The very short frame of mobile DVB-RCS: Code design and QoS Performance

Gianluigi Liva1,*,†, Marco Papaleo2, Cristina Párraga Niebla1, Stefano Cioni2, Sandro Scalise1, Alessandro Vanelli-Coralli2, Gioyanni E. Corazza2, Pansoo Kim3 and Ho-Jin Lee3

1DLR (German Aerospace Center), Institute of Communications and Navigation, Wessling, Germany
2DEIS-ARCES, University of Bologna, Viale Risorgimento, 2-40136 Bologna, Italy
3ETRI (Electronics and Telecommunication Research Institute), 305-350 Deajeon, Republic of Korea

SUMMARY

In this paper, the design of a new very short frame introduced in the upcoming version of DVB-RCS standard targeting interactive mobile services will be described. The frame design with a special focus on channel coding and performance are addressed in detail, as well as an evaluation of the QoS performance for realistic traffic scenarios, propagation channel conditions, and carrier sizes. Copyright © 2009 John Wiley & Sons, Ltd.

KEY WORDS: LDPC codes; mobile DVB-RCS; IRA codes; protographs; interactive services

1. INTRODUCTION

A new version of the DVB-RCS, targeting interactive mobile services, has been recently approved by the Digital Video Broadcasting (DVB) forum. The new standard, DVB-RCS+M in the following [1], foresees several new functionalities aimed at ensuring high performance operation in challenging propagation environments such as aeronautical, maritime, railway, and vehicular mobile scenarios. In this framework, a new very short frame hereafter referred to as 4k mode has been designed and included in the standard, including also newly designed Low-Density Parity-Check (LDPC) codes [2]. The DVB-RCS+M standard employs two different access schemes in the return link: a classical Multi-Frequency Time Division Multiple Access (MF-TDMA) and a new Continuous Carrier (CC) mode. The so-called CC scheme allows the allocation of a fixed carries to a single Return Channel via Satellite (RCS) terminal transmitting in a DVB-S2 [3] based mode. This new access scheme is meant for the provision of interactive services to mobile terminals collecting a moderate-to-high aggregated traffic in the order of some few hundred kbps. The original DVB-S2 waveform was designed for much broader carriers, hence envisaging long coding blocks that can deliver large data blocks within delay margins that are acceptable also for interactive applications. In DVB-RCS+M practical scenarios, where lower data rates are considered, the use of such coding blocks implies, depending on

*Correspondence to: Gianluigi Liva, Institute of Communications and Navigation – DLR, Postfach 1116, 82230 Wessling, Germany.
†E-mail: Gianluigi.Liva@dlr.de

Copyright © 2009 John Wiley & Sons, Ltd.
the used modulation, long transmission times that might be critical for interactive services, as seen in Figure 1. To avoid these long delays that can reduce the perceived Quality of Service (QoS) by the final user, a shorter frame, not foreseen in the DVB-S2 standard, has been introduced and new short LDPC codes designed. Details on the code design and on the framing structure, as well as a performance assessment at both physical and data link layer are presented in this work.

The paper is organized as follows. In Section 2 some insights on the code design and on the baseband framing including the definition of the pilot distribution will be presented. Section 3 will provide numerical results concerning the performance of the coding scheme on Additive White Gaussian Noise (AWGN) channel (with ideal and non-ideal carrier phase estimation) as well as simulation results for mobile (Rice) fading channels. Section 4 compares the performance of the very short frame with the normal DVB-S2 frame in terms of QoS metrics for exemplary traffic scenarios, propagation channel conditions, and carrier sizes. The concluding remarks follow in Section 5.

2. DESIGN OF THE VERY SHORT FRAME

The design of the new family of codes has been carried out following a list of requirements, which is listed next:

1. The code design shall provide performance close to the theoretical limits. The target is the random coding bound [4] for \( n = 4096 \) block codes. The design shall provide performance within 1 dB from the bound.

Figure 1. DVB-S2 frame duration versus bandwidth for different modulations and a roll-off factor of 0.35. The 64k block is here assumed.
2. The codes shall have very low error floors, that is preferably below a codeword error rate of $10^{-7}$.
3. The code design shall share some commonalities with the current DVB-S2 codes. This would allow a smooth upgrade of the DVB-S2 chipsets to support the new codes.
4. The design shall lead to decoders having a high degree of parallelism.

It is indeed well-known that the first two requirements are somehow in contradiction. The design of iteratively decodable codes with good waterfall performance usually leads to high error floors. On the other hand, achieving low error floors requires a sacrifice in terms of iterative decoding threshold, and hence in the code’s waterfall performance. Good compromises in this sense can be found in [5–7]. Dealing with requirement 3, the search of the candidate codes was narrowed to the ensemble of the Irregular Repeat-Accumulate (IRA) [8] codes, on which the DVB-S2 standard is based. Finally, to answer to the requirement 4, the code design has been based on a block-circulant parity-check matrix form with circulants of size 128 [9]. Following this approach, three IRA codes have been designed, with $n = 4096$ and (exact) code rates $1/4$, $1/2$, and $3/4$. Let us now summarize the main features of the new codes.

2.1. Parity-check matrix structure

The parity check matrices of the codes have been derived by a proper protograph [10] expansion. A protograph is a relatively small bipartite graph from which a larger graph can be obtained by a copy-and-permute procedure: the protograph is copied $q$ times, and then the edges of the individual replicas are permuted among the $q$ replicas (under some restrictions described below) to obtain a single, large bipartite graph. Suppose the protograph possesses $n_p$ variable nodes and $m_p$ constraint nodes. Then the derived graph will consist of $n = n_p q$ variable nodes and $m_c = m_p q$ constraint nodes. Note that the edge permutations cannot be arbitrary. In particular, the nodes of the protograph are labeled so that if Variable Node (VN) $V_j$ is connected to Check Node (CN) $C_i$ in the protograph, then VN $V_j$ in a replica can only connect to one of the $q$ replicated CNs $C_i$. Doing so preserves the decoding threshold properties of the protograph while permitting the design of quasi-cyclic codes. In particular, if the edge permutations are organized in a cyclic manner such that the final adjacency matrix is an array of circulant permutation matrices, the code will be quasi-cyclic. Protographs impose a structure on the derived graph, which facilitates the design of fast decoders and efficient encoders, as well as a refined control on the derived graph edge connections. Consider an LDPC code ensemble $C_{\lambda, \rho}$ defined by the degree distribution pair $\lambda(x), \rho(x)$ [11]. A protograph $G$ with these degree distribution defines the sub-ensemble of $C_{\lambda, \rho}$ comprising the LDPC codes with the edge connections imposed by the protograph structure. This refined definition allows performing a further selection within $C_{\lambda, \rho}$. In particular, it is possible to restrict the code search in the protograph sub-ensemble with the best characteristics in terms of decoding threshold [12].

The protograph expansion has been developed in two steps. In the first one, small expansion parallel edges have been eliminated, leading to an intermediate protograph with $n_p = 32$ VNs and $m_p = 24$, 16, and 8 respectively for the rate $1/4$, $1/2$, and $3/4$ codes. Then, a final protograph expansion has been performed by replacing each edge of the protograph with a $128 \times 128$ circulant permutation matrix. The permutation selection followed the girth optimization technique of [13]. In a first proposal design, the protographs for the IRA ensembles are shown in Figure 2. The iterative decoding thresholds of the ensembles have been computed by
protograph Extrinsic Information Transfer (EXIT) analysis [14]. Note that the choice of a constant information variable node degree \( d_v = 5 \) is not necessarily optimal in terms of iterative decoding threshold. However, for moderate-short code lengths, design that are merely based on the iterative decoding threshold of the asymptotic \( (n \to \infty) \) ensemble tend to lead to high error floors [15, 16]. By slightly relaxing the iterative decoding threshold requirement, it is however possible to achieve low error floors with a limited loss in the waterfall region [5].

In a second design stage, the new rate 1/4 code was introduced. The new design came from the need of aligning more strictly the code structure to that of the DVB-S2 LDPC code. The codes adopted by the DVB-S2 standard possess a constant check node degree (i.e. their parity-check matrices have uniform row Hamming weight). While for the rate 1/2 and the rate 3/4 codes presented above the check node degree was constant, the original rate 1/4 code presents a

Figure 2. Protographs for the IRA code ensembles adopted in the DVB-RCS+M standard. The rate 1/4 protograph represents the original design, replaced in a later stage by the one provided in Figure 3: (a) protograph for the rate 3/4 code; (b) protograph for the rate 1/2 code; (c) protograph for the rate 1/4 code.
non-uniform distributions (i.e. 2/3 of the check nodes has degree 4, while 1/3 has degree 3). Therefore a new rate 1/4 design was introduced, which allows a constant check node degree by adopting an irregular variable nodes distribution. The protograph of the new ensemble is depicted in Figure 3. The new protograph presents an iterative decoding threshold that outperforms one of the former ensemble \( (E_b/N_0 = -0.0410 \text{ dB} \) for the new ensemble, while for the former one \( E_b/N_0 = 0.205 \text{ dB} \)). Such a gain is reflected in a gain of \(~0.2 \text{ dB} \) in the waterfall region, while the error floor requirement has to be slightly relaxed.\(^1\) As it will be shown in Section 3, the new ensemble shows a flattening of the Codeword Error Rate (CER) curve slightly above \( 10^{-7} \).

2.2. Error floor estimation

The error floor has been estimated following the strategy proposed in [9]. This method is based on the union bound and exploits the structure of the parity check matrix. In particular, this analysis fully applies to a subclass of IRA codes, which can be defined by the three parameters, \( k, m, \) and \( w_c \), namely the number of information bits, the number of parity check equations, and the weight of the first \( k \) columns of the parity check matrix (hereafter referred to as matrix \( H_1 \)). As a consequence, this method can be applied only if the variable node degree is constant over \( H_1 \). Since this condition is satisfied only for the rate 1/2 and 3/4 (in which \( w_c = 5 \)), the error floor of the rate 1/4 case will be computed by means of numerical simulations.

The union bound on the bit error rate is given by

\[
P_b \sum_{d>0} D_d Z^d \bigg|_{Z = e^{-E_b/N_0}}
\]

\(^1\)Recall the usual trade-off between error floor and waterfall performance. Irregular LDPC ensembles allows gains in the waterfall region, but they tend to suffer by error floors, especially when the block size is moderate or small.

Copyright © 2009 John Wiley & Sons, Ltd.  
DOI: 10.1002/sat
where
\[
D_d < \sum_{w:h=k=d} \frac{w}{k} \overline{A}_{w,h}
\] (2)

Following the notation introduced in [9], \(\overline{A}_{w,h}\) indicates the average input–output weight enumerator (IOWE) over the IRA code ensemble, \(w\) and \(k\) are the Hamming weights of information and parity block, respectively. The IOWE of the IRA ensemble can be computed as:
\[
\overline{A}_{w,h} = \sum_l \overline{A}_{w,l}(1) \frac{A_{l,h}^{(\text{acc})}}{\binom{m}{l}}
\] (3)

\(\overline{A}_{w,l}(1)\) is the average IOWE of the outer code corresponding to \(H^T_1\), while \(A_{l,h}^{(\text{acc})}\) is the IOWE of the accumulator part of the IRA code ensemble, and it can be computed as:
\[
A_{l,h}^{(\text{acc})} = \binom{m-h}{\lfloor l/2 \rfloor} \binom{h-1}{\lfloor l/2 \rfloor - 1}
\] (4)

where \(l\) represents the Hamming weight of the outer code output (that is the input of the accumulator), and \(h\) is the accumulator output weight. The hard task in the computation of Equation (3) is the evaluation of the term \(\overline{A}_{w,l}(1)\), which in general can be expressed as:
\[
\overline{A}_{w,l}(1) = \binom{k}{w} P_{w,l}
\] (5)

\(P_{w,l}\) is the so-called input output weight transfer probability (IOWTP). Its computation represents a challenging task since all possible combinations of \(w\) and \(l\) must be evaluated. Nevertheless, since for high Signal-to-Noise Ratio (SNR) values the major contribution to the bit error rate is given by the low-weight encoder inputs, only small values of \(w\) can be taken into account. In particular, following [9] we will consider only \(w=1\) and \(w=2\). Under these assumptions we compute \(P_{1,w}\), and \(P_{2,w}\). Considering \(w=1\), the output of the outer code will be a row of \(H^T_1\), which means \(P_{1,l}=1\) for \(l=w_c\), and \(P_{1,l}=1\) for \(l \neq w_c\). Further, since no \(4 - \text{cycles}\) exist in \(H^T_1\), if the encoder input has weight 2 (\(w=2\)), two possible values of \(l\) can appear, that are \(2w_c\) and \(2(w_c-1)\). If we call \(N_{w,l}\) the multiplicity of output codewords of the outer code with weight \(l\) given input weight \(w\), \(P_{w,l}\) can be written using the following expression:
\[
P_{w,l} = \frac{N_{w,l}}{\sum_p N_{w,p}}
\] (6)

For the case of \(w=2\), we have
\[
N_{2,w_c} = \binom{m}{w_c} \binom{m-w_c}{w_c} / 2
\] (7)

and
\[
N_{2,(w_c-1)} = \binom{m}{w_c} \binom{m-w_c}{w_c-1} / 2
\] (8)
Since $P_{1,l}$ and $P_{2,l}$ dominate the performance in the error floor region, we will omit all the terms corresponding to $w \geq 3$. The results of this analysis are reported in Figure 4. The curves refer to the rate $1/2$ ($k = 2048$, $m = 2048$, $w_c = 5$) and rate $3/4$ ($k = 3072$, $m = 1024$, $w_c = 5$). As it can be seen, both codes present very low error floor fulfilling requirement number 2.

2.3. Outer BCH codes

As for the DVB-S2 standard, an outer algebraic code has been included in our design. The outer code shall take care of removing low-weight, residual error patterns that may appear at the output of the LDPC decoder. Such small error patterns usually lead to high error floors, and are due either to low-weight codewords or to the so-called trapping sets [17]. While our design already gives an answer to the first issue (i.e. the code’s minimum distance shall be quite large, according to the choice of a constant information VN degree equal to 5 [9]), the second issue is not fully under the control of the code designer. Trapping sets are a function of the decoder’s implementation. The inclusion in the coding scheme of an outer BCH code plays the role of an insurance with respect to sub-optimal decoder implementations. The BCH code selection has been carried out trading error correcting capabilities with the additional overhead. For the $(4096, 3072)$ rate-$3/4$ IRA code, the selected outer code is a $(3072, 3000)$ 6-errors correcting code obtained by shortening of a $(4095, 4023)$ BCH. The outer code of the $(4096, 2048)$ rate-$1/2$ IRA

Figure 4. Error Floor Performance for rate 1/2 and rate 3/4 codes computed following [9].
code is a (2048, 1992) 5-errors correcting code obtained by extension of a (2047, 1992) BCH code. The outer code for the (4096, 1024) rate-1/4 IRA code is an (1024, 993) extended BCH code with capability of correcting three errors. The overhead introduced by the outer concatenation can be measured in terms of $E_b/N_0$ penalty as 0.103 dB in the first case, 0.120 dB in the second one, and 0.133 dB in the last one.

2.4. Baseband framing and pilots distribution

The CC mode in DVB-RCS+M maintains the framing structure of the classical DVB-S2 standard [3]. In order to ensure that the new 4k mode can correctly fit into the structure of the DVB-S2 framing, a few adaptations are in order.

The output of the LDPC encoder, referred as FECFRAME in DVB-S2 terminology, can be modulated according to one of the possible four DVB-S2 modulation formats, that is QPSK, 8PSK, 16APSK, or 32APSK. This clearly requires the FECFRAME length to be a multiple of 2, 3, 4, or 5. Two/four padding bits are therefore introduced for 8PSK/32APSK modulation respectively. This leads to a modulated frame, that is XFECFRAME (complex-FECFRAME), of length 2048, 1366, 1024, or 820 complex symbols for QPSK, 8PSK, 16APSK, or 32APSK, respectively. In DVB-S2, the XFECFRAME symbols are organized into slots of 90 complex symbols each. In the case of pilot symbol insertion, which is optional in DVB-S2 but mandatory in DVB-RCS+M, 36 pilot symbols are inserted every 16 slots in order to form a physical layer frame (PLFRAME). This requires the XFECFRAME length to be a multiple of 90 symbols, which is not the case for the 4k codes. Therefore, additional pilot symbols are appended at the end of the last XFECFRAME slot so as to correctly complete the PLFRAME. These additional pilot symbols could be seen as a postamble that can be used to ease the synchronization and channel estimation algorithms at the receiver. This is essential for the 16APSK and 32APSK modulations for which, being the XFECFRAME shorter than 16 slots, the postamble is the only available pilot field in the PLFRAME.

3. PERFORMANCE OF THE LDPC CODES

In Figure 5, numerical results obtained through computer simulation are presented. The curves show the Bit Error Rate (BER) and CER performance on the AWGN channel, versus $E_b/N_0$, with $E_b$ being the energy per information bit and $N_0$ the one-sided noise power spectral density. The performance refers to the concatenation of the BCH code with the proposed IRA codes, and are obtained using several code rates and modulation schemes. In particular, the following modulation-coding pairs have been taken into account: QPSK 1/4, QPSK 1/2, QPSK 3/4, 8PSK 3/4, 16APSK 3/4, and 32APSK 3/4.

Figure 6 shows the performance of the 8PSK 3/4 modulation-coding scheme in the presence of Rice fading. The terminal speed is assumed to be 100 km/h, while the symbol rate is equal to 256 kbaud. The channel is modeled as a Ricean frequency flat fading channel with Rice factor equal to 17.4 dB. In this analysis we evaluate the effectiveness of adopting a bit interleaving (present in DVB-S2 but not in DVB-RCS+M) in such a typical scenario. As it can be noted in the figure, the interleaver does not yield any performance enhancement. This can be easily explained by taking into account the channel coherence time and the FECFRAME duration,

---

2The rate assumed in the modulation-coding pair is the LDPC code rate. The overall system code rate is slightly lower because of the concatenation with the BCH code.
which are 19.2 and 5.33 ms, respectively. The channel coherence time results to be rather larger than the physical FECFRAME duration, indicating that no time diversity can be achieved within the codeword. It is obvious that under this conditions, the adoption of a bit interleaver working on a single FECFRAME does not allow to improve the CER and BER.

Next, the performance in the presence of phase noise and non ideal carrier phase estimation is investigated. The phase tracking circuit is required due to the presence of residual frequency errors and phase noise generated by the local oscillator. The carrier phase recovery is a fundamental step in the achievement of correct receiver synchronization to obtain the maximum data decoding performance.

The DVB-S2 frame structure contains a preamble and pilot fields, which could be used in the synchronization procedure. The proposed scheme for phase recovery is a combination of data-aided (DA) and decision feedback (DF) solutions. The DVB-S2 preamble, identified as PLHEADER, is composed of 90 known symbols (26 from the SOF field and 64 from the Modulation and Code rate (ModCod) information), which are used to initialize the phase synchronizer with the Maximum Likelihood (ML) estimate:

\[
\hat{\theta}_0 = \arg\left\{ \sum_k r_k \cdot c_k^* \right\}
\]

Figure 5. Performance of the different modulation-coding schemes on the AWGN channel.
where $r_k$ contains the received preamble sequence and $c_k$ the known symbols. This DA feed-forward estimation can be considered for the initial phase acquisition and then, a feedback structure is implemented for carrier phase tracking exploiting the detector decisions, $\hat{c}_k$, in place of known data. Implementing the classical phase locked-loop, the governing equations are [18]:

$$\hat{\theta}_{k+1} = \hat{\theta}_k + \xi_k$$  

$$\xi_k = \xi_{k-1} + \alpha[\beta e(k) - e(k-1)]$$  

$$e(k) = \text{Im}\{r_k \xi_k^* e^{-j\hat{\theta}_k}\}$$

where $\alpha$ and $\beta$ are the loop gain parameters. To enhance the tracking performance, the decisions $\hat{c}_k$ are substituted with the known symbols when the pilot field is processed by the demodulator.

The BER performance of the proposed LDPC code in the presence of the carrier phase noise is summarized in Figure 7. The phase noise has been generated according to the frequency mask reported in the DVB-S2 standard [3]. For these results, the loop gain parameters have been set.
to $\alpha = 0.005$ and $\beta = 1.00001$, respectively. It shall be noted that the performance loss is in the order of 0.2 dB for both the analyzed modulation-coding schemes.

4. ASSESSMENT OF QOS SUPPORT FOR DVB-RCS CONTINUOUS CARRIER MODE

The introduction of a very short frame for the CC mode of DVB-RCS for mobile users was justified by the fact that the use of DVB-S2 (normal) frames in narrow-band channels might have a negative impact on the QoS performance due to the inherent long transmission delay of a the frame. This section is focused on comparing the QoS performance of the DVB-S2 normal frame (64 800 bits, hereafter referred to as 64k) and the very short frame (4096 bits), in exemplary traffic scenarios.

4.1. System model

We consider one collective user terminal operating in CC mode, accessing one carrier that is allocated for this user uniquely. The user terminal collects the return link traffic of several end users and transmits the cumulative traffic through this dedicated carrier. The collective terminal
The model applied for the simulations can be observed in Figure 8. The terminal has two traffic sources, each generating IP traffic of one priority class: QoS 2 and QoS 3, which represent Assure Forwarding (AF) and Best Effort (BE) traffic. The generated packets are queued in priority queues and scheduled according to a very simple priority rule: the packet scheduler attempts to empty the QoS 2 queue. It serves the QoS 3 queue only if there are no backlogged packets in the QoS 3. The scheduled packets are packed in frames using Generic Stream Encapsulation (GSE), which allows flexible fragmentation of IP packets to match the available payload length in the frame [19]. The payload length of the BBFrame depends in turn on the used ModCod. The ModCod for transmission of each frame is selected according to the propagation conditions experienced by the terminal.

![Figure 8. Collective terminal model in the return link.](image)

![Figure 9. SNIR time series used for QoS simulations.](image)
For the traffic generation, a substantial traffic aggregation has been assumed, as in [20], since the scope of the CC mode involves collective terminals only. Hence, a very simple traffic model has been adopted for simulations: an exponential packet interarrival time with variable

Table I. 4k frame parameters taken from [20].

<table>
<thead>
<tr>
<th>ID number</th>
<th>ModCod</th>
<th>Payload (bits)</th>
<th>Efficiency</th>
<th>Minimum SNIR (dB)</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>QPSK 1/4</td>
<td>913</td>
<td>0.4157</td>
<td>−1.6</td>
</tr>
<tr>
<td>2</td>
<td>QPSK 1/2</td>
<td>1912</td>
<td>0.8707</td>
<td>1.3</td>
</tr>
<tr>
<td>3</td>
<td>QPSK 3/4</td>
<td>2920</td>
<td>1.3297</td>
<td>4.3</td>
</tr>
<tr>
<td>4</td>
<td>8PSK 3/4</td>
<td>2920</td>
<td>1.9085</td>
<td>8.3</td>
</tr>
<tr>
<td>5</td>
<td>16APSK 3/4</td>
<td>2920</td>
<td>2.4957</td>
<td>10.7</td>
</tr>
</tbody>
</table>

Table II. 64 k frame parameters taken from [20].

<table>
<thead>
<tr>
<th>ID Number</th>
<th>ModCod</th>
<th>Payload (bits)</th>
<th>Efficiency</th>
<th>Minimum SNIR (dB)</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>QPSK 1/4</td>
<td>15928</td>
<td>0.4902</td>
<td>−1.6</td>
</tr>
<tr>
<td>2</td>
<td>QPSK 1/2</td>
<td>32128</td>
<td>0.9889</td>
<td>1.3</td>
</tr>
<tr>
<td>3</td>
<td>QPSK 3/4</td>
<td>48328</td>
<td>1.4875</td>
<td>4.3</td>
</tr>
<tr>
<td>4</td>
<td>8PSK 3/4</td>
<td>48328</td>
<td>2.2281</td>
<td>8.3</td>
</tr>
<tr>
<td>5</td>
<td>16APSK 3/4</td>
<td>48328</td>
<td>2.9667</td>
<td>10.7</td>
</tr>
</tbody>
</table>

Figure 10. Cumulative input rate.
packet length. Each generated packet is assigned one of the following packet lengths with its associated probability:

1. 40 bytes with probability 0.3
2. 552 bytes with probability 0.15
3. 576 bytes with probability 0.15
4. 1500 bytes with probability 0.4

The packet sizes have been selected to emulate the most commonly IP packet sizes found in Internet, which result for example from the Maximum Transfer Unit (MTU) of Ethernet networks and TCP acknowledgments. The selected probabilities are exemplary. For each QoS class, an input rate of 8 packets per second has been assumed.

The terminal is associated to a propagation region. This is a geographical region characterised by common fading conditions. For simulating variable fading conditions, an in-house fading simulator available at DLR has been used, which simulates rain fading events, interference, and scintillation. This attenuation value subtracted to a link margin in line of sight and clear sky conditions of 12 dB gives a time series of Signal-to-Noise plus Interference Ratio (SNIR) versus the simulation time. For the sake of simplicity, the effects of mobility in the propagation conditions are not considered in this paper and left for further work. A snapshot of the SNIR time series applied to the simulations are shown in Figure 9. The selected time window for the

![Empirical CDF of queuing delay](image)

**Figure 11.** CDF of queuing delay in scenario 1.
simulations features SNIR variations such that the terminal shall switch ModCod several times during the simulation time.

The use of GSE as encapsulation scheme has been selected to overcome the fact that for the lowest coding rate defined for the 4k frame, the maximum payload length is smaller than an MPEG Transport Stream (MPEG-TS) and the other available coding rates can transport an integer number of MPEG-TS only at the cost of very high padding. Hence, if MPE/MPEG encapsulation would be used, MPEG-TS fragmentation would be required, in which case GSE is more flexible and is more efficient in terms of encapsulation overhead [21].

The DVB-RCS+M implementation guidelines [20] proposes a set of 18 ModCods for the CC mode. Nevertheless, only three coding rates have been defined for the 4k coding block, namely 1/4, 1/2, and 3/4. Hence, in order to compare the performances of both frame types, we select a subset of ModCod that can be used with both frame types. These ModCod and their characteristics, including the Base-Band Frame (BBFrame) payload length, are shown in Table I for the 4k frame and in Table II for the 64k frame. It should be noted that the spectral efficiencies have been calculated without including the expected average GSE encapsulation efficiency, and for that reason it does not match the assumed efficiencies in [20]. In turn, SNIR values are taken from [20].

Finally, a packet dropping function is also implemented by dropping each packet in the queues that waited longer than 5 s to be transmitted to avoid buffer overflow (or infinity queuing times in congestion situations).

![Empirical CDF of queuing delay](image)

Figure 12. Cumulative Distribution Function (CDF) of queuing delay in scenario 2.
Simulations have been performed using 4k and 64k frames in two scenarios described below:

1. Scenario 1: carrier of 150 kHz
2. Scenario 2: carrier of 200 kHz

assuming in both cases a roll-off factor of 0.2. With the packet input rate mentioned above for each QoS class, the resulting aggregated instantaneous input rate is shown in Figure 10 together with the instantaneous maximum achievable MAC layer throughput for scenario 1 (dashed line) and the maximum achievable MAC layer throughput for scenario 2 (dotted line). The maximum achievable throughput depends on the most efficient ModCod that can be used in each instant of time (which depends on the instantaneous SNIR) and on the carrier bandwidth. The average input rate through the simulation time is 98.4 kbps, while the average maximum achievable throughput is 135.38 kbps for scenario 1 and 180.51 kbps for scenario 2, so in average, enough capacity is available for the input traffic. Nevertheless, the ModCod switches during the simulations to more robust ModCod during significant amount of time, causing that the channel capacity falls below the offered traffic. The results presented next thus show how the different frame types also behave in congestion situations.

4.2. Numerical results

The following metrics have been analyzed from the obtained simulation results: queuing delay, GSE overhead, and frame padding. Figures 11 and 12 show the empirical CDFs of the queuing delay.
delay for scenarios 1 and 2, respectively. Each of the figures shows four curves, one per QoS queue and frame type. In general, it can be observed that the queuing delay is kept lower in all cases for the 4k frame, while the QoS class 2 has better queuing delay performance than the QoS class 3 due to the selected prioritization policy. This can be especially observed in scenario 1, where lower capacity is available. When the system enters into congestion due to a deep fade, the scarce available capacity is completely used to serve the QoS 2 queue, causing that QoS 3 packets wait in their queues until they are dropped. This effect is shown in Figure 13, where the instantaneous delay performance of each QoS class is shown. In scenario 2, the queuing delay performance is much more stable due to the fact that more capacity is available through a wider carrier. Observing the behavior in scenario 1, it is recommendable to have a dynamic system where higher carrier bandwidth can be assigned on demand when the queuing delay (or the packet drops) increase and go back to a narrower carrier when possible. As it is defined, the DVB-RCS+M standard can support such a dynamic carrier allocation.

The selection of a suitable carrier bandwidth depending on the fading conditions of the terminal and the offered traffic is of high importance for achieving a cost-efficient use of the resources. On the one hand, if the carrier is too narrow, the queuing delay performance can be very poor. On the other hand, a too wide carrier might cause that the frames are often empty (full of padding bits) or carry only a few user bits. This effect can be observed in Figures 14 and 15, which show the experimental CDF of padding bits per frame for 4k and 64k frames in scenarios 1 and 2, respectively.

In scenario 1, it can be observed that the 4k frames have more padding than the 64k frames. In fact, the padding performance in both scenarios is identical for the 4k frames. This is due to

![Empirical CDF of frame padding bits](image)

**Figure 14.** CDF of padding bits in scenario 1.
the fact that when the terminal is not in congestion, the frame transmission is fast in comparison with the cumulative packet input rate. In other words, the scheduler finds the queues empty in several occasions when the terminal is not congested and thus the padding in 4k frames can vary from zero to the maximum payload length of a 4k frame. For the 64k frames, this is not the case, since the transmission time of a frame is long in comparison with the cumulative input rate. Hence, the scheduler can always fetch packets from any queue in the narrow carrier scenario. In contrast, scenario 2 offers too much capacity in general (when the ModCod used are number 2 or higher) for the input rate. In this case, the number of padding bits per frame can vary from zero to the maximum payload length of a 64k frame, as can be seen in Figure 15. Therefore, the padding performance of the 64k bits is only better if the terminal is congested. In a general case, the maximum padding bits of a 4k frame is around four times lower than the 64k frame.

Finally, the fact that higher layer packets are encapsulated using GSE also has an impact on the resulting layer-2 overhead. In general, one can expect that the GSE overhead shall be higher for 4k frames, since IP datagrams must be eventually fragmented in several GSE packets, for example in the case of 1500 bytes packets, with the corresponding GSE header per fragment. In the case of 64k, more packets can be put as a whole with a GSE header in the frame, reducing the encapsulation overhead. The simulation results confirm this prediction, as can be observed in Figures 16 and 17, which show the experimental GSE overhead per frame for each QoS class and frame type in scenarios 1 and 2, respectively. The discontinuity found for the 64k frame is

![Empirical CDF of frame padding bits](image_url)

**Figure 15. CDF of padding bits in scenario 2.**
due to the fact that in a number of occasions, the last packet in a frame must be fragmented in two pieces because it does not fit in the remaining available bits within the current BBFrame. Since the packet is fragmented to match the remaining bits of the BBFrame, it happens that some times one of the fragments has a very small payload in comparison with the header, which increases significantly the GSE overhead in the frame. From now on in this paper, this effect is referred to as marginal fragmentation. It should also be noted that in scenario 1, due to the congestion, only a few QoS 3 packets are transmitted and for those, there was no such critical fragmentation. In a non-congested scenario, the same behavior due to marginal fragmentation is also observed for QoS 3. In the case of 4k frames, the marginal fragmentation effect does not appear, since packet fragmentation in general produces fragments of similar size due to the short length of the frame.

5. CONCLUSION

This work described in detail all features of the very short frame of DVB-RCS+M and provide thorough experimental results on its performance. The new LDPC family introduced in this paper produces a coded block length of 4096 coded bits, which has been organized into a DVB-S2 like framing structure so as to be used in the CC mode of DVB-RCS+M. The new LDPC codes show no floor down CER $\approx 10^{-7}$ in both AWGN and Rice channels. Thanks to these features, the proposed LDPC code has been included in the DVB-RCS+M standard, in
addition to the classical DVB-S2 modes. In terms of QoS performance, the 4k frame has a better performance in terms of delay and padding than the 64k frame, at the cost of a slightly higher GSE overhead. Moreover, in congestion situations, the 4k frame can manage better the traffic, as can be seen in Figure 13, where the queuing delay CDF of the best effort traffic is much more stable than for 64k frames.

ACKNOWLEDGEMENTS

The authors thank Lars Erup for the useful discussions during the design of the very short frame within TM-RCS standardisation group.

REFERENCES


AUTHORS’ BIOGRAPHIES

Gianluigi Liva was born in Spilimbergo, Italy, on July 23, 1977. He received the MS degree in Electrical Engineering in 2002, and the PhD degree in 2006 at DEIS, University of Bologna (Italy). His main research interests include satellite communication systems and error control coding (with emphasis on LDPC codes and Turbo-like codes) for wireless fading channels. Since 2003 he has been involved in the research of near Shannon limit channel codes for high data rate CCSDS (Consultative Committee for Space Data Systems) missions, in collaboration with the European Space Operations Centre of the European Space Agency (ESA-ESOC). From October 2004 to April 2005 he was researching at the University of Arizona, where he was involved in the design of low-complexity coding systems for high data rate Mars links. He is currently with the Institute of Communications and Navigation, at the German Aerospace Center (DLR), in Wessling, where he is involved in the research of FEC and modulation techniques for mobile satellite systems. He is active in the DVB-SH and in the DVB-RCS Mobile standardization groups. He is IEEE member and he serves IEEE as reviewer for Transactions, Journals and Conferences. He received the 2007 IST Mobile and Wireless Communication Summit Best Paper Award.
Marco Papaleo was born in Soverato (Italy) in 1981. He received his BS and MSc degrees (summa cum laude) in Telecommunication Engineering from University of Bologna, Italy, in 2003 and in 2006, respectively. From January 2006, he joined the Advanced Research Center on Electronic Systems for Information and Communication Technologies ‘Ercole De Castro’ (ARCES) at the University of Bologna, where he started his PhD studies in 2007. In summer 2008 he was a visiting PhD student at the Institute of Communications and Navigation of the German Aerospace Center (DLR), in Wessling, where he was involved in the design and analysis of LDPC convolutional codes. In 2006 he was a visiting affiliate student at the University College of London (UCL), London. His research activities are mainly focused on the next generation wireless telecommunication systems, both the terrestrial and the satellite networks. In particular, his interests include design and performance evaluation of error control coding with emphasis on packet level coding.

Cristina Párraga Niebla graduated in Telecommunications Engineering from Universidad Politécnica de Catalunya (UPC), Barcelona, Spain, in July 2002. She joined the German Aerospace Centre (DLR) in 2001, from 2002 as research scientist. Her research activities deal with resource management for satellite systems, especially for DVB-S2/RCS systems, and aeronautical communications over satellite both for passengers and air traffic management. She is co-author of many international journal and conference papers, co-organizer of the 4th Advanced Satellite Mobile Systems Conference, and co-editor of a book chapter devoted to resource management in satellite networks.

Stefano Cioni received the Dr Ing degree in telecommunication engineering and the PhD from University of Bologna, Italy, in 1998 and in 2002, respectively. From March 2002 to October 2002, he was a visiting researcher at the European Space Agency (ESA) on adaptive coding and modulation (ACM) techniques for future broadband satellite networks. In 2002, he joined the ARCES where he is currently a Post-Doc researcher. During the summer 2006 he was a Visiting Researcher at the Agilent Labs SMRD, Belgium. During the summer 2007 he was a Visiting Researcher at the German Aerospace Center (DLR), Oberpfaffenhofen (Germany). His research activities are mainly focused on the next generation wireless telecommunication systems, both the terrestrial and the satellite networks. In particular, his interests include synchronization techniques, medium access control resource allocation algorithms, OFDM systems, and iterative decoding techniques joint to channel parameter estimation. Dr Cioni co-authored more than 40 papers and scientific conference contributions and he is a co-recipient of the Best Paper Award at IEEE ICT 2001.

Sandro Scalise graduated in Electronic Engineering specializing in Telecommunications (with honors) from University of Ferrara (Italy) in July 1999 and obtained his PhD from University of Vigo (Spain) in July 2007. Since 2001, he is within the Institute for Communications and Navigation, DLR (German Aerospace Center), Germany, where he is currently the Head of the Digital Networks Department. His research activity deals with forward error correction and synchronization schemes for mobile satellite applications, land mobile satellite channel modeling, and link performance evaluation. He is a co-author of many international journal and conference papers, co-chairman of the 3rd and 4th Advanced Satellite Mobile Systems Conference and editor of a book chapter devoted to satellite channel impairments.
Alessandro Vanelli-Coralli received the Dr Ing Degree (cum laude) in Electronics Engineering and the PhD in Electronics and Computer Science from the University of Bologna (Italy) in 1991 and 1996, respectively. In 1996, he joined the Department of Electronics, Computer Science and Systems (D.E.I.S.) at the University of Bologna where he is currently an Assistant Professor. Since 2001, he has been a Research Associate of the Advanced Research Center for Electronic Systems (ARCES) of the University of Bologna. During 2003 and 2005, he was a Visiting Scientist at Qualcomm Inc. (San Diego, CA). He participates in national and international research projects on satellite mobile communication systems and he is a Vice-Chairman of the research and development group of the Integral SatCom Initiative (ISI) technology platform. His research interests are in the area of wireless communication systems, physical layer techniques, and digital signal processing. He has been a co-editor of Special Issues of the Wiley International Journal on Satellite Communications and Network, and of Special Issues of the EURASIP Journal on Wireless Communications. Dr Vanelli Coralli has been the Technical Program Committee Chairman for the 4th IEEE ASMS2008 and Vice-Chairman for the IEEE ISSSTA 2008 Conference. Dr Vanelli-Coralli co-authored more than 100 papers and scientific conference contributions and is a co-recipient of the Best Paper Award at IEEE ICT 2001 and at IEEE ISWCS 2005. He is an IEEE Senior Member.

Giovanni E. Corazza is a Full Professor at DEIS, University of Bologna, and leads Wireless Communications inside the Advanced Research Centre for Electronic Systems (ARCES). He was a Chairman of the School for Telecommunications in the years 2000–2003, Chairman of the Advanced Satellite Mobile Systems Task Force (ASMS TF), Founder and Chairman of the Integral Satcom Initiative (ISI), a European Technology Platform. Since 1997, he is a Editor for Communication Theory and Spread Spectrum for the IEEE Transactions on Communications. He is the author of more than 200 papers, and received the Marconi International Fellowship Young Scientist Award in 1995, the 2002 IEEE VTS Best System Paper Award, the Best Paper Award at IEEE ISSSTA’98, at IEEE ICT2001, and at ISWCS 2005. He has been the General Chairman of the IEEE ISSSTA 2008, ASMS 2004, ASMS 2006, ASMS 2008 Conferences. His research interests are in wireless and satellite communications, estimation and synchronization, spread spectrum and multi-carrier transmission, upper layer coding, navigation, and positioning.

Pansoo Kim received his BS and MS degrees in the ECE from SungKyunKwan University, Korea in 2000 and 2002. Since 2002, he has worked for ETRI, Deajeon, Korea, as a senior member of engineering staff in the field of standard activities and system implementation of stationary/mobile DVB-RCS/S2 communication/broadcasting system. His main research interests are in digital satellite broadcasting/communication system and modem technology with associated equalization, synchronization and iterative decoder design and VLSI implementation.

Ho-Jin Lee received his BS, MS, and PhD degrees in the EE from Seoul National University, Korea in 1981 and 1983 and 1990. Since 1983, he has been with ETRI, Deajeon, Korea, and was involved with TDX development, satellite communication and broadcasting technology research programs. He has been leading a number of research projects, including KOMPSAT-1 ground mission control system development, DVB-RCS VSAT, a couple of broadband mobile VSATs, and S-DMB enhancement feasibility study. He is currently the director of satellite and wireless convergence research department. His interests are in VSATs, B3G satellite mobile communications, satellite broadcasting, and HAPS communications.