary rotation matrix \( U \)

\[
H' = U \cdot H
\]  

(22)
where

$$U = \begin{bmatrix} \cos \psi & -\sin \psi \\ \sin \psi & \cos \psi \end{bmatrix},$$

(23)

and $\psi$ is the tilt angle between transmit and receive antennas. If we assume that the receiver has full knowledge of this angle, it will generate the same estimated symbols as before. Hence, the average system performance will not be affected.

### 9.1.1 Decoding Strategies of 4-D TCM for Multiplexing

We assume that the receiver has perfect channel state information (CSI) and performs maximum-likelihood (ML) detection which computes the vector $\hat{d}$ according to

$$\hat{d} = \arg \min_d \|r - Hd\|^2.$$

(24)

In single-input single-output (SISO) systems the coded 4-D point is transmitted and received by the concatenation of the two constituent 2-D points. The decoder detects the two closest 2-D points individually and uses them to recover the corresponding coded 4-D point. In MIMO systems, however, the two 2-D points need to be jointly determined. In our decoding algorithm, the decoder searches the closest 4-D points in each of the sixteen 4-D types directly according to (24) and records their associated branch metrics. Note that, each 4-D type is formed by the concatenation of a pair of 2-D subsets. Next, eight 4-D subset metrics selected from the above sixteen 4-D type metrics are used to update the survivor path metrics in the trellis. The decoder then picks up the path with minimum metric and makes the final decision on the previously transmitted 4-D points. Eventually, the decoded data bits are generated.

### 9.1.2 Performance

In this section, a performance evaluation of the 4-D trellis coded PM and SM schemes for varying channel conditions is presented assuming that the antennas are aligned. A rate-4/5 16-state 4-D trellis code [52] is chosen as channel code. As a reference system we used an uncoded 16-QAM which has the same information bit rate as the trellis code.

We separate the channel conditions into two cases, NLOS and LOS. For the NLOS environment, we consider a correlated Rayleigh fading channel with $K = 0$, while for the LOS environment, a Ricean fading channel with $K = 10$ is assumed. The correlation coefficients are selected to be $\rho_t = 0.5$ and $\rho_r = 0.3$. For PM system, we choose a high XPD for the fixed component of the channel, $\beta = 0$, and $\alpha = 0.4$ as the XPD for the scattered component. For SM system, it is set to be $\alpha = \beta = 1$.

The SNR is defined as

$$\text{SNR} = 10 \log \left( \frac{2E_s}{\sigma_n^2} \right) \text{dB}$$

and

$$10 \log \frac{E_b}{N_0} = 10 \log \left( \frac{2E_s}{\sigma_n^2} \right) + 10 \log \frac{1}{2RM_c}$$

(25)

where $R$ is the code rate and $M_c$ is number of bits per symbol.

We assume that the channel is constant over a block of symbols and ML decoding with perfect channel knowledge is performed at the receiver.

First, we investigate the performance of coded PM and SM schemes over correlated Rayleigh and Ricean fading channels with typical parameters provided above. Figure 56 and Figure 57 present the simulation results, respectively.

They show that in a Rayleigh environment, the use of dual-polarized antennas in an uncoded 16-QAM system usually results in about 1 dB loss. On the other hand, there is a negligible performance degradation of the 4-D trellis coded PM scheme. It is also found that the coding gain of 4-D TCM in PM system is larger than that in SM system. 1.0 dB and 0.5 dB improvement at BER of $10^{-3}$, and 2.3 dB and 1.5 dB improvement at BER of $10^{-4}$ are achieved by 4-D TCM for PM and SM schemes, respectively. In a Ricean environment, however, the performance of PM system is significantly superior to that of SM system. Improvements of about 12 dB at BER of $10^{-3}$ and about 15 dB at BER of $10^{-4}$ are obtained for
both coded and uncoded schemes. It is important to note that 4-D TCM can offer 7.7 dB coding gain for the PM system at BER of $10^{-5}$ and this coding gain is expected to be larger for lower BER requirement.

The next two simulation examples serve to demonstrate the influences of $\alpha$ and $\beta$ on the 4-D trellis coded PM scheme for the NLOS environment and the LOS environment, respectively. The required average $E_b/N_0$ ratio at BER $= 10^{-3}$ is concerned.

As Figure 58 shows, in a correlated Rayleigh fading channel the PM system performance degrades as $\alpha$ decreases. This is because high XPD reduces the mean power of the cross-coupled component, and thus, the available polarization diversity arising from uncorrelated cross coupling diminishes. This case is much worse when $\alpha \leq 0.4$. When $\alpha \geq 0.4$ the performance loss is within 1.4 dB and 0.8 dB for uncoded 16-QAM and 4-D TCM, respectively. Since the use of dual-polarized antennas has the advantage of saving cost and space, we can consider employing polarization multiplexing in conditions in which $\alpha \geq 0.4$.

In Figure 59, we choose a LOS propagation environment with $\alpha = 1$. This condition generally occurs in satellite communications or wireless communications from High-Altitude Platforms (HAPs) [53] for large ranges. It is clear from Figure 59 that the system performance improves dramatically with decreasing $\beta$. The main reason is that in a dominating LOS environment, the channel orthogonality improves with increasing XPD. For low XPD, the LOS component of the two data streams cannot be separated at the receiver, and hence, the system performance is significantly degraded. We also notice that 4-D TCM provides significant coding gain for all the cases especially when $0.4 \leq \alpha \leq 0.8$ and $0.4 \leq \beta \leq 0.8$. This can be interpreted in that for high XPD, the coding gain which derives from joint encoding over two polarizations diminishes with the loss energy of the cross-coupled channels. The polarization diversity gain is poor. On the other hand for low XPD, the lack of polarization orthogonality brings about insufficient separability of subchannels, which makes the MIMO channel less selective. Therefore, the coding gain of 4-D TCM in PM systems benefits from a trade-off between reduced polarization diversity on the one hand, and increased polarization orthogonality on the other.

Finally, we study the influence of the $K$-factor on both PM and SM schemes. Again, $\alpha = 0.4$ and $\beta = 0$ are chosen for PM system. It can be seen from Figure 60 that while PM is inferior to SM for low $K$-factor, it is markedly superior to SM for medium and high $K$-factor. This advantage extends as $K$ increases. In the NLOS environment, the system performance depends on the diversity gain which is diminished since the deployment of dual-polarized antennas reduce the energy of the cross-coupled channels. However, in the LOS environment, which is closer to the AWGN channel, the use of dual-polarized antennas makes it easier to separate the two subchannels, and therefore improves the system performance. In fact, for $\beta \to 0$ and $K \to \infty$ the PM system becomes two independent AWGN
Figure 57: BER of the 4-D trellis-coded PM and SM schemes over Ricean fading channel ($\mathcal{K} = 10$).

channels, and hence outperforms the fading channel with the same signal power.
Figure 58: Required $E_b/N_0$ for the BER $= 10^{-3}$ as a function of $\alpha$, $(K = 0)$.

Figure 59: Required $E_b/N_0$ for the BER $= 10^{-3}$ as a function of $\beta$, $(K = 10)$. 
Figure 60: Required $E_b/N_0$ for the BER $= 10^{-3}$ as a function of $K$-factor.
9.2 Joint Space-Time Coding and Polarization Diversity

The application of dual-polarized antennas in conjunction with spatial multiplexing and space-time coding has been investigated by [50]. All of their considerations are restricted to a system with one dual-polarized antenna at each end of the link. Hereafter we investigate the performance of the hybrid transmission system by setting two dual-polarized transmit antennas and one dual-polarized receive antenna at both ends of the system. Thereby, we propose a scheme that can support two cochannel terminals while providing a diversity order of two to both terminals.

Figure 61 and Figure 62 illustrate the block diagrams for the transmitter and receiver of the hybrid spatial multiplexing and STBC (space time block code) system, respectively. Two dual-polarized transmit antennas and one dual-polarized receive antenna are set at both ends of the link. In fact, the underlying channel is a four-input two-output channel since each polarization branch can be treated as an isolated physical channel. Each terminal is equipped with two nominal transmit antennas and uses STBC over the same polarizations. The overall system can be view as a PM-STBC.

We define the received signal vectors \( r_1 = [ r_{11} \quad r_{12} ]^T \) and \( r_2 = [ r_{21} \quad r_{22} ]^T \) at the two orthogonal polarizations described as vertical and horizontal, the two code symbol vectors transmitted from the first and the second terminals \( c = [ c_1 \quad c_2 ]^T \) and \( s = [ s_1 \quad s_2 ]^T \), and the i.i.d. complex Gaussian random variables \( \eta_1 = [ \eta_{11} \quad \eta_{12} ]^T \) and \( \eta_2 = [ \eta_{21} \quad \eta_{22} ]^T \) with zero mean and covariance \( N_0 \cdot I_2 \).

The received signal vectors at the two orthogonal polarizations of the receive antenna are then

\[
    r_1 = H_1 \cdot c + G_1 \cdot s + \eta_1 \\
    r_2 = H_2 \cdot c + G_2 \cdot s + \eta_2
\]

where the elements in channel matrices \( H_1 \) and \( G_2 \) are gains between the co-polarized vertical and horizontal channels, respectively. Similarly, the elements in channel matrices \( H_2 \) and \( G_1 \) are the cross couplings from one polarization to the other, horizontal-to-vertical and vertical-to-horizontal, respectively. They are given by

\[
    H_1 = \begin{bmatrix} h_{11} & h_{21} \\ h_{21} & -h_{11} \end{bmatrix}, \quad G_1 = \begin{bmatrix} g_{11} & g_{21} \\ g_{21} & -g_{11} \end{bmatrix}
\]
As in the previous section, these channel matrices can be decomposed into the sum of a fixed component representing the line of sight (LOS) propagation and a variable Rayleigh fading component which are weighted by the Ricean $K$-factor. Furthermore, the variances of the elements in $H_2$ and $G_2$ are directly related to the cross-polarization discrimination (XPD) of the channel and thus are less than 1. Detail explanation of the polarization channel matrix can be found in [50].

The overall received signal vector can be written as

$$r = \begin{bmatrix} r_1 \\ r_2 \end{bmatrix} = \begin{bmatrix} h_{12} & h_{22} \\ h_{22} & -h_{12} \end{bmatrix} \begin{bmatrix} c \\ s \end{bmatrix} + \begin{bmatrix} \eta_1 \\ \eta_2 \end{bmatrix}$$

(30)

We adopt the zero-forcing scheme introduced by [26] to completely remove the interference between the two ST cochannel terminals. The compensated signal values for data detection are

$$W \cdot r = \begin{bmatrix} \tilde{r}_1 \\ \tilde{r}_2 \end{bmatrix} = \begin{bmatrix} \tilde{H} & 0 \\ 0 & \tilde{G} \end{bmatrix} \begin{bmatrix} c \\ s \end{bmatrix} + \begin{bmatrix} \tilde{\eta}_1 \\ \tilde{\eta}_2 \end{bmatrix}$$

(31)

where the matrices $\tilde{H}$ and $\tilde{G}$ are orthogonal and their elements are i.i.d. complex Gaussian random variables with zero mean and variance 1. Hence, using the matrix linear combiner $W$ will reduce the problem of detecting the two cochannel ST terminals into two separate problems that have a much simpler solution. It follows immediately that the optimum ML decoder for the symbols transmitted from the first and the second terminals are

$$\hat{c} = \arg \min_{\hat{c} \in C} \| \hat{r}_1 - \tilde{H} \cdot \hat{c} \|^2$$

(32)

$$\hat{s} = \arg \min_{\hat{s} \in S} \| \hat{r}_2 - \tilde{G} \cdot \hat{s} \|^2.$$  

(33)
9.2.1 Simulation Results

The performance of the hybrid spatial multiplexing and STBC scheme with uni-polarized antennas (SM-STBC) and that with the dual-polarized antennas (PM-STBC) are given by means of computer simulations. The channel conditions are separated into two cases, NLOS and LOS. For the NLOS environment, we consider a Rayleigh fading channel, while for the LOS environment, a Ricean fading channel with $K=10$ is assumed. It is also assumed that the transmitted and received signals are uncorrelated and the channel is constant over a block of symbols. ML decoding with perfect channel knowledge is performed at the receiver. The SNR is defined as $E_s/N_0$ per receive antenna (each polarization of receive antenna).

Figure 63 and Figure 64 show the BER performance of uncoded coherent BPSK for the combined SM-STBC scheme and the pure STBC scheme over Rayleigh and Ricean fading channels, respectively. Note that, SM-STBC allows for doubling the data rate. It can be seen from Figure 63 that the performance of SM-STBC with uni-polarized antennas is the same as that when only one terminal exists with two transmit and one receive uni-polarized antennas. The cochannel interference is perfectly suppressed and the system throughput is doubled. The use of dual-polarized antennas, however, generally leads to a performance degradation. For STBC, this is due to the power loss of the cross-coupled channels, the available polarization diversity arising from uncorrelated cross coupling diminishes. For SM-STBC, the reason is that although the compensation matrix $W$ completely removes the cochannel interference it inevitably enhances the noise and thus the system performance is degraded. The performance loss due to the application of dual-polarized antennas for SM-STBC and STBC is 1.2 dB and 1.9 dB, respectively. Since employing dual-polarized antennas has the advantage of saving cost and space, it can be taken into consideration where the performance degradation is acceptable.

In a high LOS environment such as satellite communications and some forms of terrestrial broadband fixed wireless access (BFWA), however, the performance of PM-STBC (with dual-polarized antennas) is greatly superior to SM-STBC (with uni-polarized ones). It can be interpreted in that the cochannel interference from other terminals is extremely high when uni-polarized antennas are deployed. Although $W$ can effectively suppress such interference, the enhanced noise it brings is intolerable. On the other hand, the polarization orthogonality provided by the dual-polarized antennas makes it easier to separate the two terminals at the receiver and therefore minimizes the cochannel interference. It is clear from Figure 64 that the performance of PM/SM-STBC system is close to that of STBC while providing double data rate.
Figure 63: BER performance of PM-STBC compared to that of STBC over Rayleigh fading channel.

Figure 64: BER performance of PM-STBC compared to that of STBC over Ricean fading channel ($K = 10$).
10 Channel Estimation and Equalization

Kalman filter based algorithms have been investigated in the framework of analyzing implementation aspects of signal processing algorithms developed in software radio suitable to the aerial platform, as one of the most serious candidate for channel estimation and synchronization.

Kalman based equalisers are well-known solutions and many works have already been disseminated regarding both theories and practical implementation. However, most of the reported works are guided to bandwidths lower than the required ones in HAPs applications. For example, the use of IEEE 802.16-SC standards in HAPs could require up to 28 MHz of base-band bandwidth. This high bandwidth imposes new restrictions to Kalman filter programming, mainly related with the possible number of instructions executable by DSP devices between consecutive samples of the received data. There is a compromise between processing speed and algorithm stability (i.e, filter parameter stability), or, in other words, among algorithm optimisation, MIPS in the DSP, and the selection of the most suitable DSP arithmetic (fixed/ floating point algebra). Besides, other implementation issues such as quantization noise, modularisation, adaptability, computational efficiency or consumed power are also of capital importance. All these issues affect to the cost of HAP user’s terminals and their time of autonomy (in the battery operated case).

The evaluation of DSP capabilities to support Kalman equalisation algorithms regarding IEEE802.16 data-rates will be assessed by reproducing real working conditions. The bench for testing such algorithms is composed by analog microwave devices capable to reproduce propagation channel conditions (i.e, multipath) taking into account the significant features of the physical layer of the communication standard (i.e, according to IEEE 802.16-2004, doc: IEEE Std 802.16-2004, Part 16: Air Interface for Fixed Broadband Wireless Access Systems).

Only DSP implementation will be considered here. Nevertheless, if final experimental results show important DSP limitations, then the use of other devices, such as FPGA’s, could be considered in further work. Currently, available last generation FPGA’s devices are capable of doing typical DSP tasks (multiply/accumulate).

10.1 Channel Models

The discrete multipath channel can be baseband modelled as follows:

\[ x(t) = \sum_{i=0}^{N-1} \alpha_i s(t - \tau_i (t)) \]  

(34)

where \( N \) in the number of rays (scatters) impinging the receiver, \( s(t) \) in the bandpass input signal, \( \alpha_i \) is the \( i \)-path attenuation, and \( \tau_i \) is the \( i \)-path delay. Consequently, the channel impulse response (CIR) may be represented as:

\[ h(t) = \sum_{i=0}^{N-1} \alpha_i(t) \delta(t - \tau_i) \]  

(35)

The discrete time channel output \( x(k) \) may be written in terms of the channel input \( s(k) \) and the noise \( n(k) \), assumed zero mean, as:

\[ x(k) = h^T s(k) + n(k) \]  

(36)

where the superscript \( T \) indicates transpose, \( h \) is the \( N \)-point impulse response vector \( h^T = [h_0, h_2, \ldots, h_{N-1}] \), and \( s(k) \) contains the last \( N \) inputs to the channel. This output \( x(k) \) is often used like the observation equation in Kalman filter formulation.

The channel taps may be written as:

\[ \alpha_i(t) = \sigma_i + \beta_i(t) \]  

(37)

In this work the channel multipath will be hardware-reproduced (analog simulation), so we don’t need this model for computer simulation. Hence, no comments about Bello’s or Jake’s models will be done in this report. The interest of the model is just for Kalman filter formulation.
being $\sigma_i$, the constant part of $\alpha_i$. The time-varying part $\beta_i$ is a zero-mean wide-sense stationary Gaussian process uncorrelated with any other $\beta_{i,j}$, and has time-autocorrelation properties governed by the Doppler rate $f_d T_s$ ($T_s$ is the baud duration).

As previously introduced, the impulse response of the channel, $h(t, i)$ may be modelled as a time invariant (mean response) component, $\bar{h}(t)$, plus a zero-mean randomly time-varying part, $\tilde{h}(t, i)$:

$$h(t, i) = \bar{h}(t) + \tilde{h}(t, i)$$ (38)

The time-varying part $\tilde{h}(t, i)$ follows an AR model (Gauss-Markov process) which represents the time-varying components of the channel tap gains because any stationary random process can be represented as an infinite tap AR process. However, an infinite tap AR process model is impractical, so it is truncated in practice to a $P$-tap form, being possible to represent it by a difference equation of the form:

$$\tilde{h}(k) = \sum_{i=0}^{P-1} \phi_i \tilde{h}(k-1) + w(k)$$ (39)

being: $h(k)$ a complex Gaussian process, $\phi_i$ the parameters of the model, $P$ the number of delays in the autoregressive model, $w(k)$ a sequence of identically distributed zero-mean complex Gaussian random variables (white noise).

This difference equation may be reduced to a state model in the vector form:

$$\tilde{h}(k) = F \tilde{h}(k-1) + w(k)$$ (40)

where $\tilde{h}(k)$ and $w(k)$ are column vectors of size $P \times 1$, and $F$ is a squared $P$-matrix (state transition matrix associated with the channel), resulting the following equation:

$$\tilde{h}(k) = F_1 \tilde{h}(k-1) + \ldots + F_P \tilde{h}(k-P) + w(k)$$ (41)

where $\tilde{h}(k) = [\tilde{h}(k, 0), \ldots, \tilde{h}(k, N-1)]^T$. The AR parameters are in matrices $F_i$.

### 10.1.1 Doppler Effect

We will consider a 20 km altitude HAP and a coverage radius of 35 km. The carrier frequency is 28 GHz and we suppose a train at average velocity of $v_T = 200$ km/h. Then the Doppler velocity $v_d$ related to this situation will be:

$$v_d = v_T \cos^{-1} \left( \tan^{-1} \left( \frac{20}{35} \right) \right) = 200 \cdot 0.868 = 173.64 \text{ km/h} = 48.2 \text{ m/s}$$ (42)

and the corresponding Doppler frequency:

$$f_d = \frac{v_d}{\lambda} = v_d \frac{f_c}{c} = (48.2) \frac{28 \cdot 10^9}{3 \cdot 10^8} = 4.5 \text{ kHz}$$ (43)

### 10.1.2 Number of path rays to be considered

In the receiver the effect of antenna directivity implies a significant attenuation, according to the following equation:

$$P_R = P_T G_T G_R \left( \frac{c}{4\pi f_d} \right)^2$$ (44)

where $G_R$ is the receiver antenna gain.

The considered model will be up to three rays (in a pessimistic model) because of the attenuation due to the arrival angle and antenna directivity, especially for the higher paths. The second ray is assumed to be delayed 1 times the symbol time ($T_s$). The third ray 2 times ($2T_s$). In this case the relation between transmitted and received signal will be:

$$s_R(t) = \alpha_0 s_T(t) + \alpha_1 s_T(t - T_s) + \alpha_2 s_T(t - 2T_s)$$ (45)
or in discrete time:

\[ s_R(k) = \alpha_0 s_T(k) + \alpha_1 s_T(k - 1) + \alpha_2 s_T(k - 2) \]  

(46)

For the case of a symbol time of 50 ns the maximum delay that is considered in the previous model is:

\[ T_{delay} = 2T_s = 100 \text{ ns} \]  

(47)

corresponding to a maximum difference between the main and the third propagation path of:

\[ \Delta d = c \cdot T_{delay} = (3 \cdot 10^8) \cdot (100 \cdot 10^{-9}) = 30 \text{ m} \]  

(48)

In this case, the fading effects (modelled as a 3-tap FIR) may be mitigated with a 3-tap AR (IIR) equaliser, that, thanks to its adaptability, is able to identify the correct parameters.

10.2 Adaptive Equalisation

The channel parameters identified for the Capanina contexts are a first approach because, up to date, there are not allowable detailed models of multipath for HAPs operating at Ka band, where rain and clouds attenuations are the predominant fade mechanisms. Because of this natural uncertainty about some aspects of propagation channel, we decided to face channel equaliser with adaptive capability, thus having results more insensible to future advances in channel modelling.

The considered scenarios for evaluating the designed algorithms are compliant with the reference parameters discussed in Section 5. Modulation formats and bit rates are compliant with the Standard IEEE 802.16-SC.

The channel estimation and equalization methods based on training sequences assume that the receiver has a-priori knowledge on the sequences being sent over the channel during particular time intervals. This knowledge is used to compare actually received sequences with the transmitted ones, and the minimisation process of the resulting error is used to adjust the channel equaliser parameters. Main advantages of training sequences are the conceptual simplicity of the method, which is facilitated in some new communication standards that includes frames for sending training sequences. Obvi-ously, the drawback is because the time-slots occupied in the transmission of these training sequences, reducing throughput.

Blind methods have not the transportability of the methods based on training sequences, being sensitive to the particular communication system because of their dependence on the transmitted information over the channel. The blind methods use certain underlying mathematical information about the kind of data being transmitted, not requiring a deterministic knowledge of the actually transmitted sequence. They are bandwidth efficient because does not occupy time-slots in which the normal data communication is interrupted. However, they are significantly slow to converge (more than 1000 symbols may be required for a FIR channel with 10 coefficients) and require important computational capacity.

In Figure 65 we can see the probability error (BER) of a number of different modulations as a function of \( \frac{E_b}{N_0} \) for the case of a channel free of multipath (AWGN). This figure can be considered as a reference for testing the equaliser capabilities. In fact, the objective of the desired equaliser is to mitigate (ideally, to nullify) the multipath effects. Hence, theoretical values in Figure 65 are the asymptotic behaviour of the signal once equalised.

10.2.1 Equaliser Structures

We expose in this section some possibilities of how solve the equalization problem.

[Case a] Indirect equalisation (estimation+equalisation)  

The idea is to run a channel estimation algorithm and then to use the estimated parameters to update the equaliser coefficients (see Figure 66). Both processes of estimation and equalisation are carried out independently and consecutively in time.

The channel model is a 3 taps FIR, as:

\[ s_R(k) = \alpha_0 s_T(k) + \alpha_1 s_T(k - 1) + \alpha_2 s_T(k - 2) \]  

(49)
Thus being the channel impulse response as:

\[ h_c(k) = \alpha_0 \delta(k) + \alpha_1 \delta(k-1) + \alpha_2 \delta(k-2) \] (50)

in a time-invariant case, the channel transfer function will be:

\[ H_c(z) = \alpha_0 + \alpha_1 z^{-1} + \alpha_2 z^{-2} \] (51)

If the channel model is a 3 taps FIR, then the estimator has to be coherent in structure and order:

\[ h_{id}(k) = h_0 \delta(k) + h_1 \delta(k-1) + h_2 \delta(k-2) \] (52)

being \( h_0, h_1, \) and \( h_2 \) the time-variant coefficients which should converge to \( \alpha_0, \alpha_1, \) and \( \alpha_2. \) LMS and Kalman filter are suitable solutions for updating these coefficients, and both will be tested below in this study.

The ideal equalizer, in the time-invariant case, will correspond to a second order AR filter with transfer function:

\[ H_{eq}(z) = \frac{1}{\alpha_0 + \alpha_1 z^{-1} + \alpha_2 z^{-2}} \] (53)

Then, in the time-variant case the time domain equaliser equation will be:

\[ y(k) = \frac{1}{h_0} (x(k) - h_1 y(k-1) - h_2 y(k-2)) \] (54)

where \( x(k) \) corresponds to the received signal \( s_R(k) \) and \( y(k) \) is the equaliser output. The \( h_0, h_1 \) and \( h_2 \) are the coefficients obtained after the identification process during every training sequence.

**[Case b] Direct equalisation** According to Figure 67, the idea is to use an equaliser capable to simultaneously perform equalisation and coefficients updating. In some contexts this is known as an "implicit model" structure.
As in the previous case, the channel equation is:

\[ s_R(k) = \alpha_0 s_T(k) + \alpha_1 s_T(k - 1) + \alpha_2 s_T(k - 2) \]  

(55)

which corresponds to the following impulse response:

\[ h_c(k) = \alpha_0 \delta(k) + \alpha_1 \delta(k - 1) + \alpha_2 \delta(k - 2) \]  

(56)

Or, alternatively, to the transfer function:

\[ H_c(z) = \alpha_0 + \alpha_1 z^{-1} + \alpha_2 z^{-2} \]  

(57)

The ideal transfer function (time invariant) of the equaliser will be:

\[ H_{eq}(z) = \frac{1}{\alpha_0 + \alpha_1 z^{-1} + \alpha_2 z^{-2}} \]  

(58)

Then, the proposed time-variant equaliser equation, realized by a 3 tap AR filter, in time-domain is:

\[ y(k) = h_0 x(k) + h_1 y(k - 1) + h_2 y(k - 2) \]  

(59)
Notice that in this case the coefficients $h_0$, $h_1$ and $h_2$ are continuously updated, not only after the training sequences. This is the key difference between this case and the previous one, where an identification process is carried out during each training sequence.

So $h_0$, $h_1$ and $h_2$ are time-variant (adaptive) coefficients, and the final objective is to select the adequate updating algorithm in order to achieve the correct coefficients convergence. We will try a Kalman filter for this purpose in order to have assured the stability of the time variant AR equaliser. This stability is not assured when using a LMS algorithm.

The $h_0$, $h_1$ and $h_2$ are the coefficients obtained after the identification process during every training sequence.

### 10.2.2 Complex LMS Adaptive Filter

The equations for the well known adaptive LMS filter are:

\[
\begin{align*}
    y(k) &= \bar{h}^H(k) \bar{x}(k) \\
    e(k) &= d(k) - y(k) \\
    \bar{h}(k+1) &= \bar{h}(k) + \bar{e}^*(k) \mu \bar{x}(k),
\end{align*}
\]

being $\bar{x}(k)$ the information vector, $\bar{h}(k)$ the filter coefficients vector, $d(k)$ the reference signal (scalar), and $y(k)$ the filter output (scalar).

### 10.2.3 Complex Adaptive Kalman Filter

Kalman and Bucy proposed an extremely powerful recursive estimation technique, commonly described as Kalman filtering [2]. The use of basic Kalman filtering presupposes that the system under consideration can be described by a set of linear difference equations. Given this description, optimal estimates of the states can be obtained from noisy observations in a recursive manner.

In general the optimum transversal equaliser for a channel (FIR or IIR), assuming the minimisation of a quadratic criterion, is subject to the constraint that the impulse response be finite (FIR), causal and stable. The motivation for considering an equaliser with an IIR structure was to try and overcome the limitations of conventional FIR equalisers structures. In many applications FIR solutions have been found to be perfectly adequate, and they are generally preferable since they are unconditionally stable. However, these FIR filters suffer from indeterminate order when are required to model transfer function poles, especially poles close to the unit circle. The obvious alternative has been IIR filters, but these are not unconditionally stable. The Wiener filter (Kalman filter is a recursive implementation of Wiener optimal solution) offers advantages over the conventional linear (FIR) equaliser, in terms of the required
equaliser order for achieving the same level of performance in minimum phase channels. However, the realisation of such filters would require a spectral factorisation. This would present a major difficulty, since it is often an ill conditioned problem. Moreover, conventional Kalman filters have the requirement of the a priori knowledge of a transition matrix of the system to be identified or equalised.

Godard [3, 4] showed that it is possible an improved application of Kalman filtering techniques to adaptive equalisation by appropriately selecting the states to configure the algorithm used in each concrete problem. The algorithm proposed by Godard was based in the use of a Kalman filter to perform the coefficients update of a IIR (or FIR) adaptive filter. In the adopted solution Godard selects the tap weight vector of the equaliser as his state vector and assumes the state transition matrix to be the identity matrix. So, it’s not necessary the a priori knowledge of a transition matrix. Moreover, the result is a fast equaliser structure relaxing the constraint that the filter have to be FIR.

The discrete-time Kalman filter results in a linear and recursive system whose implementation is well suited for a digital computer.

The equations of the adaptive Kalman filter are (the structure corresponds to Figure 68):

\[
\begin{align*}
g(k) &= \bar{h}^H(k) \bar{x}(k) \\
e(k) &= d(k) - y(k) \\
\bar{h}(k+1) &= \bar{h}(k) + \bar{x}^\ast(k) \bar{g}(k) \\
\bar{g}(k) &= \frac{K(k-1) \bar{x}(k)}{\bar{x}^H(k) K(k-1) \bar{x}(k) + Q_M} \\
K(k) &= K(k-1) - \bar{g}(k) \bar{x}^H(k) K(k-1) + Q_p,
\end{align*}
\]

being: \(\bar{x}(k)\) the information vector, \(\bar{h}(k)\) the filter coefficients vector, \(d(k)\) the reference signal (scalar), and \(y(k)\) the filter output (scalar).

10.2.4 AR Equaliser with Kalman-Updating. Analysis and Matlab Simulations

We choose the implementation of an adaptive AR equaliser making the coefficients update by means of a Kalman algorithm.

The performance results for 64-QAM and QPSK modulations are shown in the following figures, which are not comparable among them. The simulation objectives have been to test the adaptive Kalman filter performance in front of 3-ray multipath (third order filters). Because taps are settled, as exposed above, at times delayed 1 and 2 times the symbol time \((T_s, 2T_s)\), actual delay profiles of simulated multipaths depend on the selected symbol rate.

10.2.5 Programming Issues

The fast Kalman algorithm has a complexity proportional to \(N\) (filter order) and offers similar convergence times than the conventionally-implemented Kalman algorithm whose complexity is proportional to \(N^2\). For a linear equaliser, the fast Kalman algorithm requires a number of operations per iteration only about 5 times those of the simple gradient algorithm.

These figures are only indicative, depending on several factors. One of the main aspects which affects the DSP performance is the complexity of the assembler code produced by the C-compiler. Some DSP devices whose assembler software requires a high number of basic operations to run a certain filter may compensate this drawback from its hardware, i.e, offering higher floating point operations per second and per instruction.

Once the code is arranged in the specific memory it have to be hand-optimised in order to improve some limitations of the automatically generated code. Furthermore, most of current DSP devices allow pipe-line programming, thus requiring additional software optimisation in order to arrange the assembler program to the parallelisation procedures.
Figure 69: BER for 64-QAM, bit rate 120 Mbit/s, $f_d = 4500$. Rice factor $K = 15$ dB (x–x) and $K = 20$ dB (*–*). Training sequence length: 10000.

Figure 70: Received constellation (64-QAM, bit rate 120 Mbit/s, $f_d = 4500$. Rice factor $K = 20$ dB, $E_b/N_0 = 20$ dB).
Figure 71: Equalised constellation (64-QAM, bit rate 120 Mbit/s, $f_d = 4500$. Rice factor $K = 20$ dB, $E_b/N_0 = 20$ dB).

Figure 72: BER for QPSK, bit rate 32 Mbit/s, $f_d = 4500$, Rice factor $K = 10$ dB (x–x) and $K = 15$ dB (*–*).
Figure 73: Received constellation (QPSK, bit rate 32 Mbit/s, $f_d = 4500$. Rice factor $K = 15$ dB, $E_b/N_0 = 20$ dB).

Figure 74: Equalised constellation (QPSK, bit rate 32 Mbit/s, $f_d = 4500$. Rice factor $K = 15$ dB, $E_b/N_0 = 20$ dB).
Figure 75: BER for QPSK, bit rate 8 Mbit/s, $f_d = 4500$, Rice factor $K = 10$ dB (x=x) and $K = 15$ dB (*-*).

Figure 76: Received constellation (QPSK, bit rate 8 Mbit/s, $f_d = 4500$, Rice factor $K = 15$ dB, $E_b/N_0 = 20$ dB).
Figure 77: Equalised constellation (QPSK, bit rate 8 Mbit/s, $f_d = 4500$. Rice factor $K = 15$ dB, $E_b/N_0 = 20$ dB).
Considering a 300 MHz \( \frac{300}{10^6} = \frac{300}{10^9} \text{ ns} \) clock DSP, the maximum possible number of instruction cycles within a symbol duration are:

<table>
<thead>
<tr>
<th>Symbol time</th>
<th>Considering a 300 MHz ( \frac{300}{10^6} = \frac{300}{10^9} \text{ ns} ) clock DSP, the maximum possible number of instruction cycles within a symbol duration are:</th>
</tr>
</thead>
<tbody>
<tr>
<td>( T_s = 50 \text{ ns} ) (20 MBaud/s)</td>
<td>( \frac{50 \text{ ns}}{\frac{300}{10^6} \text{ ns}} = 15 )</td>
</tr>
<tr>
<td>( T_s = 62.5 \text{ ns} ) (16 MBaud/s)</td>
<td>( \frac{62.5 \text{ ns}}{\frac{300}{10^6} \text{ ns}} = 18.75 \rightarrow 19 )</td>
</tr>
<tr>
<td>( T_s = 250 \text{ ns} ) (4 Mbaud/s)</td>
<td>( \frac{250 \text{ ns}}{\frac{300}{10^6} \text{ ns}} = 75 )</td>
</tr>
</tbody>
</table>

Table 27: Maximum number of DSP instruction cycles.

10.3 Implementation on a TMS320C6711 Digital Signal Processor

10.3.1 Description of the DSP Benchmark

The selected processor for implementing the equaliser algorithms has been a TMS320C6711. With performance of up to 1200 million floating-point operations per second (MFLOPS) at a clock rate of 200 MHz or 1350 MFLOPS at a clock rate of 250 MHz (for 6711D), the C6711C/C6711D device also offers cost-effective solutions to high-performance DSP programming challenges. The C6711C/C6711D DSP also possesses the operational flexibility of high-speed controllers and the numerical capability of array processors. This processor has 32 general-purpose registers of 32-bit word length and eight highly independent functional units. The eight functional units provide four floating-/fixed-point ALUs, two fixed-point ALUs, and two floating-/fixed-point multipliers. The C6711C/C6711D can produce two MACs per cycle for a total of 400 MMACS. Figures 78 and 79 show the components for the benchmark:

- SignalWare AED-101 AD/DA board
- Texas Instruments TMSC6711 Digital Signal Processor
- Texas Instruments Code Composer C compiler

10.3.2 C-Program Structure

The developed program consists in a main file and a sort of callable routines for doing the most important algebraic operation: dot product of complex vectors, complex vectors multiplication, complex...
10.3.3 Achieving Improvement by Using the Optimised DSPLIB Library

The Texas Instruments C67x DSPLIB [5] is an hand-coded assembly-optimized DSP library of functions for C programmers using TMS320C67x devices. It includes C-callable, assembly-optimized general-purpose signal processing routines. These routines are typically used in computational intensive applications where optimal execution speed is critical. By using these routines it is possible to achieve execution speeds considerably faster than equivalent code written in standard ANSI C language.

Some of commonly used DSP routines that include the DSPLIB are listed in Figures 80 and 81.

10.3.4 DSP Performance Results

Table 28 shows the number of required instruction cycles for running a hand-optimised adaptive Kalman filter and those for the same filter with a posterior debugging by using the DSPLIB. Results are compared with the required instruction cycles for programming an LMS algorithm with the same programming tools.

The measured computational cost (using the optimal library DSPLIB) for programming the following minimum equation (without coefficient’s updating)

\[ c = a_1b_1 + a_2b_2 + a_3b_3 \]  \hspace{1cm} (68)

is 90 instruction cycles (being \( a_i \), \( b_i \) and \( c \) complex numbers).

Regarding a 3-tap adaptive Kalman filter program optimised by means of the DSPLIB library, the
maximum allowable symbol rate with a 300 MHz DSP device can be computed as:

\[
\frac{\text{symbol time} \ [\text{ns}]}{\frac{10^3 \ [\text{ns}]}{\text{symbol}}}} = 2051
\]  \(\text{(69)}\)

so that

\[
\text{symbol rate} = 1/6836 = 146\text{kBaud/s}.
\]  \(\text{(70)}\)

This value is far from the symbol-rates specified in IEEE802.16 standard. Even assuming an optimistic expectative to reduce the instruction cycles to the half of the abovementioned figures, the resulting tolerable symbol-rate is extremely low regarding our application.

A simple DSP-based LMS algorithm, once DSPLIB optimised, is only capable to approximately tolerate

\[
\frac{\text{symbol time} \ [\text{ns}]}{\frac{10^3 \ [\text{ns}]}{\text{symbol}}}} = 624
\]  \(\text{(71)}\)

so that

\[
\text{symbol rate} = 1/2080 = 480\text{kBaud/s}.
\]  \(\text{(72)}\)
Thus also becoming not enough for IEEE802.16. Obviously, once obtained these results, it has not sense to evaluate improved alternatives (i.e., extended Kalman filters, or filter capable to automatically updating the noise covariance matrices).

The objective of this analysis was not only to assess the feasibility of a DSP implementation of a practical equalizer structure, but also, more important, in case of infeasibility, to have figures indicating how far the DSP capability to support the equaliser is, and consequently to decide if the further work has to be guided to improve algorithms or to search for alternative technologies. As a consequence of the analysis presented in this section, we may conclude that a DSP device has not enough capability to support symbol rates specified in IEEE802.16, thus being necessary to develop further research aiming to evaluate the possibilities of alternative devices, such as new FPGAs.

### Table 28: Measured instruction cycles.

<table>
<thead>
<tr>
<th>order</th>
<th>LMS (reference algorithm)</th>
<th>Adaptive Kalman</th>
<th>Adaptive Kalman (using DSPLIB)</th>
</tr>
</thead>
<tbody>
<tr>
<td>3</td>
<td>624</td>
<td>6613</td>
<td>2051</td>
</tr>
<tr>
<td>5</td>
<td>1220</td>
<td>13175</td>
<td>3849</td>
</tr>
<tr>
<td>10</td>
<td>1424</td>
<td>39240</td>
<td>11116</td>
</tr>
</tbody>
</table>
Appendix: C-Code of the Implemented Adaptive Kalman Filter

```c
#include "parametros.h"
#include "funciones.h"
#include "c6x11dsk.h"
#include <c6x.h>
#include <stdlib.h>
#include <ctime>
#include <math.h>

float s_R[Tf][2];
float s_T[Tf][2];
float error_T[Tf][2];
float w[N][2];
int Qm[2]={1,0};
float Qp[N][N][2];
float K[N][N][2];
float u[N][2];
float d[2];
float y[2];
float error[2];
float g[N][2];
float vec[N][2];
float m[2];
float n[2];
float v[2];
float z[2];
float im, re;
float u_conj[N][2];
float constellation[S][2];
int s_aleatoria[Tf];
float mod_error[Tf];

void main()
{
    int i, j, t;
    for(i=0; i<Tf; i++)
        for(j=0; j<2; j++)
        {
            // s_R[i][j] = 0.0;
            // s_T[i][j] = 0.0;
            error_T[i][j] = 0.0;
        }
    gen_const(constellation);
    gen_mensaje(s_aleatoria);
    gen_modulacion(s_aleatoria, s_T, constellation);
    anade_ruido(s_T, s_R);
    for(i=0; i<N; i++)
        for(j=0; j<2; j++)
        {
            w[i][j] = 0.0;
        }
    for(i=0; i<N; i++)
        for(j=0; j<N; j++)
        for(t=0; t<2; t++)
            K[i][j][t] = 0.0;
    for(i=0; i<N; i++)
```
for (j=0; j<N; j++)
{
  if (i==j)
  {
    Qp[i][j][0] = 1.0;
    Qp[i][j][1] = 0.0;
  }
  else
  {
    Qp[i][j][0] = 0.0;
    Qp[i][j][1] = 0.0;
  }
}
for (t=N-1; t<Tf; t++)
{
  j=0;
  for (i=t; i>t-N; i--)
  {
    u[j][0] = s_R[i][0];
    u[j][1] = s_R[i][1];
    u_conj[j][0] = s_R[i][0];
    u_conj[j][1] = -s_R[i][1];
    j++;
  }
  d[0] = s_T[t][0];
  d[1] = s_T[t][1];
  conjCVdotCV(u_conj, w, y);
  error[0] = d[0] - y[0];
  error_T[t][0] = error[0];
  error_T[t][1] = error[1];
  CMxCV(K, u, vec, 0);
  //vec_mul
  conjCVdotCV(vec, u_conj, n);
  for (i=0; i<N; i++)
  {
    re = n[0] + Qm[0];
    im = n[1] + Qm[1];
    CSdivCS(vec[i][0], vec[i][1], re, im, z);
    g[i][0] = z[0];
    g[i][1] = z[1];
  }
  conjCVdotCV(g, u_conj, m);
  for (i=0; i<N; i++)
  {
    for (j=0; j<N; j++)
    {
      CSxCS(m[0], m[1], K[i][j][0], K[i][j][0], v);
      K[i][j][0] = K[i][j][0] + Qp[i][j][0] - v[0];
      K[i][j][1] = K[i][j][1] + Qp[i][j][1] - v[1];
    }
    CSxCS(error[0], -error[1], g[i][0], g[i][1], z);
    w[i][0] += z[0];
    w[i][1] += z[1];
  }
for(i=0;i<Tf;i++)
{
    re = error_T[i][0] * error_T[i][0];
    im = error_T[i][1] * error_T[i][1];
    mod_error[i] = re+im;
}
while(1);

#include <stdio.h>
#include <math.h>
#include <stdlib.h>
#include <ctime>
#include "QAM.h"
#include "parametros.h"

void gen_const (float (*constel)[2])
{
    int i,j;
    int m,n,t,r,x=0,z;
    int bigLen;
    int smallLen;
    n = (int)floor(sqrt(2*S));
    m = (int)ceil(sqrt(S));
    z = (int)floor(sqrt(S));
    r = (int)sqrt(S);
    if(S == 2)
    {
        constel[0][0] = -1.0;
        constel[0][1] = 0.0;
        constel[1][0] = 1.0;
        constel[1][1] = 0.0;
    }
    else
    {
        if(z == m)//Square constellation
        {
            for(i=1;i<=r-1;i=i+2)
            {
                for(j=1;j<=r-1;j=j+2)
                {
                    constel[x][0]=(float)i;
                    constel[x][1]=(float)j;
                    x++;
                }
            }
        }
        else if(S==8)
        {
            constel[0][0] = 1.0;
            constel[0][1] = 1.0;
            constel[1][0] = 3.0;
            constel[1][1] = 1.0;
        }
        else //Cross constellation
        {
smallLen = 1;
for (bigLen = m; bigLen < n; bigLen++)
{
    while (bigLen > sqrt(S + 4 * (smallLen * smallLen)))
        smallLen++;
    if (bigLen == sqrt(S + 4 * (smallLen * smallLen)))
        break;
    else
    {
        bigLen++;
        smallLen = 1;
    }
} 
t=0;
for (i=1; i<=bigLen-1; i=i+2)
for (j=1; j<=bigLen-1; j=j+2)
if (i < bigLen - (2 * smallLen -1) || j < bigLen - (2 * smallLen -1))
{
    constel[t][0] = (float)i;
    constel[t][1] = (float)j;
    t++;
}
for (i=0; i<S/4; i++)
{
    constel[(S/4)+i][0] = constel[i][0];
    constel[(S/4)+i][1] = -constel[i][1];
}
for (i=0; i<S/4; i++)
{
    constel[2*(S/4)+i][0] = -constel[i][0];
    constel[2*(S/4)+i][1] = -constel[i][1];
}
for (i=0; i<S/4; i++)
{
    constel[3*(S/4)+i][0] = -constel[i][0];
    constel[3*(S/4)+i][1] = constel[i][1];
}
}
void gen_modulacion(int *mensj, float (*s_mod)[2], float (*constel)[2])
{
    int i;
    for (i=0; i<Tf; i++)
    {
        s_mod[i][0] = constel[mensj[i]][0];
        s_mod[i][1] = constel[mensj[i]][1];
    }
}
void gen_mensaje(int *mensj)
{
    int i, j;
    srand(time(NULL));
    for (i=0; i<Tf; i++)
    {
j = rand() % S;
mensj[i] = j;
}
}
void anade_ruido(float (*s_modulada)[2], float (*s_mod_ruido)[2])
{
    int i, j, t;
srand(time(NULL));
    for(i=0; i<Tf; i++)
    {
        j = (rand() % 7) - 4;
        t = (rand() % 7) - 4;
        s_mod_ruido[i][0] = s_modulada[i][0] + j * ampli_ruido;
        s_mod_ruido[i][1] = s_modulada[i][1] + t * ampli_ruido;
    }
}
#include "parametros.h"
#include "funciones.h"
#include <stdio.h>
#include <c6x.h>
void CMxCV(float (*x)[N][2], float (*y)[2], float (*vec)[2], int t) //t nos dice
    si es el vector normal o traspuesto 0: normal, 1: traspuesto
{
    int i, j;
    float sum1, sum2;
    float r[2];
    if(t==0)
    {
        for(i=0; i<N; i++)
        {
            sum1 = 0;
            sum2 = 0;
            for(j=0; j<N; j++)
            {
                CSxCS(y[j][0], y[j][1], x[i][j][0], x[i][j][1], r);
                sum1 += r[0];
                sum2 += r[1];
            }
            vec[i][0] = sum1;
            vec[i][1] = sum2;
        }
    }
    else
    {
        for(i=0; i<N; i++)
        {
            sum1 = 0;
            sum2 = 0;
            for(j=0; j<N; j++)
            {
                CSxCS(y[j][0], y[j][1], x[j][i][0], x[j][i][1], r);
                sum1 += r[0];
                sum2 += r[1];
            }
            vec[i][0] = sum1;
        }
    }
vec[i][1]=sum2;
}
}
void conjCVdotCV(float (*x)[2], float (*y)[2], float *mul)
{
    int i;
    float res[2];
    mul[0]=0;
    mul[1]=0;
    for(i=0;i<N;i++)
    {
        CSxCS(x[i][0],x[i][1],y[i][0],y[i][1],res);
        mul[0] += res[0];
        mul[1] += res[1];
    }
}
void CSxCS(float re1,float im1, float re2, float im2, float *r)
{
    float real, imag;
    real=re1*re2-im1*im2;
    imag=re1*im2+im1*re2;
    r[0]=real;
    r[1]=imag;
}
void CSdivCS(float rel,float iml, float re2, float im2, float *r)
{
    float n;
    float num[2];
    float den[2];
    float a,b;
    n = -im2;
    CSxCS(rel,iml, re2, n, num);
    CSxCS(re2,im2, re2, n, den);
    a = num[0]/(den[0]+den[1]);
    b = num[1]/(den[0]+den[1]);
    r[0]=a;
    r[1]=b;
11 Conclusions

In BWA systems, the wireless link is likely to be the bottleneck in any end-to-end mobile wireless broadband system. Therefore, the improvements in wireless link performance, in terms of spectral efficiency, channel utilization and QoS capability, directly translate to the overall improvement of the end-to-end system.

The CAPANINA project aims at the delivery of broadband services through a collective terminal able to serve all the passengers on board of a train via a HAP system. Passengers travelling on the train are equipped with WLAN terminals and access the network via a collective terminal through the HAP and a gateway located in the ground station, requesting broadband multimedia services from various content providers. In this scenario HAPs are used as a collective access network, providing BWA services to the users.

The aim of this report is to propose advanced signal processing techniques especially tailored to cope with the time variant response of the HAP-to-train / train-to-HAP wireless communication channel with reference, in particular, to the IEEE 802.16 Standard (both in the Single Carrier and OFDM flavors), which is suited to the HAP environment. In particular, adaptive modulation and coding, highly efficient variable-rate, variable-length channel codes, spatial diversity, polarization diversity coupled with trellis-coded modulation and space-time codes, adaptive channel estimation based on a Kalman filtering approach are analyzed in this report, as some of the most powerful techniques to provide QoS support in HAP-based broadband communications.

Simulation results obtained over Rice/Rayleigh fading channels showed the effectiveness of the proposed approaches to achieve target QoS levels, often exceeding the performance of other common modulation and coding schemes, even those proposed in the IEEE 802.16 standard. Indeed, the adaptive selection of the coding and modulation scheme has been shown to be able to increase the bandwidth efficiency with respect to fixed allocation schemes. If standard IEEE 802.16 coding formats are applied, the average information rate is expected to vary between 25 to 75 Mbit/s. Furthermore, if advanced coding solutions are considered, such as serial or parallel concatenated convolutional codes, 1.4 to 2 dB gain are expected with respect to the standard solutions, given the same code length. It is evident that such a choice would represent a modification to be applied to the selected communications standard.

Finally, a Kalman-filtering based channel estimator and equalizer appears able to ensure quasi-ideal working conditions in a few-multipath channel (or flat fading channel), even for high bit rates, high-level modulations and medium/high Doppler frequency shift. This is a very promising approach for improving the performance of the abovementioned coding and modulation schemes toward the ideal bound obtained over the AWGN channel.

Synchronization algorithms are able to perform Doppler recovery are not expected to be a critical issue with single carrier modulations. However, since the IEEE 802.16 standard has not been expressly intended for high-speed terminals, this topic will be object of further investigations within WP2.3, with particular attention to the performance of the selected communication standard.
References


