Adaptive coding and modulation for satellite broadband networks: From theory to practice

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SUMMARY

This paper presents the detailed design and the key system performance results of a comprehensive laboratory demonstrator for a broadband Ka-band multi-beam satellite system exploiting the new DVB-S2 standard with adaptive coding and modulation (ACM). This complete demonstrator allows in-depth verification and optimization of the ACM techniques applied to large satellite broadband networks, as well as complementing and confirming the more theoretical or simulation-based findings published so far. It is demonstrated that few ACM configurations (in terms of modulation and coding) are able to efficiently cope with a typical Ka-band multi-beam satellite system with negligible capacity loss. It is also demonstrated that the exploitation of ACM thresholds with hysteresis represents the most reliable way to adapt the physical layer configuration to the spatial and time variability of the channel conditions while avoiding too many physical layer configuration changes. Simple ACM adaptation techniques, readily implementable over large-scale networks, are shown to perform very well, fulfilling the target packet-error rate requirements even in the presence of deep fading conditions. The impact of carrier phase noise and satellite nonlinearity has also been measured. Copyright © 2009 John Wiley & Sons, Ltd.

KEY WORDS: Broadband satellite communications; fading mitigation; adaptive coding and modulation; DVB-S2; DVB-RCS; multi-beam satellite networks; ka-band

1. INTRODUCTION

The successful deployment of broadband interactive services via satellite as a complement to the terrestrial services requires achieving user experience and servicing cost comparable to the terrestrial networks. Although initially broadband interactive services have been provided at Ku-band using conventional fixed satellite services, it quickly became evident that a move to

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Ka-band was needed (see [1,2]), where more bandwidth is available and where dedicated multi-beam payloads with high-frequency reuse can be developed. It is clear that interactive services, aiming at the provision of different data to the different user terminals, can benefit significantly from small satellite beams to increase the satellite performance in terms of EIRP and G/T, as well as from the increased frequency reuse factor, which today allows achieving unprecedented throughputs from a single satellite in the order of 100 Gbps or even more [3,4]. However, this great Ka-band multi-beam satellite payload potential requires exploiting the state-of-the-art techniques in terms of physical layer, fading mitigation techniques (FMT) and radio resource management.

The approval in 2004 of the DVB-S2 standard [5–7] opened up a major breakthrough in the commercial standards for satellite broadcasting, interactive and professional applications as it made possible to achieve, at low cost, near Shannon bound FEC performance combined with a wide range of modulation and coding schemes. Furthermore, the built-in DVB-S2 support of adaptive coding and modulation (ACM) in addition to constant coding and modulation (CCM) and variable coding and modulation allowed the provision of interactive broadband services to large user communities at affordable prices. ACM makes it possible to operate Ka-band multi-beam systems with reduced link margins, optimizing the physical layer configuration for every user at all times, thanks to a distributed adaptation of the optimum physical layer configuration. As a consequence, the system performance is no longer dominated by the worst-case link conditions, but rather by the average link conditions of the system. Thus, the rare deep fading events have no practical impact on the overall system capacity, but are just causing a reduction of the individual physical layer link peak data rate. This effect can be counteracted by proper resource management techniques [8]. A similar approach can be adopted for the return link, although the current DVB-RCS standard [6] only allows for adaptive coding (AC) and dynamic rate adaptation (DRA) implementation. Previous work has been focusing on the system performance analysis in the presence of ACM for the forward [9] and the return link [10] as well as in-depth investigations of the ways to practically implement ACM [11].

This paper describes a comprehensive DVB-S2/RCS ACM laboratory demonstrator developed by a team of companies and research institutions with ESA ARTES R&D funding [12] to demonstrate the ACM capabilities in a realistic Ka-band multi-beam satellite system environment. The demonstrator allows experimental verification of the expected ACM advantages and acquires an in-depth understanding of the design and optimization of an ACM-based broadband satellite network. Although the demonstrator covers both the forward and the return link, emphasis is given to the forward link ACM implementation aspects. It is felt that the demonstrator described in this paper represents a good complement to the standardization and analytical work mentioned in Section 1, confirming through laboratory experiments the high ACM potential in a realistic satellite system environment. The rest of the paper is organized as follows: following this introduction, in Section 2 a satellite multi-beam system reference scenario for the demonstrator is introduced and key system performance is derived. In Section 3 the ACM-demonstrator key elements design is illustrated. In Section 4 the main demonstrator experimental performance results are reported. Finally, in Section 5 the conclusions are drawn.

\[^{1}\text{With ACM worst-case fading link margin is in principle not required as the waveform is adapting to the individual user current reception conditions.}\]
2. REFERENCE SYSTEM SCENARIO

The objective of the system scenario definition is to provide a reference satellite system scenario for a realistic evaluation of the ACM performance, to be used in particular for the system tests of the ACM system demonstrator. The proposed system targets a typical multi-beam Ka-band satellite system, with DVB-S2 and ACM on the forward link and DVB-RCS with FMT on the return link. The mission is to offer a realistic broadband communication system that could be operational in the mid-term time frame.

System features: The main features of the reference system are:

- The reference system is based on a medium to large satellite platform.
- The hub-to-spoke network architecture is centralized, characterized by a minimum number of gateways (7) with a fixed connectivity to groups of beams.
- Ka-band operations for both the feeder and user link.

- Gateways’ main characteristics:
  - 250 W transmit high power amplifier (HPA) with 2 GHz bandwidth.
  - Antenna size: 8.1 m.
  - EIRP: 80 dBW per HPA.
  - Clear sky receive G/T: 37.4 dB/K.

- Satellite main characteristics:
  - 100 user beams of 250 MHz each, with frequency reuse 1:4.
  - Platform: Alphabus.
  - Receive reflector size: 1.76 m.
  - Transmit reflector size: 2.64 m.
  - G/T: 23.2 dB/K.
  - EIRP: 66 dBW per 250 MHz TWTA on the forward link.

- Two types of user terminals are being considered:
  - Consumer: terminals with 65 cm antenna, transmitting at 1 W, 512 kbaud symbol rate.
  - Small and medium enterprises (SME): terminals with 90 cm antenna, transmitting at 2 W, 2048 kbaud symbol rate.

Applications: The mission consists of transparent broadband access for applications such as Internet. The typical applications envisaged are web browsing, E-mail, audio/video streaming, voice over IP, video conferencing and LAN interconnection. The accessible market targets 3% of the population. The users consist of 85% consumers and 15% SME. The total number of connected user terminals per beam is typically 1000, possibly up to 2000 [13].

System coverage and frequency plan: The main features of the reference satellite network are summarized in the following:

- Satellite orbital position: 33° east.
- Frequency band: Ka-band.
- Number of user beams: 100.
- Beam size: 0.4°.
- Polarization: circular.
- Bandwidth: 250 MHz per beam with four-colour scheme.
- Total bandwidth: 25 GHz (100 × 250 MHz).
- Connectivity: each gateway is connected with up to 16 pre-defined beams (4GHz/gateway).

The target extended European coverage is shown in Figure 1. The footprints of the 100 user beams for the targeted coverage are shown in Figure 2.

**Frequency plan:** For the gateway links a 2 GHz bandwidth is exploited in both polarizations. More precisely, the forward uplink is in the 27.5–29.5 GHz band, while the reverse downlink occupies the 17.7–19.7 GHz band. The 0.5 GHz user link spectrum is reused in both polarizations with a reverse uplink band from 29.5 to 30.0 GHz and forward downlink from 19.7 to 20.2 GHz.

The associated user link frequency reuse scheme padding is represented in Figure 3. Note that the LHCP and RHCP polarizations are inverted between the uplink and the downlink.

![Figure 1. Targeted user coverage.](image-url)
The four-colour scheme defined in this document is based on a pre-defined pattern (see paragraph on user links). The distance between beam centres of the same colour is 0.8° (typically 500 km) for the nearest six beams.

**Connectivity:** For the link budget simulations, seven gateways have been arbitrarily distributed on the main areas, located on existing teleports (Table I).

The gateway feeder beams are connected to a fixed number of pre-defined user beams. Each gateway serves a maximum of 16 user beams of 250 MHz, using twice the two 2 GHz/polarization allocated feeder link band. The gateway to beam connectivity used for the simulations is presented in Figure 4.
The traffic distribution has been computed according to the methodology described in [13]. Figure 5 graphically presents the aggregated user traffic per beam. It is apparent that the beam loading is very uneven throughout the coverage with few very hot beams and many cold beams. For simplicity, the reference system and simulations assume that spot beams are evenly sized, each beam being loaded on the forward link with all DVB-S2 carriers active and 70% loaded on the return link. Customized architectures can be envisaged for operational satellites where different EIRP resources are allocated to zones with higher user density, in a static or programmable basis. Another technique, when a spot beam is saturated, consists of allocating the users at the edge of the beam to neighbouring beams.

**Air interface:**

*Forward link:* The 250 MHz beam band is divided into four DVB-S2 carriers of 45 Mbaud each. Using four carriers per beam gives a reasonable granularity while keeping a centralized architecture. The key physical layer parameters are the following:

- Standard: DVB-S2.
- FMT: ACM.
- Number of carriers per beam: 4.
- Carrier symbol rate: 45 Mbaud.
- Square-root raised-cosine roll-off factor: 0.25.
Figure 5. Traffic aggregation per beam.

Figure 6. CDF of forward link $C/N$, $C/I$ and $C/(N+I)$ in dB.
MODCODs: the 28 MODCODs of the DVB-S2 standard are available, of which 22 are initially retained (see Figure 7).

Reverse link: The reverse link is based on standard DVB-RCS with the following configuration:

- Standard: DVB-RCS.
- Number of carriers per beam: 324 for consumers and 81 for SME users.
- Carrier peak symbol rates:
  - Consumers: 512 kbaud.
  - SME: 2048 kbaud.
- Square-root raised-cosine roll-off factor: 0.35.
- Modulation: QPSK.
- Coding: turbo coding with code rates 1/3, 2/5, 1/2, 2/3, 3/4, 4/5, 6/7.
- FMT technique: AC+DRA.

Service availability: The targeted service availability ranges from 99.5 to 99.9%. The PER targets are based on quasi-error-free (QEF) links, which are usually considered for DVB-S/S2 and DVB-RCS links. At the transport stream level, the associated MPEG packet-error rate target is equal to or better than:

- Forward link: $10^{-7}$, which corresponds to approximately 0.5 packet errors per hour on average for a user forward link rate of 2048 kbps.
- Reverse link: $10^{-6}$, which corresponds to approximately one packet error per hour on average for a user reverse link rate of 512 kbps.

System simulations: Multi-dimensional link budget simulations have been performed based on the reference system, to anticipate the performance that would be obtained in a real system. The simulation tool described in [14] takes into account the geographical distribution of the
users and the atmospheric conditions. The simulations have been performed separately for the two types of terminals (consumer and SME). The results obtained by simulation will be compared with the performance measured on the ACM modem test bed (see Section 4). Following the findings of Reference [13], the gateways transmit at the maximum power level, whatever the atmospheric conditions on the uplink (no uplink power control).

In Figure 6 the forward link $C/N$, $C/I$ and $C/(N+I)$ cumulative distribution function (CDF) is plotted for both types of terminals.

It is remarked that consumers and SME $(C/I)_{\text{down}}$ distributions are almost identical, while the resulting total $C/(N+I)$ is approximately 1 dB higher for the SME users compared with consumers. $(C/N)_{\text{down}}$ is generally equal to or better than $(C/I)_{\text{down}}$, which means that the satellite EIRP has been correctly sized.

The time availability corresponds to the $C/(N+I)$ values that are above the ACM minimum modulation/coding threshold ($-0.86$ dB). The time availability CDF across the users' distribution is summarized in Table II.

The previous availability results have been obtained assuming that the physical layer has the capability to continuously adapt to the variable SNIR over the coverage region and the time. In reality DVB-S2 allows for a discrete number of coding rates and modulation configurations (MODCODs). The DVB-S2 defines 28 MODCODs, but this large number of MODCODs complicates radio resource management and forward link scheduling. It is therefore desirable to reduce the number of MODCODs without sacrificing system throughput (spectral efficiency) [14]. From the FEC measurements performed during the initial phase of the project, the required $E_s/N_0$ values to achieve the target packet-error rate have been extracted (see column 4 in Table V). In addition to this, we add the transmitter (see Table III) and receiver (see Table IV) implementation losses, where an important component is the synchronization loss. The synchronization losses have also been experimentally derived based on modem simulation

<table>
<thead>
<tr>
<th>Availability in time (%)</th>
<th>Consumer coverage (%)</th>
<th>SME coverage (%)</th>
<th>Weighted average (%)</th>
</tr>
</thead>
<tbody>
<tr>
<td>98.00</td>
<td>99.93</td>
<td>99.93</td>
<td>99.93</td>
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<td>99.00</td>
<td>99.93</td>
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<td>99.50</td>
<td>99.93</td>
<td>99.93</td>
<td>99.93</td>
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<tr>
<td>99.80</td>
<td>99.89</td>
<td>99.93</td>
<td>99.90</td>
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<td>99.90</td>
<td>98.93</td>
<td>98.93</td>
<td>98.93</td>
</tr>
<tr>
<td>99.95</td>
<td>82.61</td>
<td>86.46</td>
<td>83.19</td>
</tr>
<tr>
<td>99.98</td>
<td>40.79</td>
<td>49.75</td>
<td>42.13</td>
</tr>
<tr>
<td>99.99</td>
<td>25.86</td>
<td>32.61</td>
<td>26.87</td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>Table III. Estimate of transmitter loss for some cases.</th>
</tr>
</thead>
<tbody>
<tr>
<td>Modulation</td>
</tr>
<tr>
<td>----------------</td>
</tr>
<tr>
<td>QPSK (1/2)</td>
</tr>
<tr>
<td>8-PSK (2/3)</td>
</tr>
<tr>
<td>16-APSK (3/4)</td>
</tr>
<tr>
<td>32-APSK (4/5)</td>
</tr>
</tbody>
</table>
results. We have assumed a simple, static pre-compensation technique in order to reduce the HPA nonlinearity loss [15].

Table V gives the effective decision SNIR thresholds for DVB-S2 MODCODs based on the actual demonstrator performance. The SNIR Δ steps between consecutive MODCODs are also indicated in the table. The results are graphically summarized in Figure 7.

From Figure 7 we can easily see that some MODCODs should not be used in a particular scenario. Let us consider the performance for the nonlinear channel. We can see that MODCOD 12 is clearly better than MODCODs 10 and 11, and similarly MODCOD 18 is better than MODCODs 16 and 17. Further, the difference between 22 and 23 and between 27 and 28 is small; therefore, we can remove one of them. One can perform similar exercises for the linear channel. The basic MODCOD sets based on this initial selection are given in bold in Table V and summarized in Table VI.

Note that the sets found are not unique and that there are other combinations with almost identical performance. However, it is important that the list of possible MODCODs is such that when one plots the $E_b/N_0$ versus spectral efficiency this should be a monotonic function in order for the ACM control mechanism to work.

In order to limit the scheduler complexity, it is interesting to aggregate the traffic for many terminals on a small number of MODCODs. When one considers the overall system performance, it is intuitively important to have reduced MODCOD granularity for the most probable $C/(N+I)$ values, i.e. the MODCODs corresponding to clear sky conditions for the different terminals. For faded conditions the MODCOD granularity is much less important, being a low-probability event. Thus, what is important is to have the most protected MODCODs to achieve the required system availability. In order to numerically perform this MODCOD subset optimization, we used global fading statistics, i.e. the $C/(N+I)$ distributions computed before (see Figure 6).

Figures 8 and 9 contain the MODCOD distribution, excluding the MODCODs previously removed. The professional users have a larger antenna and can thus support higher MODCODs. It is apparent that MODCODs more protected than 8-PSK $r = 2/3$ are used with very little probability, most likely just in the case of fading events. Thus, there is room for further MODCOD pruning.

We will now see that the size of the MODCOD family $M$ set affects the throughput. The subsets are formed by first selecting the lowest MODCOD ($j = 1$) in order to achieve the required link availability and then sequentially adding MODCODs by adding the one that maximizes the overall spectral efficiency $\eta_{TOT}$. This can be formalized by the following set of

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Table IV. Maximum estimated loss for the DVB-S2 receiver 25 Msp.s.

<table>
<thead>
<tr>
<th>Modulation</th>
<th>Linear estimated loss</th>
<th>Non-linear estimated loss</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>PER $= 10^{-4}$</td>
<td>PER $= 10^{-4}/10^{-7}$</td>
</tr>
<tr>
<td>QPSK</td>
<td>0.1/0.2</td>
<td>0.35/0.45</td>
</tr>
<tr>
<td>8-PSK</td>
<td>0.15/0.35</td>
<td>0.65/0.95</td>
</tr>
<tr>
<td>16-APSK</td>
<td>0.35/0.45</td>
<td>1.5/1.7</td>
</tr>
<tr>
<td>32-APSK</td>
<td>0.65/0.80</td>
<td>2.5/2.7</td>
</tr>
</tbody>
</table>

Note that for this assessment we used demodulator simulation results down to PER $= 10^{-4}$ and extrapolated the simulation results for PER $= 10^{-7}$.
Table V. MODCOD thresholds for DVB-S2 (linear and non-linear channel).

<table>
<thead>
<tr>
<th>MODCOD #</th>
<th>Modulation/Coding</th>
<th>Spect efficiency</th>
<th>Linear</th>
<th>Non-linear w/pre-compensation</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td></td>
<td></td>
<td>FEC-only AWGN</td>
<td>Required $E_s/N_0$ (dB) @ PER = $10^{-7}$</td>
</tr>
<tr>
<td>1</td>
<td>QPSK (1/4)</td>
<td>0.49</td>
<td>-1.79</td>
<td>-1.59</td>
</tr>
<tr>
<td>2</td>
<td>QPSK (1/3)</td>
<td>0.66</td>
<td>-0.68</td>
<td>-0.47</td>
</tr>
<tr>
<td>3</td>
<td>QPSK (2/5)</td>
<td>0.79</td>
<td>0.12</td>
<td>0.33</td>
</tr>
<tr>
<td>4</td>
<td>QPSK (1/2)</td>
<td>0.99</td>
<td>1.07</td>
<td>1.28</td>
</tr>
<tr>
<td>5</td>
<td>QPSK (3/5)</td>
<td>1.19</td>
<td>2.32</td>
<td>2.53</td>
</tr>
<tr>
<td>6</td>
<td>QPSK (2/3)</td>
<td>1.32</td>
<td>3.17</td>
<td>3.38</td>
</tr>
<tr>
<td>7</td>
<td>QPSK (3/4)</td>
<td>1.49</td>
<td>4.10</td>
<td>4.31</td>
</tr>
<tr>
<td>8</td>
<td>QPSK (4/5)</td>
<td>1.59</td>
<td>4.72</td>
<td>4.94</td>
</tr>
<tr>
<td>9</td>
<td>QPSK (5/6)</td>
<td>1.65</td>
<td>5.23</td>
<td>5.45</td>
</tr>
<tr>
<td>10</td>
<td>QPSK (8/9)</td>
<td>1.77</td>
<td>6.30</td>
<td>6.52</td>
</tr>
<tr>
<td>11</td>
<td>QPSK (9/10)</td>
<td>1.79</td>
<td>6.51</td>
<td>6.73</td>
</tr>
<tr>
<td>12</td>
<td>8-PSK (3/5)</td>
<td>1.78</td>
<td>5.62</td>
<td>5.99</td>
</tr>
<tr>
<td>13</td>
<td>8-PSK (2/3)</td>
<td>1.98</td>
<td>6.63</td>
<td>7.00</td>
</tr>
<tr>
<td>14</td>
<td>8-PSK (3/4)</td>
<td>2.23</td>
<td>7.97</td>
<td>8.34</td>
</tr>
<tr>
<td>15</td>
<td>8-PSK (5/6)</td>
<td>2.48</td>
<td>9.37</td>
<td>9.75</td>
</tr>
<tr>
<td>16</td>
<td>8-PSK (8/9)</td>
<td>2.65</td>
<td>10.79</td>
<td>11.19</td>
</tr>
<tr>
<td>17</td>
<td>8-PSK (9/10)</td>
<td>2.68</td>
<td>11.04</td>
<td>11.44</td>
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<tr>
<td>18</td>
<td>16-APSK (2/3)</td>
<td>2.64</td>
<td>9.02</td>
<td>9.50</td>
</tr>
<tr>
<td>19</td>
<td>16-APSK (3/4)</td>
<td>2.97</td>
<td>10.26</td>
<td>10.76</td>
</tr>
<tr>
<td>20</td>
<td>16-APSK (4/5)</td>
<td>3.17</td>
<td>11.06</td>
<td>11.56</td>
</tr>
<tr>
<td>21</td>
<td>16-APSK (5/6)</td>
<td>3.30</td>
<td>11.66</td>
<td>12.18</td>
</tr>
<tr>
<td>22</td>
<td>16-APSK (8/9)</td>
<td>3.52</td>
<td>12.98</td>
<td>13.51</td>
</tr>
<tr>
<td>23</td>
<td>16-APSK (9/10)</td>
<td>3.57</td>
<td>13.25</td>
<td>13.78</td>
</tr>
<tr>
<td>24</td>
<td>32-APSK (3/4)</td>
<td>3.70</td>
<td>13.00</td>
<td>13.88</td>
</tr>
<tr>
<td>25</td>
<td>32-APSK (4/5)</td>
<td>3.95</td>
<td>13.86</td>
<td>14.76</td>
</tr>
<tr>
<td>26</td>
<td>32-APSK (5/6)</td>
<td>4.12</td>
<td>14.40</td>
<td>15.32</td>
</tr>
<tr>
<td>27</td>
<td>32-APSK (8/9)</td>
<td>4.40</td>
<td>15.71</td>
<td>16.67</td>
</tr>
<tr>
<td>28</td>
<td>32-APSK (9/10)</td>
<td>4.45</td>
<td>16.07</td>
<td>17.04</td>
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</tbody>
</table>
Table VI. Basic MODCOD set.

<table>
<thead>
<tr>
<th>Channel</th>
<th>Basic MODCOD set</th>
</tr>
</thead>
<tbody>
<tr>
<td>Linear channel</td>
<td>Full set excluding {10, 11, 15, 16, 17, 23 and 28}</td>
</tr>
<tr>
<td>Non-linear channel</td>
<td>Full set excluding {10, 11, 16, 17, 23 and 28}</td>
</tr>
</tbody>
</table>

Figure 8. MODCOD distribution FL, maximum power (nonlinear channel).

Figure 9. MODCOD distribution FL, maximum power (linear channel).
equations:

$$\eta_{\text{TOT}} = \sum_{j=1}^{M} p_{S(j)} \eta_{S(j)}$$  \hspace{1cm} (1)$$

$$\hat{p}_{S(j)} = \frac{\sum_{S(j)}^{(j+1)-1} p_{j}}{\sum_{S(j)}^{(j+1)-1}}$$  \hspace{1cm} (2)$$

where $p_j$ and $\eta_j$ are the probability and the corresponding spectral efficiency of MODCOD $j$ and $S$ is the sorted set of possible MODCODs. $\hat{p}_{S(j)}$ gives the accumulated probability for a certain MODCOD for that particular set. As an example, if the set $S$ contains the MODCODs 1 and 14, $\hat{p}_{S(1)}$ will be the sum of all probabilities from MODCOD 1 up to 13, and since all these MODCODs will be mapped to MODCOD 1 in the reduced set, $\hat{p}_{S(2)}$ will be the accumulated probabilities from 14 and up.

By performing this optimization for the nonlinear case we get the results given in Table VII. The table indicates in which order the different MODCODs should be added to the MODCOD set. The resulting loss in spectral efficiency is shown in Figure 10. Even with only four MODCODs the spectral efficiency loss is no more than 2% for both the consumer and professional user, and when we increase to 8–10 MODCODs the loss is negligible. These results are based on global data that are averaged over a very long time and for all geographical locations.

<table>
<thead>
<tr>
<th>User</th>
<th>MODCOD selection order</th>
</tr>
</thead>
<tbody>
<tr>
<td>Consumer, non-linear</td>
<td>{11 4 18 15 19 12 20 5 13 4 7 2 9 6 8 3}</td>
</tr>
<tr>
<td>Professional, non-linear</td>
<td>{11 8 14 19 20 15 2 1 5 1 2 1 3 8 6 2 9 4 7}</td>
</tr>
</tbody>
</table>

Figure 10. Spectral efficiency loss for the different MODCOD subset size, nonlinear channel.
3. DEMONSTRATOR DESIGN

Figure 11 shows the overview of the ACM-demonstrator testbed architecture. The demonstrator testbed consists of the traffic simulator, the gateway, the channel emulator, the user terminal and the control and monitoring unit. The functional blocks are defined in Figure 12.

Traffic simulator: It consists of an FL traffic simulator and an RL traffic simulator. The FL traffic simulator is responsible for user traffic generation on the forward link and traffic reception and analysis on the return link. The RL traffic simulator is used for user traffic generation on the return link and reception and analysis of the traffic received by the user terminal on the forward link.

Gateway: The gateway consists of the NCC, DVB-S2 transmitter, DVB-RCS receiver and hub/switch. The NCC is responsible for providing the network control and management functions. The DVB-S2 transmitter performs the traffic routing, encapsulation, scheduling, FEC encoding, baseband modulation and up-conversion functions. The DVB-RCS receiver is a multi-carrier burst receiver (MCR) developed by Nera. The DVB-RCS receiver (MCR) used in the ACM-demonstrator testbed has the same functionality as the return link subsystem (called RLS hereafter) of Nera’s commercial DVB-RCS gateway solution (now owned by STM Norway). The hub/switch is used for providing the Ethernet interface for the user traffic.

Channel emulator: The FL and RL channel emulator performs the noise, interference, fading and HPA/TWTA nonlinearity simulations.

User terminal: The user terminal consists of two components: the DVB-S2 receiver developed and validated by Fraunhofer IIS during this project and the DVB-RCS transmitter, which is Nera’s (now STM’s) SatLink 1910-DVB-RCS user terminal. The output from the DVB-S2

Figure 11. ACM-demonstrator subsystems.
receiver is connected to the SatLink 1910 modulator, through the interface Nera I/O. The SatLink
1910 unit also performs the necessary handling and processing of higher-layer functions.

The overall ACM demonstrator has been integrated in two racks as shown in Figure 12. In
the following sections a summary description of the demonstrator subsystem elements’ design
aspects is provided.

3.1. ACM control

The ACM control and the method for selecting MODCODs are similar for the FL ACM and
RL AC+DRA, although RL specificities described in [11] are not further developed here. In
principle, the ACM control only determines the optimum MODCOD based on the measured
$C/(N+I)$. However, the physical layer configuration and the performance are quite different for
the FL and RL links and also the channel conditions are different, but this will only affect the
parameters used for ACM control such as margin and hysteresis, and algorithms used for
channel estimate post-processing.

The implementation of the ACM control in a real system needs to be simple since the GW
will have to manage a very large set of terminals, but yet effective in order to utilize the high
potential of the ACM adaptation.

Figures 13 and 14 show the basic functions of an ACM control implementation for the FL
and RL, respectively. The MODCOD selection can be done either in the terminal or the GW.
Basically, the demodulator derives the channel estimates, which are then sent to the post-
processing module. The post-processing module can be omitted, but can be used to improve the
channel estimate from the demodulator by performing the averaging, worst-case SNIR
estimation over a given time window, or even by implementing more advanced algorithms. The
post-processing module will generate a table with the last valid carrier-to-noise plus interference ratio \( \frac{C}{(N+I)} \) (called CNI in the following) value for each terminal. This CNI table is used by the MODCOD selection module in order to select the appropriate MODCOD for each terminal. The MODCOD to be used is found from a pre-calculated table, which contains the decision threshold for each MODCOD. In the case that hysteresis is included, both up- and down-thresholds must be used. The selected MODCOD is then sent to the ACM manager. The ACM/RRM manager might decide to use a more protected MODCOD when performing the overall resource optimization if there is available capacity in a certain physical layer frame (FL) or on a certain physical carrier (RL). However, the ACM manager should never use a less protected MODCOD than that indicated by the MODCOD selection unit.

There are multiple options for signalling MODCOD/CNI from the terminal to the gateway. For the demonstrator, a combination of periodic DVB-RCS synchronization bursts (SYNCs), contention MINISYNCs and traffic bursts (TRF) has been used for sending MODCOD/CNI reports.
The following strategy represents the baseline implementation adopted for the demonstrator:

```java
if (CNI report required) {
    if (exists a SYNC-burst to be transmitted within the current SF) {
        send CNI/MODCOD_RQ in the next SYNC-burst;
    }
    else if (exists at least one TRF-burst to be transmitted within the current SF) {
        send CNI/MODCOD_RQ in in-band in the next TRF-burst
    }
    else {
        send CNI/MODCOD_RQ using an available contention-based mini-slot (choose random)
    }
}
else {
    send CNI/MODCOD_RQ in the next SYNC-burst
}
```

*CNI report required* could be either because *\{CNI_{last,reported} - \Delta C\}* or because *\{MODCOD_RQ changed to more robust\}* where \(\Delta C\) represents the fixed margin.

The demonstrator supports both of the following options:

(a) CNI reporting without MODCOD values from the terminal (i.e. MODCOD selection in the gateway).
(b) CNI reporting with MODCOD value from the terminal (i.e. MODCOD selection in the terminal).

Option (b) is baseline. When using adaptive margin, only option (b) is applicable.

*ACM threshold margin*: The design supports two different types of margin: (a) fixed constant/variable MODCOD margin and (b) adaptive margin. With the fixed constant margin one uses the same margin for all MODCODs. With the fixed variable MODCOD margin one uses a fixed margin for each MODCOD, but the margin depends on the actual MODCOD. The fixed/variable margin is included in the MODCOD table used for MOCOD selection; see Figures 13 and 14. This design allows for a separate margin for each MODCOD. The FL and RL tables are configured during set-up and kept fixed during the measurement. We will further refer to this method as the fixed margin approach.

*Adaptive margin*: This method was derived first in [16], taking inspiration from the outer loop design for mobile power control. The idea is to adapt the margin to the channel conditions. The proposed scheme uses the same margin for all MODCODs. One can start with a large margin, which is reduced by a small step as long as the receiver decodes successfully and increased with a large step when a packet error is detected. This requires that a successful decoding flag be signalled to the MODCOD selection process. If we consider adaptive margin used in the FL link, this decoding flag is not included in the present update to the return link signalling for DVB-S2 (DVB-RCS) where only a CNI report or a MODCOD request is possible. This means that the margin adaptation must be done by MODCOD selection in the terminal. For the RL, this can be more easily supported in the GW. However, we propose only to use adaptive margin for the FL link.

The idea is to minimize the required margin, ensuring that the required PER is achieved. This could be done for each MODCOD separately, but for simplicity here it is proposed to adapt the margin without considering the MODCODs used.

The algorithm described in Table VIII uses margin step-up and step-down to control the overall packet-error rate and is in general applicable to any packet-error rate. If we assume that
the target PER is achievable by adjusting the margin, then \(M(k)\) will vary over time around an average margin controlled by the ratio between error packets and correct packets. The period of this fluctuation will depend on packet rate and target PER. A low target PER will result in both (a) a long time to 'stabilize' \(M(k)\) on the correct long-term level; (b) long period of the fluctuation around this level.

Adaptive margin is implemented in the MODCOD selection unit using separate variables: Margin_Adaptive, Delta_Step_Up and Delta_Step_Down. For the demonstrator, a Delta_Step_Up of 0.5 dB has been chosen after some simulation optimization. Owing to the long feedback delay one might get several packet errors when approaching the MODCOD threshold, resulting in multiple step-ups. Therefore, an additional constraint was implemented where one only allows a single step-up, even if more packet errors occur within the round-trip delay of the adaptation algorithm. However, in this case the relation \(\text{Target\_PER} = \Delta M_{\text{down}} / \Delta M_{\text{up}}\) is not 100% correct since the actual PER will be somewhat higher than the value given by the above step-down/up relation. For the demonstrator, a target PER of \(10^{-4}\) was chosen in order to have a reasonable convergence time.

The adaptive margin can be combined with the fixed margin in a common design as shown in Figure 15. The MODCOD tables include fixed/variable margin as described above, and the adaptive margin may or may not be added to the received CNI values. Note that when one uses adaptive margin, the margin will, during clear sky conditions, adapt to the actual demodulator threshold, independent of the thresholds set in the MODCOD table. However, it might still be desirable to use hysteresis and to use variable margin (i.e. larger margin for lower MODCODs)
during a fade. The proposed design will support this. During a fade, one will step down to low MODCOD, which has a large margin given by the MODCOD table. Typically, the adaptive algorithm will be back to clear sky conditions before the algorithm has reduced the margin too much, because the margin adaptation rate is quite slow. The successful decoding flag, triggering the adaptive margin, can be based on validating BBFRAME, MPEG or IP packets. In the demonstrator, IP packet flag will be used.

**Hysteresis:** Hysteresis is used to avoid the oscillation between two code rates/symbol rates during steady-state conditions. ACM adaptation with hysteresis, which has been investigated in depth in [11], is implemented by having different up and down decision thresholds in the MODCOD table used for MOCOD selection, see Figures 13 and 14. This means that the down-threshold for MODCOD($N+1$) is not equal to the up-threshold for MODCOD($N$). In practice, the down-threshold for MODCOD($N+1$) is lower than the up-threshold for MODCOD($N$), providing an overlapping region between MODCOD($N$) and ($N+1$) where the $C/(N+I)$ estimate can fluctuate without causing unwanted MODCOD changes. No slope detection is

\[
\text{ModCod} = \text{Threshold} \ (\text{CNI}) ;
\]

![Figure 16. CNI to MODCOD table.](image)

![Figure 17. ACM control post-processing.](image)
necessary with this implementation, but the previous MODCOD used must be stored, in order
to determine which threshold should be used for MODCOD selection. To implement hysteresis,
two tables are used. The UpThreshold table is shifted by a pre-defined margin relative to the
DownThreshold table, as shown in Figure 16. The hysteresis can be simply removed by setting
the up- and down-thresholds to be identical. In-depth investigations indicated that a hysteresis
interval on the order of 0.25 dB should be used for the forward link. However, when using
adaptive margin, the hysteresis interval may be set to zero.

**Post-processing:** Basically, this module performs post-processing on the last \( N \) channel
estimates (Figure 17). The parameter \( N \) should be configurable and selected according to the
physical layer configuration (symbol rate, number of bursts received in the RL). The post-
processing can be either the window averaging or worst-case SNIR estimation, but simulation
findings favoured the averaging approach for both the FL and RL links.

**Forward link:** Here the CNI will be continuously calculated, based on averaging the CNI for
the \( N \) last pilot symbols. The last valid CNI is transmitted to the terminal CPU, with a repetition
period \( T_r \) between each measurement. The \( N \) and \( T_r \) parameters are configurable from the
demonstrator control and monitoring system (CMS).

**Return link:** The existing implementation is based on measuring the CNI for the SYNC-burst,
and the measurements are sent to the NCC. The NCC then performs averaging over the last \( N \)
measurements.

### 3.2. Traffic generator

The main components of the traffic simulator are two traffic generator/analyzer (TGA) PCs.
These two PCs have the same configuration and are both able to act as traffic generator and
analyzer at the same time. Each TGA has two Ethernet interfaces, one of which connects to the
demonstrator and the other connects to the CMS. The first interface is used to send/receive the
useful traffic, while the second interface is used to control the TGAs. Both links use standard IP
protocols for all purposes.

Between the TGA and the CMS, network time protocol (NTP) is used to synchronize the
clocks for delay and jitter measurements. The simple network management protocol is used to
transfer control information and measurement results between the TGA and the CMS as well as
to control the traffic generation and measurement process. The overall architecture is depicted
in Figure 18.

A particular note has to be reserved to the NTP in virtue of the synchronization delays that it
may introduce. In fact, NTP protocol implementation has been properly tuned so as to reduce
synchronization drift to less than 1 ms. In particular, the performed tests showed that in the
presence of light traffic load, synchronization delay of 40 \( \mu \)s was achieved, whereas in the
presence of higher load it could be increased up to 1 ms. In this light, it is worth noting that even
a synchronization delay of 1 ms can be considered negligible because of the propagation delay
experienced on the satellite side and the available link bandwidth present on all the segments
composing the testbed.

The TGA implements two kinds of traffic generators: application-like traffic generators and
statistical traffic generators. Application-like traffic generators simulate single application
instances, while statistical traffic generators simulate the traffic characteristics of links with a
multitude of connections to applications and users. Both types of generators can be instantiated
multiple times and are multiplexed together to form the resulting traffic going to the
demonstrator. This allows simulating a broad range of network applications from single user/single application home user to corporate users with a lot of applications and users. In particular, voice, video, web browsing and data transfer traffics, shortly described in the following, can be generated by TGAs:

- **Voice**: The traffic generator can reproduce both constant bit rate and talk spurt VoIP sources. In the first case, several bit rates can be set, depending on the encoding algorithm: 64 kbps for G.711, 8 kbps for G.729, 5.3 kbps for G.723.1 and 13 kbps for GSM, just to cite a few, although a larger set of speech codecs is available [17]. On the other hand, a talk spurt VoIP model reproduces more adequately the effect of silence detection. In practice, the audio model implements an active voice source with silence detection active, according to an MMPP distribution [18].

- **Video**: TGA reproduces video flows implementing both H.263 and MPEG-4 codec [19], which correspond to some of the most used ones. As far as H.263 is concerned, the following parameters are used: the size of the I-frames ($S_I$), the size of the P-frames ($S_P$), the size of the PB-frames ($S_{PB}$), the duration of the I-frames ($D_I$), the duration of the P-frames ($D_P$) and the duration of the PB-frames ($D_{PB}$). The I-frames are sent at a fixed rate, whereas the rate at which P- and PB-frames are transmitted is variable because the codec tries to adapt to the target bit rate. In the case of MPEG-4 flows, self-similarity propriety of flows is implemented as well through a long range dependence model.

- **Web browsing**: It is simulated by TGAs through a heavy-tailed packet size model, in order to reproduce the self-similar nature of this kind of traffic. Besides, the connection inter-
arrival time is modelled by an exponential distribution, whereas the number of packet calls is derived from an exponential distribution.

- **Data transfer**: TGAs can reproduce file transfer protocol (FTP), peer-to-peer, E-mail and real-time access (e.g. Telnet and SSH). Their implementation follows specific models developed in the literature [20–24], to which the reader may refer for further details.

### 3.3. Channel emulator

The channel emulator supports the real-time emulation of:

- propagation delay;
- nonlinearity of the satellite transponder;
- phase noise;
- thermal noise;
- interfering signals;
- propagation fading.

Two modes of operation have been implemented. For the characterization of the demodulation performance in the case of stationary reception, the parameters such as $E_b/N_0$ or $C/I$ are set directly while the fading simulator is switched off. In the second mode, the signal levels of the main signal and the interfering signals are time variant. The time variable signal levels have been implemented by a fading simulator. The fading simulator is mainly used for testing the ACM mode.

The power level profiles (‘time series’) generated by the channel model described below are converted into discrete time series dubbed as ‘fading coefficients’. The fading coefficients representing the channel propagation characteristics are stored with a sampling rate of 2 kHz to faithfully support slow and fast variations of the signal level. A digital interpolation filter (‘up-
sampling') converts the fading coefficient to the sampling frequency of the main channel emulator signal (80 MHz for the selected implementation). The method allows fading data to be generated in advance and stored in data files. The fading data can be derived from measurements or generated by software-based fading models described in the following.

Figure 19 shows an overview of the channel emulator, while Figure 20 shows a more detailed block diagram. The block diagram represents one subsystem. Two identical subsystems are integrated in the testbed to simulate the channel for the forward link and the return link. The input signal is an IF-band signal that is converted to digital form and down-converted to baseband. Full digital implementation ensures a high level of accuracy. It is a complex signal (indicated by a double line). Single lines represent real signals. A first-level adjustment (\#1) normalizes the power of the input signal to a reference value. All other signal levels can be set relative to this reference value. The interference can be added before (interference generator \#1) or after the nonlinearity (interference generators \#2 and \#3). If the interference is added before the nonlinearity, a second-level adjustment (\#2) may be useful to achieve a constant HPA input back-off (IBO), if desired. The level adjustment also provides a signal-level measurement for monitoring purposes. To correctly drive the HPA model the unit power signal is scaled down by \(10^{-\text{IBO}/20}\) where IBO is the HPA IBO. The HPA model is implemented as a memoryless device characterized by a spline interpolation of the experimental AM/AM \(A(\rho)\) and AM/PM \(\delta(\rho)\) points. The input signal envelope \(\rho\) and phase \(\varphi\) are measured and transformed as \(A(\rho)e^{i(\varphi+\delta(\rho))}\) through HPA AM/AM and AM/PM look-up tables. At the HPA output there is a Level Control \#3 to measure the output power and to normalize the signal to unitary power again. This level detector/adjustment provides the reference value for the thermal noise generator. When this level control is set to manual and gain A3 is set to 1, the measured power corresponds to the HPA output back-off (OBO), provided that the used HPA characteristic is normalized. After level adjustment (\#3), co-channel and adjacent channel interference is added.
according to the setting \((C/I)\) output), and eventually the signal is multiplied by the fading factor and the phase noise jitter before adding the Gaussian noise.

In the manual mode the Gaussian noise PSD \(N_0\) can be set directly by the user. In automatic mode the \(N_0\) is set relative to the signal level such that the useful signal power to noise ratio corresponds to the desired \(C/N_0\). Different types of interfering signals can be generated to simulate various applications and usage scenarios. For example, the following scenarios can be generated:

- **Constant \(C/I\):** This represents the forward link case whereby an interfering signal is emitted from the same satellite. In this case the fading affecting the interfering signal and the main signal is the same. The interferer is added before the fading (or the same fading profile is applied to the interferer).
- **Time-variant \(C/I\):** In this case a separate level profile for the interfering signal can be generated and applied independently from the main signal. The signals are added after the fading process (this mode is not shown in the block diagram, Figure 20).
- **The channel emulator includes an arbitrary signal generator.** This allows generation of an interfering signal with time division multiple access (TDMA) structure, for example. Signals with arbitrary level profiles can be generated to also simulate temporal interference.

The centre frequency of the interfering signal can be set to simulate co-channel interferer, adjacent channel interferer (ACI) or combinations of the two.

The channel emulator is based on commercially available hardware from IZT GmbH. Within the project new firmware (FPGA code and related software running on a built-in industrial PC) has been developed.

As stated at the beginning of this section, characterization of both interference and fading has been implemented in the channel emulator in order to have complete channel modelling on both forward and reverse links. As far as interference is concerned, two different models are available for distributing the interferers: (a) uniformly distributed and (b) distributed according to population density. The model is chosen at configuration time together with the number of interferers \((N_I)\) to generate the coordinate range for the users. In the case of uniformly distributed interferers, the users are equally distributed inside the satellite coverage region. There is no provision to remove users from sea or other uninhabited areas. The coordinates are generated using the following formulas:

\[
L_i = r(L_{i_{\text{max}}} - L_{i_{\text{min}}}) + L_{i_{\text{min}}} \\
l_i = r(l_{i_{\text{max}}} - l_{i_{\text{min}}}) + l_{i_{\text{min}}}
\]

(3)

where:

- \(L_i, l_i\) are the longitude and latitude of the \(i\)th interferer, \(L_{i_{\text{max}}}, L_{i_{\text{min}}}\) the upper and lower limits of the range of interferer longitudes and \(l_{i_{\text{max}}}, l_{i_{\text{min}}}\) the upper and lower limits of the range of interferer latitudes, respectively.
- \(r\) is a random number equally distributed in \([0, 1]\).

On the other hand, when the interferer distribution according to population density is considered, population densities available from the Gridded Population of the World project
are used as the input distribution for the channel emulator. Note that for the reverse link all interferers generated according to the two aforementioned models are in turn classified into three groups: interferers with negligible interference, interferers with non-negligible interference that fall into a certain spatial range around the useful user and all other interferers (see Table IX).

In other words, the first class of interferers is removed from all following calculations—its total amount of interference is assumed to be small enough so that it can be neglected. For the second group the rain attenuation is computed from a dynamic rain attenuation model including spatial correlation, and for the third group a statistical model for rain attenuation is assumed without correlation between the interferers according to the ONERA-CNES spectral model [23].

The classification is done in two steps. Initially, the first group of interferers (the ‘No interference’ class, as described in Table IX) is sorted out. This group actually consists of three sub-groups. The first sub-group consists of all interferers that use the same carrier and polarization within the same beam as the useful user. The motivation underlying this choice is that only one among the interferers and useful user can be active as a consequence of the TDMA operation mode. It follows that these interferers cannot interfere with the useful user. The second sub-group consists of all interferers that are two or more carrier frequencies away from the useful user. It is assumed that the suppression of these carriers in the receive equipment is good enough so that it can be neglected. For the third sub-group the clear sky interference power at the satellite for all interferers and the useful power for the useful user are computed. The useful user received carrier power is computed with the rain attenuation set to 0 dB (and is therefore not dependent on time):

\[
C_{\text{up, useful user, clear sky}} = \frac{\text{EIRP}_{\text{Terminal}} - L_{\text{Path}} + G_{\text{RxSat}}}{C_0} \text{ (dBW)}
\]

(4)

The appropriate formula for the interferer \(i\) is

\[
I_{\text{up, interferer } i, \text{ clear sky}} = \frac{\text{EIRP}_{\text{Terminal}} - L_{\text{Path}}}{C_i} + G_{\text{RxSat}} - f_i \cdot \frac{(C/I)_{ACI}}{C_0} - p_i \cdot \frac{(C/I)_{Pol}}{C_0} \text{ (dBW)}
\]

(5)

Table IX. Reverse link interferer classes.

<table>
<thead>
<tr>
<th>Interferer class</th>
<th>Conditions</th>
<th>Computations</th>
</tr>
</thead>
<tbody>
<tr>
<td>No interference</td>
<td>All interferers in the same beam with the same polarization and on the same carrier as the useful user All interferers that are more than one carrier frequency away from the useful user All interferers whose clear-sky interference power is larger than useful users' received carrier power (C_{\text{up, useful user, clear sky}}) divided by the reference threshold ((C/I)_{\text{up, thresh}})</td>
<td>—</td>
</tr>
<tr>
<td>Interferers with spatially correlated rain attenuation</td>
<td>All interferers not in the above class which are inside a certain geographical range from the useful user</td>
<td>Interferers with spatially correlated rain attenuation Interferers not in the above class which are inside a certain geographical range from the useful user Rain attenuation computed from dynamic rainmap</td>
</tr>
<tr>
<td>Interferers without spatially correlated rain attenuation</td>
<td>All other interferers</td>
<td>Interferers without spatially correlated rain attenuation All other interferers Rain attenuation uncorrelated between the interferers (ONERA-CNES spectral model, [25])</td>
</tr>
</tbody>
</table>
where \((C/I)_{ACI}\) is the adjacent channel interference suppression in dB and \((C/I)_{Pol}\) is the suppression of the other polarization (on the same carrier) in dB. \(p_i\) is zero if the interferer \(i\) and the useful user use the same polarization and one otherwise. \(f_i\) is one if the interferer \(i\) and the useful user use adjacent frequencies and zero if they use the same frequency. If the resulting \(C_{up, useful user, clear sky}/I_{up, interferer i, clear sky}\) is below a certain threshold \((C/I)_{up, thresh}\) (configurable with a default of 50 dB), the interferer is assumed to be negligible and put into the first class. For the remaining users their horizontal and vertical distance from the useful user is computed, and if both distances are smaller than a certain, configurable parameter, they are put into the second class of interferers for which spatial correlation is considered. All other interferers go into the third class. Relying on this classification, the time series of the interference power spectral density \(I_{0, up}(t)\) in the uplink is calculated as:

\[
I_{0, up}(t) = \frac{1}{R_{c, up}} \sum_{i, all uplink interferers} I_{up, interferer i}(t) (W/Hz)
\]

\(R_{c, up}\) is the carrier symbol rate of the uplink. \(I_{up, interferer i}(t)\) is the uplink interference power generated by the interferer \(i\). \(I_{up, interferer i}(t)\) can be calculated as

\[
I_{up, interferer i}(t) = \frac{C_{up, interferer i}(t)}{(C/I)_{up, interferer i, to useful signal} \cdot \text{activity}_{interferer i}(t) (W)}
\]

\(C_{up, interferer i}(t)\) is the carrier power received by the satellite antenna from a transmitting interferer \(i\). \((C/I)_{up, interferer i, to useful signal}\) is the ratio between the carrier power received by the satellite antenna from the interferer \(i\) (in beam of the useful user) and the interference power generated by the interferer \(i\) with respect to the useful user. Finally, as the interferers are not always transmitting, an activity time series \(\text{activity}_{interferer i}(t)\), which switches between 0 and 1, is taken into account as well. Its dynamics depend on the characteristics of the aggregated traffic transported by each carrier.

The forward link also deserves some attention. In practice, the time series of the interference power spectral density \(I_{0, down}(t)\) in the downlink is calculated as

\[
I_{0, down}(t) = \frac{1}{R_{c, down}} \sum_{i, all uplink signals} I_{down, interfering signal i}(t) (W/Hz)
\]

\(R_{c, down}\) is the downlink carrier symbol rate of the forward link. \(I_{down, interfering signal i}(t)\) is the terminal antenna received interference power of interfering signal \(i\) in the downlink. This can be calculated as

\[
I_{down, interfering signal i}(t) = \frac{C_{Terminal, interfering signal i}(t)}{(C/I)_{down, interfering signal i, w.r.t. useful signal} (W)}
\]

where \(C_{down, interfering signal i}(t)\) is the carrier power received by the terminal antenna from the interfering signal \(i\) and \((C/I)_{down, interfering signal i, w.r.t. useful signal}\) is the ratio between the carrier power received by the terminal antenna from the interfering signal \(i\) and the interference power of the interfering signal \(i\) with respect to the useful signal.

As far as fading characterization is concerned, particular attention has been paid to rain components. Dynamic rainfield maps use a stochastic-dynamic synthesizer of rain attenuation.
based on the work of Grémont [24,25] and Bell [26]. The outline of the algorithm is shown in Figure 21, and is briefly described in the following.

The rainmap computation takes place on a matrix of equally spaced grid points with a configurable grid granularity. For the second step it is assumed that the rain attenuation is lognormal distributed (see [24]). The range computed in the first step is now rasterized with a configurable number of points, and the lognormal parameters $\mu(x, y)$ and $\sigma(x, y)$ are computed for each grid point by computing the value of the attenuation exceeded probability for a range of probabilities from ITU-T P.618, and then curve-fitting a lognormal CDF function to these values.

In the fourth initial step, the FFT of the cross-correlation function $C_{pq}$ for the grid point $(p, q)$ is computed. From this the initial Gaussian field $a_0(x, y)$ with unit variance and zero mean is computed by multiplying the FFT of the cross-correlation function by a complex normal two-dimensional (2D) noise random variable (rv) $n_{-1}(p, q)$ with unit variance and zero mean. The rainfield is then the inverse FFT of that product $a_0(p, q)$. That operation in the Fourier domain corresponds to filtering the noise with the cross-correlation function.

To obtain a dynamic field an iteration is applied to $a_0(p, q)$, starting with the value $a_0(p, q)$. In this iteration, advection and birth/death of raincells is modelled. With a (constant) velocity $
u$ the following steps are repeated:

1. Define grid for the Rainfield computation
2. Compute location dependent $\mu$ and $\sigma$ for the lognormal distributed attenuation
3. Select spatial cross-correlation function $C_{pq}$
4. Compute Gaussian field $a_{pq}$ based on $C_{pq}$
5. Compute attenuation field $A_{pq}$
6. Apply advection and birth/death of raincells and get new Gaussian field $a_{pq}$ from $A_{pq}$
7. Compute attenuation field $A_{pq+1}$

Last time step?

Figure 21. Dynamic rainfield map algorithm.
\[ \mathbf{V} = \begin{bmatrix} V_x & V_y \end{bmatrix} \] and parameter \( \beta \)—a temporal constant (in \( s^{-1} \))—describing the birth/death rate of a raincell—the process can be mathematically described as

\[
a_{t+\Delta t}(p, q) = e^{-\beta \Delta t} \left( a_t(p, q) \otimes e^{-1 \Delta t (x V_x + y V_y)} \right) + \sqrt{1 - e^{-2 \beta \Delta t}} \otimes C_g(p, q) \otimes \mathbf{n}(p, q)
\]

The terms outside of the advection term are an AR(1) process describing the birth/death of raincells. The \( \mathbf{n}(p, q) \) are uncorrelated 2D Gaussian noise rv, as described above.

For the spatially uncorrelated users, the attenuation time series is generated with the ONERA-CNES spectral model. The block diagram of this synthesizer is shown in Figure 22.

For the computation of the downlink rain attenuation, the modified instantaneous frequency scaling model from Grémont and Filip (see [27,28]) is used. The scheme for frequency scaling is shown in Figure 23.

The lognormal distributed attenuation delivered by the reverse link attenuation model (dynamic rainmap) is de-logarithmized and normalized with the lognormal parameters from the

---

Figure 22. Block diagram of the time-series generator.

Figure 23. Frequency scaling.
reverse link:

\[ X_{rev} = \frac{\ln(A_{rev}) - \mu_{rev}}{\sigma_{rev}} \]  
\[ (11) \]

Another particular aspect related to fading characterization concerns the scintillation effect, whose time-series generator is based on the time-series generator proposed by Polytechnic of Milano [29] and can be used both for clear sky and rainfall conditions. The block diagram of the time-series generator is shown in Figure 24.

First, the rain attenuation \( A \) in dB has to be calculated. Then, the scintillation time series can be generated as filtered white Gaussian noise with corner frequency \( f_C \) and standard deviation \( s \). The standard deviation \( s \) is calculated according to the Matricciani–Riva model [29,30]:

\[ \sigma = \max(\sigma_0, \sigma_0 \cdot (A/\text{dB})^{5/12}) \]  
\[ (12) \]

where \( \sigma_0 \) is the standard deviation of the scintillation for \( A = 1 \) dB (almost clear sky). \( \sigma_0 \) is a configurable parameter.

3.4. Gateway

The gateway is based on the commercial Nera (now STM) SatLink gateway solution. The main components of the gateway are shown in Figure 25.

**DVB-S2 transmitter**: The DVB-S2 transmitter is the forward link subsystem (called FLS hereafter) of the gateway and has been developed ad hoc during the project. The main functions of the FLS include routing of user traffic, encapsulation, scheduling and forwarding of the traffic to the modulator for encoding and modulation. The functionality of the DVB-S2 transmitter can be considered to consist of three main functions: the analogue up-conversion, the modem and the higher-layer SW, Figure 26. It is basically using the same HW platform as the MCR receiver.

The modulator supports from 10 to 25 Msymb/s, but the higher-layer SW is only designed to support a single terminal with a useful bit rate of approximately 4 Mbps. Normally, one should then insert dummy PLFRAMEs to fill up the carrier, but this is not a realistic test scenario for the S2 demodulator in a real system. Instead, the S2 modulator adds dummy BBFRAMEs with time-varying MODCOD in order to emulate a fully loaded system.

During the ACM-demonstrator validation and tests, the traffic simulator generates the user traffic destined to the test terminal and the background traffic. Owing to the processing capacity limitations, the background traffic is filtered out as early as possible. To emulate a realistic scenario for forwarding/scheduling of user traffic to the terminal, the scheduling module
performs the scheduling of the user traffic depending on the background traffic load, i.e. if the generated traffic exceeds the capacity available, the user terminal should experience congestion.

Figure 27 shows the functional block diagram of the DVB-S2 modulator. Modulation is implemented as a table look-up, supporting all S2 modulation schemes. In the case where we are assuming a nonlinear HPA in the satellite and use pre-compensation in the transmitter, the pre-compensated constellation points are pre-calculated and stored in separate tables. The input stream synchronizer and null-packet deletion are not implemented.

**DVB-RCS receiver**: The DVB-RCS receiver is the burst receiver of the return link channels, and it constitutes the RLS of the gateway. The Nera SatLink 8500, hereafter denoted as SatLink 8500, is a general-purpose QPSK-burst demodulator, especially suited for DVB-RCS gateways. The compact, 19in single rack unit (1 RU) operates at 140 MHz IF and can support...
simultaneous demodulation of up to 12 carriers within a 40 MHz frequency band. This burst receiver is used for the ACM demonstrator without modifications.

The coarse block diagram of the RLS is shown in Figure 28.

Network control centre and network management system (NCC/NMS): The NCC/NMS is the core for control and management of the satellite network. It consists of an application server
and a database server. The solution is based on the NCC/NMS solution of the Nera SatLink gateway, with the necessary updates to incorporate the DVB-S2 elements. Figure 29 shows the architecture of the NCC.

**Terrestrial interface subsystem (TIS):** The TIS provides the interface towards the terrestrial network. It normally consists of routers or switches for the Internet connection. Other interfaces may also be supported, such as ISDN or SS7 when a VoIP option is required. For the ACM-demonstrator testbed, the TIS is chosen to be a standard Ethernet hub/switch.

### 3.5. User terminal

A functional overview of the FL receiver module is given in Figure 30. The module includes an RF tuner (converting the input analogue signal from L-band to baseband I-Q), a DVB-S2 demodulator with I-Q baseband inputs and an FEC decoding module.

The DVB-S2 demodulator consists of three main parts:

- The demodulator core, including symbol matched filter and symbol clock time loop.
- The frame synchronization.
- The feed-forward frequency/phase synchronization modules and channel estimation.

For clock time recovery non-data-aided algorithms suitable for PSK and APSK modulation described in [15] have been adopted. To meet the demonstrator’s tight requirements, data-aided algorithms based on DVB-S2 pilot fields are mandatory for frequency and phase synchronization. This is particularly true for the most protected MODCODs operating at low SNR, and for carrier phase noise scenarios combined with higher-order modulations. These data-aided algorithms, detailed in [15], are exploiting the pilot fields included in the DVB-S2 framing structure. For frame synchronization a novel approach has been developed, using both the start of frame (SOF) inside the PLHeader and the pilot fields [12]. An overview of this approach is given below. By this approach a very reliable algorithm, suitable for ACM (variable frame lengths) operating at low SNR (down to $-3$ dB), was derived.
Figure 31 details the demodulator core module architecture and functional block diagram implemented in the demonstrator.

The demodulator unit is equipped with a dual channel ADC connected to the tuner. The I-Q generation sub-module takes the ADC input as a complex or a real signal. In the latter case, to avoid problems caused by floating input ports, the imaginary part of the complex signal is forced to be zero. The IF carrier mixing is performed jointly with the coarse carrier frequency correction. Digital front-end filtering and down-sampling have to be configured according to the sampling rate at the ADC and the modulation symbol rate. The filtering/down-sampling can be achieved by cascaded stages. Matched filtering is assumed to be working at two times the oversampling rate. A first digital gain correction stage, normalizing the signal power, is located directly after the matched filter. All modules after the matched filter output decimation to symbol rate are operating at symbol rate.

---

Figure 30. FL receiver module, functional overview.

Figure 31. FL receiver module detailed block diagram of demodulator unit.
The SOF synchronization unit is internally split into two parts: (a) the constant delay buffer and (b) the SOF correlation processor. Thanks to the delay buffer the physical layer SOF location can be derived from past and ‘future’ information. By reading back the delay buffer, the frame boundaries are well identified, leading to robust SOF identification even in the ACM mode. PL de-scrambling is logically located after frame synchronization and prior to coarse- and fine-frequency synchronization. Functionally, the PL de-scrambling can be considered as the output stage of the frame synchronization; therefore, it is not shown in the figure. Final channel estimation is done on the remaining sub-modules. Phase correction is done in the fine carrier frequency and phase synchronization module, whereas amplitude correction is performed in the AGC2 module that normalizes the APSK symbol constellation. The $C/(N+I)$ estimation is derived from the fully corrected pilot symbols. PL header decoding is done inside the demodulator core, although it is only necessary for FEC decoding and not relevant for the demodulator core itself.

Frame synchronization is achieved by correlating the received baseband symbols with a known pattern, which occurs repeatedly in the symbol stream. From the peak locations the frame start can be derived. AWGN and carrier frequency offset (CFO) deteriorate the result. The correlator should be robust against CFOs within certain limits. Frame synchronization for DVB-S2 may use only the SOF and the PLSCODE field for correlation, but not the pilot blocks, which are scrambled. However, the PL scrambling sequence is known, although the number of pilot fields present in each frame is not fixed and depends on the current frame modulation format (MODCOD). As the current MODCOD received is not yet known by the demodulator, a simple solution has been devised to exploit part of the frame pilots to enhance frame synchronization as described in the following. The pilot blocks are enumerated starting with 1, i.e. pilot block no. 1 denotes the first pilot block within a PL frame. Depending on the modulation, there are at least two pilot blocks per PL frame. By correlating over the first 2 pilot blocks and by correlating over the 26 SOF symbols and by appropriately combining the respective output signals, a signal is generated that contains peaks in those locations where the last symbol of the second PB occurs. Figure 32 shows how this is done. The blocks indicated by ‘D1’ and ‘D1+D2’ denote delays by 1441 and 1441+64 symbols, respectively. After adding the squared magnitudes of the three correlator output signals, peak detection is applied in order to

Figure 32. FL receiver frame synchronization with SOF and two pilot blocks.
identify the peaks with a low probability of false alarm \( (P_{fa}) \). In the final step the peaks are validated.

This approach is quite robust against CFO. It has been verified that CFOs (normalized to symbol rate) of up to \( 1E−2 \) can be handled for \( E_{s}/N_{0} = −3 \) dB. The initial CFO is typically higher. Thus, the approach is applied in connection with a frequency sweep, where we have to test several frequency hypotheses in normalized frequency steps of \( 2E−2 \). For the assumed maximum CFO of \( ±0.1 \), we have to test 10 frequency hypotheses \( (CFO = −9E−2, −7E−2, ..., 9E−2) \).

During frame acquisition the peak validation takes an observation interval of several PL frames to increase the synchronization probability. Table X lists the synchronization probabilities versus the length of the observation interval for the example of \( E_{s}/N_{0} = −3 \) dB and \( CFO = 1E−2 \). For tracking mode with lower CFO the synchronization probabilities are even higher.

After initial acquisition, frame synchronization switches into tracking mode. Correlation values above a given threshold are taken as validated peaks. During bad signal conditions, i.e. when MODCOD is not correctly detected, all possible DVB-S2 frame lengths are used for SOF correlation to detect possible weaker correlation values, which are nevertheless representing local maxima. That sequence of local maxima that starts and ends at a validated correlation peak and shows the strongest intermediate correlation values is selected; thus, a short period of bad signal quality can be overcome.

The synchronization, channel estimation and decoding subsystems have been analyzed in the presence of channel impairments (AWGN, phase noise, nonlinearities and interference), clock offsets and carrier frequency/phase offsets. The analysis included performance and estimation errors in both acquisition and tracking mode.

As an example of the performed analysis, some simulation results are shown in Figure 33.

The FEC module has been derived from TurboConcept’s commercial IP Core TC4000, for the considered Virtex4 FPGA platform. All MODCODs are supported in full ACM mode. The functional coverage of the encoder module is:

- BCH encoding.
- LDPC encoding.
- Bit interleaving.

The bit to I/Q mapping is not performed in the module.

The functional coverage of the decoder module is:

- Bit to I/Q demapping.
- Bit deinterleaving.
- LDPC decoding.
- BCH decoding.

### Table X. Frame synchronization probability for \( E_{s}/N_{0} = −3 \) dB and \( CFO = 1E−2 \).

<table>
<thead>
<tr>
<th>( N )</th>
<th>4</th>
<th>5</th>
<th>6</th>
<th>7</th>
<th>8</th>
<th>9</th>
<th>10</th>
</tr>
</thead>
<tbody>
<tr>
<td>( P_{sync} (N) )</td>
<td>0.9688</td>
<td>0.9919</td>
<td>0.9980</td>
<td>0.9995</td>
<td>0.9999</td>
<td>1–2.8E−5</td>
<td>1–6.5E−6</td>
</tr>
</tbody>
</table>

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DOI: 10.1002/sat
The decoder performance with 50 iterations is within 0.1 dB loss from the ideal DVB-S2 performance for the vast majority of the MODCODs. In order to obtain a receiver of reasonable complexity, the number of iterations can be reduced for high-throughput links. The internal architecture of the decoder allows a reduced number of iterations with small losses in performance. Typically, 15 iterations can be used for high code rates with negligible degradation (0.05 dB additional degradation for rates $>3/4$). The demodulator monitoring and control interface is connected to the CMS. Its task is on the one hand to configure the demodulator and FEC core based on the given system configuration, and on the other hand provide extensive monitoring information about the demodulator and FEC status to the CMS.

4. TEST RESULTS

The ACM modem demonstrator has been used to perform an extensive system test campaign, the purpose of which was to evaluate the performance of a DVB-S2+ACM/DVB-RCS+AC+DRA system in the context of an emulated Ka-band multi-beam satellite system configuration, described in Section 2. The main outcome and findings of the system test results and analysis are summarized in the following sections.

Test conduction has been automated to a large extent in order to handle the high number of parameters of the platform in a reliable way and to program test sequences. The control and monitoring station (CMS) equipment has been developed by Astrium to perform testbed
automation tasks, to monitor the platform in real time and to perform post-processing of the results.

4.1. CCM performance

In Figures 34 and 35 forward link performance has been characterized by laboratory experiments for each MODCOD separately in CCM mode, with constant link conditions. Each MODCOD has been tested by transmitting between 1 and 10 million frames until QEF behaviour was obtained. Two QEF levels have been tested: FER = 5E−4 (approximately

Figure 34. FER performance versus SNIR for QPSK MODCOD.

Figure 35. FER performance versus SNIR for each MODCOD.
1 frame error every 30 s) and FER = 2E−6 (approximately 1 frame error per hour). Frame error rate curves, represented below, tend to be very steep when close to the QEF point because of the powerful FEC combined with the long physical layer frames. Consequently, the difference between the two levels of thresholds tends to be very small.

As shown in Figure 36 the resulting QEF thresholds are relatively close to the DVB-S2 standard reference thresholds [31]. As expected, degradation increases with the order of modulation: from 0.25 dB in QPSK up to 0.62 dB in 32-APSK for an FER = 5E−4. Other degradation effects (phase noise, nonlinearities, etc.) have been evaluated separately with sensitivity tests.

4.2. ACM performance

ACM loop mechanisms have been extensively validated in different modes, and against a variety of time series. The ACM loop response time is as low as 1 s, due to the exploitation of high-priority ACM feedback messages on the return link. ACM adaptation performance has been measured and compared by running 30 cases from a combination of 3 parameters:

- Three types of ACM margin: fixed, variable and adaptive.
- Two sets of MODCODs: basic set with 22 MODCODs and reduced set with 4 MODCODs. Note that, in practice, the higher MODCODs are not used because the link budget does not allow the $C/N+I$ to be better than the MODCOD 19 threshold.
- Five different fading time series from clear sky to deep fading.

The example reported in Figure 37 shows how the ACM loop adapts to an extreme fading event (case tested with variable ACM margin and 22 MODCODs activated). When SNIR changes on the link, the ACM loop adapts the forward link MODCOD. The corresponding
MODCOD threshold is always below the SNIR curves, which shows that no ACM outage events occurred.

ACM performance results for the 30 cases tested are summarized in Table XI. The following measurements have been performed based on the frame-by-frame statistics:

- Frame error rate, characterizing outage events that may appear when the ACM loop does not react fast enough compared with the fading event tested.
- Link availability: proportion of time when the link is active, i.e. when MODCOD 1 or higher can be correctly decoded. Losses should occur only when SNIR is lower than MODCOD 1 threshold.
- Spectral efficiency averaged on the event, compared with three reference cases:
  - CCM MODCOD reference at 99.9% availability.
  - CCM MODCOD reference at 99.7% availability (note that the 99.7% target is higher than the 99.5% target foreseen when defining the reference system).
  - Ideal ACM case where the ACM decision is taken instantaneously without ACM margin (impossible in practice, of course).

Statistics have been collected from the five different time series tested and averaged with the following weights to derive realistic global statistics: clear sky = 50% of time, light fading = 30%, moderate fading = 18%, extreme fading = 1.9%, deep fading = 0.1%.

Based on the reference system settings, the link budget allows for MODCOD 18 (16-APSK 2/3) in clear sky, which occurs approximately 50% of the time. An important gain of spectral efficiency is obtained with ACM compared with CCM, when variable margin and 22 MODCODs are activated:

- +130% compared with MODCOD 5 with a 99.9% availability target.
- +53% compared with MODCOD 12 with a 99.7% availability target.
Table XI. Synthesis of ACM performance results.

<table>
<thead>
<tr>
<th>Margin</th>
<th>MODCOD set</th>
<th>Time series</th>
<th>Frame error rate</th>
<th>Link availability</th>
<th>Spectral efficiency (bit/s/Hz)/comparison with reference cases</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td></td>
<td></td>
<td>Outage</td>
<td>Events</td>
<td>Losses</td>
</tr>
<tr>
<td>Fixed</td>
<td>Basic set</td>
<td>Clear sky</td>
<td>0 0.0E+0</td>
<td>0 0.0E+0</td>
<td>0 100.00%</td>
</tr>
<tr>
<td></td>
<td>(22 MODCOD)</td>
<td>Light fading</td>
<td>0 0.0E+0</td>
<td>0 0.0E+0</td>
<td>0 100.00%</td>
</tr>
<tr>
<td></td>
<td></td>
<td>Moderate fading</td>
<td>0 0.0E+0</td>
<td>0 0.0E+0</td>
<td>0 100.00%</td>
</tr>
<tr>
<td></td>
<td></td>
<td>Extreme fading</td>
<td>0 0.0E+0</td>
<td>751 99.86%</td>
<td>1.591</td>
</tr>
<tr>
<td></td>
<td></td>
<td>Deep fading</td>
<td>0 0.0E+0</td>
<td>806 99.84%</td>
<td>1.763</td>
</tr>
<tr>
<td></td>
<td>Reduced set</td>
<td>Clear sky</td>
<td>0 0.0E+0</td>
<td>0 0.0E+0</td>
<td>0 100.00%</td>
</tr>
<tr>
<td></td>
<td>(4 MODCOD)</td>
<td>Light fading</td>
<td>0 0.0E+0</td>
<td>0 0.0E+0</td>
<td>0 100.00%</td>
</tr>
<tr>
<td></td>
<td></td>
<td>Moderate fading</td>
<td>0 0.0E+0</td>
<td>0 0.0E+0</td>
<td>0 100.00%</td>
</tr>
<tr>
<td></td>
<td></td>
<td>Extreme fading</td>
<td>0 0.0E+0</td>
<td>744 99.88%</td>
<td>0.465</td>
</tr>
<tr>
<td></td>
<td></td>
<td>Deep fading</td>
<td>0 0.0E+0</td>
<td>803 99.86%</td>
<td>1.219</td>
</tr>
<tr>
<td></td>
<td>Variables</td>
<td>Clear sky</td>
<td>0 0.0E+0</td>
<td>0 0.0E+0</td>
<td>0 100.00%</td>
</tr>
<tr>
<td></td>
<td>Basic set</td>
<td>Light fading</td>
<td>0 0.0E+0</td>
<td>0 0.0E+0</td>
<td>0 100.00%</td>
</tr>
<tr>
<td></td>
<td>(22 MODCOD)</td>
<td>Moderate fading</td>
<td>0 0.0E+0</td>
<td>653 99.88%</td>
<td>1.624</td>
</tr>
<tr>
<td></td>
<td></td>
<td>Extreme fading</td>
<td>0 0.0E+0</td>
<td>849 99.84%</td>
<td>1.833</td>
</tr>
<tr>
<td></td>
<td></td>
<td>Average</td>
<td>0 0.0E+0</td>
<td>100.00%</td>
<td>2.668</td>
</tr>
<tr>
<td></td>
<td>Reduced set</td>
<td>Clear sky</td>
<td>0 0.0E+0</td>
<td>0 0.0E+0</td>
<td>0 100.00%</td>
</tr>
<tr>
<td></td>
<td>(4 MODCOD)</td>
<td>Light fading</td>
<td>0 0.0E+0</td>
<td>0 0.0E+0</td>
<td>0 100.00%</td>
</tr>
<tr>
<td></td>
<td></td>
<td>Moderate fading</td>
<td>0 0.0E+0</td>
<td>0 0.0E+0</td>
<td>0 100.00%</td>
</tr>
<tr>
<td></td>
<td></td>
<td>Extreme fading</td>
<td>0 0.0E+0</td>
<td>703 99.88%</td>
<td>0.479</td>
</tr>
<tr>
<td></td>
<td></td>
<td>Deep fading</td>
<td>0 0.0E+0</td>
<td>795 99.86%</td>
<td>1.364</td>
</tr>
<tr>
<td></td>
<td></td>
<td>Average</td>
<td>2.1E−7</td>
<td>100.00%</td>
<td>2.573</td>
</tr>
<tr>
<td></td>
<td>Adaptive</td>
<td>Clear sky</td>
<td>342 8.2E−4</td>
<td>0 100.00%</td>
<td>2.902</td>
</tr>
<tr>
<td></td>
<td>Basic set</td>
<td>Light fading</td>
<td>7031 9.6E−3</td>
<td>0 100.00%</td>
<td>2.593</td>
</tr>
<tr>
<td></td>
<td>(22 MODCOD)</td>
<td>Moderate fading</td>
<td>17362 2.2E−2</td>
<td>0 100.00%</td>
<td>2.745</td>
</tr>
<tr>
<td></td>
<td></td>
<td>Extreme fading</td>
<td>1902 3.7E−3</td>
<td>710 99.86%</td>
<td>1.744</td>
</tr>
<tr>
<td></td>
<td></td>
<td>Deep fading</td>
<td>5686 1.1E−2</td>
<td>785 99.84%</td>
<td>1.980</td>
</tr>
<tr>
<td>Reduced set</td>
<td>Clear sky</td>
<td>Light fading</td>
<td>Moderate fading</td>
<td>Extreme fading</td>
<td>Deep fading</td>
</tr>
<tr>
<td>-------------</td>
<td>-----------</td>
<td>--------------</td>
<td>-----------------</td>
<td>---------------</td>
<td>------------</td>
</tr>
<tr>
<td>(4 MODCOD)</td>
<td>0 0.0E+0</td>
<td>3495 4.7E-3</td>
<td>4934 5.9E-3</td>
<td>3125 5.6E-3</td>
<td>1082 2.0E-3</td>
</tr>
<tr>
<td></td>
<td>100.00%</td>
<td>100.00%</td>
<td>100.00%</td>
<td>99.87%</td>
<td>99.85%</td>
</tr>
<tr>
<td></td>
<td>2.896 149.68%</td>
<td>2.574 121.87%</td>
<td>2.640 127.62%</td>
<td>0.818 -29.46%</td>
<td>1.663 43.35%</td>
</tr>
<tr>
<td></td>
<td>+66.50%</td>
<td>+47.95%</td>
<td>+51.78%</td>
<td>-52.96%</td>
<td>-4.41%</td>
</tr>
<tr>
<td></td>
<td>+0.00%</td>
<td>+2.58%</td>
<td>-2.19%</td>
<td>-52.96%</td>
<td>-15.37%</td>
</tr>
</tbody>
</table>

Average 7.3E-3 100.00% 2.758 +137.74% +58.53% +1.32%
These large gains are explained by the fact that ACM can adaptively compensate for the large fading, which can sporadically occur at Ka-band. ACM allows error-free links without outage events, provided that ACM margins are correctly set.

According to Section 3.1 three types of ACM margins have been tested (see Figure 38):

- Fixed margin: all MODCODs use the same margin for handling variations of SNIR in a 1 s period. The ACM loop with fixed margin works correctly, although spectral efficiency is not optimal.
- Variable margin: each MODCOD has a different margin, which optimizes the ACM loop and spectral efficiency (approximately +10% compared with fixed margin). The margin must be calibrated for each MODCOD.
- Adaptive margin: consists of reducing the ACM margin progressively until a frame is lost. Adaptive margin gives a further gain in efficiency, while bursts of packets may be lost because of the slow margin adaptation loop characteristics. The resulting PLR ranges from 1E−3 to 1E−2, depending on the number of MODCODs and the fading severity. This performance may be acceptable for certain types of UDP applications, which are usually tolerant against packet losses, but not for all types of applications. The adaptive algorithm could be improved with a better channel estimator for the distance to QEF—based, for instance, on the number of bits corrected by the LDPC decoder—and making sure that the selected MODCOD threshold is always higher than channel SNIR.

The loss of efficiency when reducing the number of MODCODs from 22 to 4 is theoretically evaluated to 2% and was measured between 3 and 5% on the different cases tested. The loss of efficiency of implementing the ACM loop, compared with an ideal ACM system that reacts immediately (i.e. without ACM margins), is approximately 2% (variable margin and 22 MODCODs case). In practice it is suggested to use the fixed or variable margins with hysteresis. In this case the cost of implementing ACM on top of CCM is rather low compared with the gain obtained in terms of spectral efficiency. In conclusion, the good performance measured with a practical implementation ACM proves that an important gain of spectral efficiency can actually be achieved with the simple proposed ACM loop, in particular for Ka-band, while maintaining a very high link availability.
4.3. Satellite nonlinearity impact

The satellite nonlinearity impact using the DVB-S2 HPA models reported in [6] has been measured by activating the channel estimator. As expected, losses were limited for QPSK and 8-PSK, which are almost constant envelope constellations. Significant HPA impact is instead measured for 16-APSK and is even more pronounced for 32-APSK (see Figures 39 and 40). Static, memoryless pre-compensation in the FLS gives significant improvement for nonlinearized tubes, and almost no improvement for linearized tubes. Further improvements can be achieved by exploiting the pre-distortion with memory that has not been implemented in the demonstrator but that is described in detail in [15].

Figure 39. Nonlinearized TWTA B1 at IBO = 6 dB without pre-compensation (32-APSK modulation).

Figure 40. Degradation on nonlinear transponder versus IBO on 16-APSK 3/4 MODCOD.
Whatever configuration is used, the IBO/OBO point can be optimized by a parametric study searching for the maximum spectral efficiency, which usually corresponds to maximizing the MODCOD number. The optimization point generally depends on terminal geographical position, atmospheric conditions at a given time (involving ACM) or other link budget parameters; therefore, global optimization would require statistical analysis. Anyway, it is interesting to note, observing Figure 41, that the spectral efficiency performance curve usually behaves as a quadratic function around optimum, which implies that small IBO changes (or almost equivalently, link budget changes) impact system performance in the second order only. The figure also shows that activating pre-compensation allows using a MODCOD with higher spectral efficiency by operating closer to saturation.

4.4. Phase noise impact

The impact of phase noise on forward link performance has been measured at a symbol rate of 10 Mbaud. Phase noise was generated by the channel emulator based on the DVB-S2 typical and critical masks (see Appendix H of Reference [5]). The performance has also been measured with the typical mask degraded by +5 dB.

As expected, degradation increases progressively with the amount of phase noise and with the order of the modulation. QPSK and 8-PSK MODCOD are only slightly degraded by phase noise. Significant degradation is measured in 32-APSK. The impact of phase noise on the link performance depends not only on the modulation but also on the coding: a higher degradation is expected when a higher coding rate is used (Figure 42).

4.5. Performance of IP applications with ACM

Different types of traffic have been modelled at IP level by the TGA, generating random traffic sequences representative of the behaviour of real applications over a satellite link. Each model has been tested individually. As shown in Figure 43 a mix of models with different priorities has
also been tested with the following distribution:

- Web browsing (HTTP): five simultaneous models with servers on the gateway side and clients on the terminal side.
File transfer (FTP): two simultaneous sources are generated. An FTP return link control channel runs in parallel to adapt FTP data rate to the capacity of the link.

Video streaming of MPEG-4/MPEG-2 contents.

Audio application: 10 simultaneous users with and without talk spurt/silence.

As shown in Figure 44 end-to-end IP statistics are correct as long as the application data rate is below the instantaneous link capacity: timing and jitter are relatively small for the satellite link, no packet losses occur.

In the case of a fading event, where the MODCOD becomes more robust due to the ACM loop, the capacity of the link decreases and may become less than the application bit rate. In Figure 45 example:

- the link is lost around 1000 s because SNIR is below the MODCOD 1 threshold (unavailability state);
- the loss of a few IP packets before and after the event corresponds to the limited capacity with MODCOD 1;
- all other MODCODs provide sufficient capacity for the link, which is error-free and with reasonable jitter and timing.
QoS mechanisms have shown a correct behaviour when testing VoIP applications (high priority) together with background traffic. During fading events, generating congestion, the VoIP stream has been correctly transmitted, whereas background traffic packets in excess were discarded.

5. SUMMARY AND CONCLUSIONS

This paper describes the design and the capabilities of a comprehensive laboratory testbed, developed under European Space Agency funding. Using this testbed, one is able to assess and optimize the performance of a two-way hub and spoke communication system exploiting the DVB-S2 ACM over a multi-beam satellite payload. The testbed includes faithful traffic emulation as well as satellite channel modelling in terms of satellite nonlinearity, propagation fading and interference. The key design drivers for the various elements of the testbed have been provided with particular emphasis on ACM adaptation, DVB-S2 demodulator implementation aspects and satellite channel modelling. A realistic study case based on a Ka-band 100 beams satellite system has been described and exploited for the dimensioning of system and physical layer key parameters. Key performance results obtained by exploiting the ACM testbed

Figure 45. Example of video streaming application IP statistics during a deep fading event (with congestion).
confirmed the previous analytical and simulation findings and demonstrated the important performance gains that can be achieved by implementing ACM in practical satellite networks. In particular, the ACM adaptation loop mechanisms have been extensively validated and optimized through synthetic, yet realistic fading time series. ACM spectral efficiency gain with practical implementation for various MODCOD sets has been assessed and shown to be remarkably high compared with CCM approach. Nonlinearity impact has been evaluated for typical satellite TWTA characteristic and gain obtained by using simple static pre-compensation derived by measurement. As expected, phase noise impact has been shown to be particularly critical not only for 32-APSK but also for high coding rates. A weakness of the rate 2/3 FEC has been observed in the presence of phase noise. Finally, the impact of ACM on some IP application has been measured.

The next tests with the ACM modem demonstrator are being performed as part of a different ESA project. This project, the DVB-S2 Satellite Experiment project [32], assesses the performance of DVB-S2 in a real environment, i.e. over a real satellite link. The results obtained in the ACM modem project, partly reported in this paper, are used as a reference for comparison.

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**AUTHORS’ BIOGRAPHIES**

**Hermann Bischl** received the Dipl.-Ing. degree in electrical engineering from the Technical University Munich in 1989, and the Dr.-Ing. degree from the University of the Federal Armed Forces, Munich, in 1994. Since 1989 he has been with the Digital Networks section of the Institute of Communications and Navigation of the German Aerospace Center (DLR). He is currently head of the Satellite Communications group. In 1997 he received an innovations award from the Society of Friends of DLR and in 2001 the DGLR Lectureship Award in the area space on the occasion of the German Aerospace Congress 2001. He participated in many national and international research projects related to satellite communications. His main research interests include satellite communication protocols and standards (e.g. DVB-S2/RCS), medium access control, resource management, error control, channel modelling, and system dimensioning.
Hartmut Brandt graduated in Computer Science from the Novosibirsk Institute for Electrotechnical Engineering in 1988. Following a research at the Institute’s Laboratory for Laser Interferometry he joined the Commerzbank in Frankfurt, Germany, to work on expert systems for financing. In 1996 he started to work at the Fraunhofer Institute for Open Communication Systems (FOKUS) in Berlin, Germany, in research on ATM networks and real-time multimedia communications. Since 2004 he is with the Institute for Communications and Navigations at the German Aerospace Center and works in DVB-S2/RCS, channel modeling, traffic modeling, protocol design, implementation and testing and on TCP over satellite performance.

Tomaso de Cola was born in Manosque, France, on April 28, 1977. He received the “Laurea” degree (summa cum laude) in telecommunication engineering from the University of Genoa, Genoa, Italy, in 2001 and the Qualification degree as Professional Engineer in 2002. From 2002 until 2007, he has worked with the Italian Consortium of Telecommunications (CNIT), University of Genoa Research Unit, as scientist researcher. Since 2008, he has been with the German Aerospace Centre (DLR), where he is involved in different European Projects focusing on different aspects of DVB standards, CCSDS protocols and testbed design. He is co-author of more than 20 papers, including international conferences and journals. His main research activity concerns: TCP/IP protocols, satellite networks, transport protocols for wireless links, interplanetary networks as well as delay tolerant networks.

Riccardo De Gaudenzi was born in Italy in 1960. He received his Doctor Engineer degree (cum Laude) in electronic engineering from the University of Pisa, Italy in 1985 and the PhD from the Technical University of Delft, The Netherlands in 1999. From 1986 to 1988 he was with the European Space Agency (ESA), Stations and Communications Engineering Department, Darmstadt (Germany) where he was involved in satellite telecommunication ground systems design and testing. In particular, he followed the development of two new ESA’s satellite tracking systems. In 1988, he joined ESA’s Research and Technology Centre (ESTEC), Noordwijk, The Netherlands where in 2000 he has been appointed head of the Communication Systems Section and since 2005 he is Head of the RF Payload and Systems Division. The division is responsible for the definition and development of advanced satellite system, subsystems and technologies for telecommunications, navigation and earth observation applications. In 1996 he spent one year with Qualcomm Inc., San Diego U.S.A., in the Globalstar LEO project system group under an ESA fellowship. His current interest is mainly related with efficient digital modulation and access techniques for fixed and mobile satellite services, synchronization topics, adaptive interference mitigation techniques and communication systems simulation techniques. From 2001 to 2005 he has been serving as Associate Editor for CDMA and Synchronization for IEEE Transactions on Communications. He is co-recipient of the VTS Jack Neubauer Best System Paper Award from the IEEE Vehicular Technology Society.
Ernst Eberlein is Head of the Communications Department at Fraunhofer IIS. He graduated as Electrical Engineer (Dipl.-Ing.) from the University of Erlangen-Nürnberg in 1985. He worked from 1987 until 1994 in the audio and multimedia department of Fraunhofer IIS. He was member of the core team, which developed the audio coding scheme MP3. He was involved in the development of prototypes for the EU147 DAB system and the deployment of the first pilot networks in Germany. Since 1995 Ernst Eberlein has been focusing on the transmission parts of the Satellite and terrestrial broadcasting systems. From 1995 to 1997 he was in charge for the receiver development for the WorldSpace system. In the years 1997 to 1999 he was responsible for the XM satellite radio system development and validation at Fraunhofer IIS. Since 2001 Ernst Eberlein is responsible for the coordination of the research activities in the area of digital communication. In 2008 he was appointed head of the Communications Department.

Nicolas Girault joined Astrium in 1997 to design and develop on-board processor algorithms for regenerative satellites. He is currently responsible of the telecommunication systems department validation group and laboratory, and leads several activities experimenting new broadband solutions by satellite.

Eric Alberty has more than 25 years experience in satellite telecommunications. He joined Astrium Satellite in 1989 and he has been in charge of several VSAT projects for professional customers in Europe and Africa. In 1999, he took the responsibility of system engineer for the Astrium WeB/WEST initiative for developing an innovative broadband satellite system. He is currently the head of the Telecommunication System Engineering at Astrium Satellite. Before joining Astrium, he was consulting engineer to foreign telecommunications agencies for large earth stations and domestic networks deployment.

Stefan Lipp received his diploma in Electrical and Electronics Engineering from the Friedrich-Alexander University of Erlangen-Nürnberg, Germany in 1997. In 1997, he joined the Communications Department of Fraunhofer Institute for Integrated Circuits (Erlangen, Germany), where he has been working on projects related to digital satellite and terrestrial broadcasting. His research interests are in the area of digital signal processing and system design and analysis.
Rita Rinaldo has obtained her Master Degree in Telecommunications Engineering (summa cum laude) from the University of Bologna in March 2000; in May 2005 she has received her PhD with a thesis titled “Adaptive Coding and Modulation Techniques for Broadband Satellite Systems”. From June 2000 to January 2001 she was with TiLab, the technology research centre of TelecomItalia. Since February 2001 she has been with the Communication System Section in the Technical and Quality Management Directorate of the European Space Agency. Her interests are mainly in the fields of communication satellite system design, physical layer algorithms and resource optimization techniques.

Bjarne Rislow was born in 1964, he has a MSc in Electrical engineering at the Norwegian Institute of Technology (N.I.T.) in 1988. Worked as research scientist at SINTEF (1989-December 1992), ABB Corporate Research (1993–1994), Nera Research (1995–2006) and then as Senior Research Engineer in Thrane & Thrane from August 2006 until May 2008. He now works as private consultant. His main occupation has been with satellite communication and has worked on several system studies and development projects for the Inmarsat System (Inmarsat M/B, mini-M, M4, BGAN) and for DVB-RCS. Further he has been working on various research and development projects related to radio link communications and on wireless broadband access systems, including LMDS. He participated in the early phase of the DVB-RCS standardisation and on various ESA projects including acting as project leader for the Phase 1 of the ACM project.

Dr John Arthur J Skard received his PhD in Theoretical High Energy Physics in 1973 from the University of California, Santa Barbara. As an Assistant Professor at University of Bergen (Norway) and University of Maryland (U.S.A.) he was involved in High Energy research projects at DESY (Hamburg) and at CERN (Geneva). From 1987 to 1995 he worked on scientific software and data analysis for NASA’s COBE (Cosmic Background Explorer) satellite. Since 1995 Dr. Skard has been employed by Nera Satcom/Nera Broadband Satellite, where his main work and interest has been the quality and optimization of communications via satellite.

Jacky Tousch received his MSc degree in telecommunications and electronics from ENST Bretagne, Brest, France, in 1999. In 1999, he cofounded TurboConcept SAS, France, where he is currently the Chief Technical Officer. He participated in the development of more than 20 commercial products implementing turbo codes and LDPC encoder/decoder modules, in satellite communications and broadband wireless. He is the holder of a patent on LDPC receiver architecture. In 2007, TurboConcept joined Newtec Cy, a leading provider of satcom solutions and services. He is now responsible for the IP Core product line within Newtec.
Gerald Ulbricht received his Dipl.-Ing. degree in electrical engineering from the Friedrich-Alexander University Erlangen, Germany in October 1990. After University he joined the Fraunhofer Institute for Integrated Circuits (Fraunhofer IIS) in Erlangen, designing RF transceiver front-ends for digital audio broadcasting (DAB), and for telemetry systems. Between 1995 and 1998 he designed CATV wide band amplifier modules at Temic GmbH and transmitters for GSM base stations at Ericsson Eurolab in Nuremberg. 1998 he returned to the RF and Microwave department of the Fraunhofer IIS as a leader of the design group “Radiocommunicationsystems”. His main research interests are the design of multi-standard RF frontends and linear power amplifiers, as well as the linearization of RF power amplifiers.