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**Digital Video Broadcasting (DVB);  
Implementation Guidelines for a second generation  
digital cable transmission system (DVB-C2)**

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## Foreword

This Technical Specification (TS) has been produced by Joint Technical Committee (JTC) Broadcast of the European Broadcasting Union (EBU), Comité Européen de Normalisation ELECTrotechnique (CENELEC) and the European Telecommunications Standards Institute (ETSI).

**NOTE:** The EBU/ETSI JTC Broadcast was established in 1990 to co-ordinate the drafting of standards in the specific field of broadcasting and related fields. Since 1995 the JTC Broadcast became a tripartite body by including in the Memorandum of Understanding also CENELEC, which is responsible for the standardization of radio and television receivers. The EBU is a professional association of broadcasting organizations whose work includes the co-ordination of its members' activities in the technical, legal, programme-making and programme-exchange domains. The EBU has active members in about 60 countries in the European broadcasting area; its headquarters is in Geneva.

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The Digital Video Broadcasting Project (DVB) is an industry-led consortium of broadcasters, manufacturers, network operators, software developers, regulatory bodies, content owners and others committed to designing global standards for the delivery of digital television and data services. DVB fosters market driven solutions that meet the needs and economic circumstances of broadcast industry stakeholders and consumers. DVB standards cover all aspects of digital television from transmission through interfacing, conditional access and interactivity for digital video, audio and data. The consortium came together in 1993 to provide global standardization, interoperability and future proof specifications.

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## Modal verbs terminology

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## Introduction

DVB-C2 is a standard for the digital transmission of digital signals in broadband cable and cable television systems commonly referred to CATV networks. It defines techniques of the physical layer (e.g. error protection, interleaving, modulation) and lower layer protocols required for data packaging and signalling. Compared to its predecessor DVB-C, which was originally standardized in 1994, DVB-C2 offers significant benefits with regard to transmission performance (e.g. spectral efficiency) and operational flexibility (e.g. variable bandwidth, improved ability to adapt to dedicated channel conditions).

Further background information on DVB-C2 is given in the subsequent scope of the present document. Clauses 2 and 3 provide the reader with information on references (normative and informative references) and definitions, symbols and abbreviations respectively, which are used for the description of the technology in the main part of the present document and thus are helpful for the understanding of the related explanations.

Clauses 4 to 6 give general information about DVB-C2 to enable an understanding of the general concept followed during the development of the DVB-C2 standard [i.1]. Commercial Requirements are introduced, which were created to steer the technical development. An overview on the DVB-C2 system summarizes its key features. Clause 5 describes the anatomy of the signal on the level of the physical and the logical frame structures. Clause 6 justifies why dedicated parameter sets were chosen for individual elements of the DVB-C2 standard [i.1].

The detailed description of individual DVB-C2 elements starts in clause 7 with the Input Processing and the Multiplex Structure. Mechanisms for bundling of the so-called Physical Layer Pipes - a concept of transparent delivery of data streams from the transmitting to the receiving end, the common use of global information by a number of **PLPs** (Common **PLP**), stuffing algorithms and the second layer (L2) signalling are explained in great detail.

Clause 8 complements the description of the encoding techniques defined in the DVB-C2 standard [i.1] which are: Preamble Generation, Pilot Structure, Peak to Average Power Ratio (PAPR), two level (L1 and L2) signalling structure, Interleaving, Framing, OFDM and Spectral Shaping. These techniques are implemented for instance in devices such as modulators, edgeQAMs, etc.

In clause 9, the reader receives information about the delivery media, e.g. the cable network and its transmission characteristics. This clause is not intended to be an all-embracing pool of information about all kinds of CATV network infrastructures, however it provides a thorough understanding of the major characteristics of the transport media which are important to understand when developing hardware or software systems compliant with DVB-C2.

While the receiving end is not discussed by the DVB-C2 standard [i.1], clause 10 gives guidelines being tailored to the particular needs of implementers of CPE and other devices/units designed to receive DVB-C2 signals. The following techniques are explained in particular: synchronization procedure, Time and Frequency De-interleaving, use of Pilots, phase-noise requirements, mechanism for tuning to a **Data Slice**, buffer management, **FECFrame** header detection, LDPC and BCH decoding, output processing and power saving.

Having in mind the information on the entire transmission chain provided by the preceding clauses, the transmission performance supported by the DVB-C2 signal is analysed in clause 11. Next to theoretical estimations, simulation results are presented.

Last but not least clause 12 contains information about realistic examples on how DVB-C2 could be operated in CATV networks. Various network scenarios are described and possible modifications of today's headend architecture are introduced. Information on possible migration scenarios towards DVB-C2 systems complements the described scenarios.

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# 1 Scope

The present document gives guidelines for the implementation of DVB-C2 based cable transmission systems. The DVB-C2 standard [i.1] contains a detailed and precise description of techniques used at the transmitting end. The target of the DVB-C2 standard document [i.1] has been to provide an instruction manual allowing a non-ambiguous implementation of the technology. Explanations on transmission aspects and matters of receiver implementation have not been subject to standardization. Since the developers of the DVB-C2 standard [i.1] felt that such information would be essential for an accurate implementation of an end-to-end DVB-C2 system, the present document was prepared covering the following topics:

- Descriptions of techniques implemented at the transmitting end complementing the explanations given by the DVB-C2 standard [i.1] and extending the focus area to matters of headend architectures, for instance.
- Explanations on CATV network aspects to an extent important for an implementation of a DVB-C2 device.
- Guidelines for the implementation of equipment installed at the receiving end.

DVB-C2 was developed in coincidence with the DVB philosophy to make use of state-of-the-art technologies if possible rather than to invented technologies with almost no advantages compared to existing ones. However, a number of elements of DVB-C2 are based on newly invested techniques which have been used neither in the first generation DVB family of transmission systems nor in the standards of the second generation agreed prior to DVB-C2 (e.g. DVB-S2, DVB-T2), nor have been implemented in a comparable manner in any other transmission technology known. Techniques of the combined **PLP** and **Data Slice** multiplex concept are an example for such a novelty. These and other techniques had to be invented to ensure that DVB-C2 not only meets its commercial and technical requirements, but provides an optimized solution with regard to operational flexibility and transmission efficiency to an extent possible today. These techniques have not been discussed at the time of publication of this paper in the wider literature but are presented in the present document in a detail which is important to know for an implementer.

The present document has the objective to make available to DVB-C2 implementers as much as possible of the common understanding captured during the work of the experts group developing the DVB-C2 standard [i.1]. The present document was prepared by these experts with the intention to provide know-how complementary to the explanations of the DVB-C2 standard [i.1] itself and about the environment in which a DVB-C2 system will be operated. Network aspects and matters of receiver implementation are mentioned by way of an example.

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## 2 References

### 2.1 Normative references

References are either specific (identified by date of publication and/or edition number or version number) or non-specific. For specific references, only the cited version applies. For non-specific references, the latest version of the reference document (including any amendments) applies.

Referenced documents which are not found to be publicly available in the expected location might be found at <http://docbox.etsi.org/Reference>.

NOTE: While any hyperlinks included in this clause were valid at the time of publication, ETSI cannot guarantee their long term validity.

The following referenced documents are necessary for the application of the present document.

Not applicable.

## 2.2 Informative references

References are either specific (identified by date of publication and/or edition number or version number) or non-specific. For specific references, only the cited version applies. For non-specific references, the latest version of the reference document (including any amendments) applies.

NOTE: While any hyperlinks included in this clause were valid at the time of publication, ETSI cannot guarantee their long term validity.

The following referenced documents are not necessary for the application of the present document but they assist the user with regard to a particular subject area.

- [i.1] ETSI EN 302 769: "Digital Video Broadcasting (DVB); Frame structure channel coding and modulation for a second generation digital transmission system for cable systems (DVB-C2)".
- [i.2] ETSI EN 300 429: "Digital Video Broadcasting (DVB); Framing structure, channel coding and modulation for cable systems".
- [i.3] ETSI EN 302 755: "Digital Video Broadcasting (DVB); Frame structure channel coding and modulation for a second generation digital terrestrial television broadcasting system (DVB-T2)".
- [i.4] ETSI TS 102 831: "Digital Video Broadcasting (DVB); Implementation guidelines for a second generation digital terrestrial television broadcasting system (DVB-T2)".
- [i.5] ETSI EN 302 307 (all parts): "Digital Video Broadcasting (DVB); Second generation framing structure, channel coding and modulation systems for Broadcasting, Interactive Services, News Gathering and other broadband satellite applications".
- [i.6] ETSI EN 300 421: "Digital Video Broadcasting (DVB); Framing structure, channel coding and modulation for 11/12 GHz satellite services".
- [i.7] ETSI TS 102 606 (all parts): "Digital Video Broadcasting (DVB); Generic Stream Encapsulation (GSE)".
- [i.8] ETSI TS 102 771: "Digital Video Broadcasting (DVB); Generic Stream Encapsulation (GSE) implementation guidelines".
- [i.9] ETSI EN 300 468: "Digital Video Broadcasting (DVB); Specification for Service Information (SI) in DVB systems".
- [i.10] ISO/IEC 13818-1: "Information technology - Generic coding of moving pictures and associated audio information: Systems".
- [i.11] ETSI EN 300 744: "Digital Video Broadcasting (DVB); Framing structure, channel coding and modulation for digital terrestrial television".
- [i.12] J. van de Beek, M. Sandell, and P. O. Borjesson: "ML estimation of time and frequency offset in OFDM systems," IEEE Transactions on Signal Processing, Vol. 45, No. 7, Jul. 1997, pp. 1800 - 1805.
- [i.13] Stott, J. H., Summer 1998: "The effects of phase noise in COFDM", EBU Technical Review, (276), pp. 12 -25.

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- [i.16] Thomas J. Kolze: "HFC Channel Model Submission", IEEE 802.14a/98-12, May 1998.
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- [i.19] "A new cable frequency plan and power deployment rules", ReDeSign Deliverable D14, 2009.  
NOTE: Available at: [www.ict-redesign.eu](http://www.ict-redesign.eu).
- [i.20] "The HFC channel model", ReDeSign Deliverable D8, 2008.  
NOTE: Available at: [www.ict-redesign.eu](http://www.ict-redesign.eu).
- [i.21] "Methodology for specifying HFC networks and components", ReDeSign Deliverable D10, 2009.
- [i.22] Thomas Proakis: "Digital Communications", New York Mc Graw-Hill, ISBN 0-07-232111-3.
- [i.23] Jaeger, D.; Schaaf, C. (editors): DVB-C2: High Performance Data Transmission on Cable - Technology, Implementation, Networks -. Mitteilungen aus dem Institut für Nachrichtentechnik der Technischen Universität Braunschweig, Band 13, Shaker Verlag, Aachen 2010).
- [i.24] NorDig Unified Requirements for Integrated Receiver Decoders v2.3 (2012) .
- [i.25] ETSI TS 103 287: "Digital Video Broadcasting (DVB); Modulator Interface (C2-MI) for a second generation transmission system for cable systems (DVB-C2)".
- [i.26] ETSI TS 101 211: "Digital Video Broadcasting (DVB); Guidelines on implementation and usage of Service Information (SI)".

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## 3 Definitions, symbols and abbreviations

### 3.1 Definitions

For the purposes of the present document, the terms and definitions given in [i.1] and the following apply:

**0xkk:** digits 'kk' should be interpreted as a hexadecimal number

**active cell:** OFDM Cell carrying a constellation point for L1 signalling or a **PLP**

**auxiliary data:** sequence of cells carrying data of as yet undefined modulation and coding, which may be used for stuffing **Data Slices** or stuffing **Data Slice packets**

**BBFrame:** signal format of an input signal after mode and stream adaptation

**BBHeader:** header in front of a baseband data field

NOTE: See clause 5.1.

**C2 frame:** fixed physical layer TDM frame that is further divided into variable size **Data Slices**

NOTE: **C2 frame** starts with one or more **preamble symbols**.

**C2 Modulator Interface (C2-MI):** interface between a centralized and decentralized part of a DVB-C2 modulator

NOTE: **C2-MI** is specified in ETSI TS 103 287 [i.25].

**C2 system:** complete transmitted DVB-C2 signal, as described in the **L1-part 2** block of the related **preamble symbols**

**common PLP:** special **PLP**, which contains data shared by multiple **PLPs (Transport Stream)**

**data cell:** OFDM Cell which is not a pilot or tone reservation cell

**data PLP:** **PLP** carrying payload data

**Data Slice:** group of OFDM Cells carrying one or multiple **PLPs** in a certain frequency sub-band

NOTE: This set consists of OFDM Cells within a fixed range of consecutive cell addresses within each **data symbol** and spans over the complete **C2 frame**, except the **preamble symbols**.

**Data Slice packet:** **XFECFrame** including the related **FECFrame** Header

**data symbol: OFDM symbol** in a **C2 frame** which is not a **preamble symbol**

**Dependent Static Data Slice (DSDS): Data Slice** between broadband **notches** or **C2 system** edges with a bandwidth smaller than 3 408 **OFDM cells**

**div:** integer division operator, defined as:

$$x \text{ div } y = \left\lfloor \frac{x}{y} \right\rfloor$$

**elementary period:** time period which depends on the channel raster and is used to define the other time periods in the C2 System

**FECFrame:** set of  $N_{\text{LDPC}}$  (16 200 or 64 800) bits of one LDPC encoding operation

NOTE: In case of **Data Slice** carrying a single **PLP** and constant modulation and encoding is applied, **FECFrame** Header information may be carried in Layer1 part 2 and the **Data Slice** Packet is identical with the **XFECFrame**.

**FFT size:** nominal FFT size for a DVB-C2 receiver is 4 K

NOTE: Further details are discussed in clause 10.1.

**for i=0..xxx-1:** when used with the signalling loops, this means that the corresponding signalling loop is repeated as many times as there are elements of the loop

NOTE: If there are no elements, the whole loop is omitted.

**guard-interval fraction:** ratio  $T_g/T_u$  of the guard interval duration to the active **symbol period**

**Im(x):** imaginary part of x

**Layer 1 (L1):** name of the first layer of the DVB-C2 signalling scheme (signalling of physical layer parameters)

**L1 block:** set of **L1-part 2** COFDM Cells, cyclically repeated in the frequency domain

NOTE: **L1 blocks** are transmitted in the **preamble symbols**.

**L1-part1:** signalling carried in the header of the **Data Slice packets** carrying modulation and coding parameters of the related **XFECFrame**

NOTE: **L1-part1** parameters may change per **XFECFrame**.

**L1-part2:** Layer 1 Signalling cyclically transmitted in the preamble carrying more detailed L1 information about the **C2 system**, **Data Slices**, **Notches** and the **PLPs**

NOTE: **L1-part2** parameters may change per **C2 frame**.

**Layer 2 (L2):** name of the second layer of the DVB-C2 signalling scheme (signalling of transport layer parameters)

**mod:** modulo operator, defined as:

$$x \text{ mod } y = x - y \left\lfloor \frac{x}{y} \right\rfloor$$

**mode adapter:** input signal processing block, delivering **BBFrames** at its output

**nn<sub>D</sub>:** digits 'nn' should be interpreted as a decimal number

**notch:** set of adjacent OFDM Cells within each **OFDM symbol** without transmitted energy

**null packet:** MPEG Packet with the Packet\_ID 0x1FFF, carrying no payload data and intended for padding

**OFDM cell:** modulation value for one OFDM carrier during one **OFDM symbol**, e.g. a single constellation point

**OFDM symbol:** waveform  $T_S$  in duration comprising all the active carriers modulated with their corresponding modulation values and including the guard interval

**Physical Layer Pipe (PLP):** logical channel carried within one or multiple **Data Slice(s)**

NOTE 1: All signal components within a **PLP** share the same transmission parameters such as robustness, latency.

NOTE 2: A **PLP** may carry one or multiple services. In case of **PLP Bundling** a **PLP** may be carried in several **Data Slice**. Transmission parameters may change each **XFECFrame**.

**PLP bundling:** transmission of one **PLP** via multiple **Data Slice**

**PLP\_ID:** this 8-bit field identifies uniquely a **PLP** within a **C2 system**

**preamble header:** fixed size signalling transmitted in the first part of the Preamble, carrying the length and Interleaving parameters of **Layer 1 part 2** data

**preamble symbol:** one or multiple **OFDM symbols**, transmitted at the beginning of each **C2 frame**, carrying Layer 1 part 2 signalling data

**Re(x):** real part of  $x$

**reserved for future use:** value of any field indicated as "reserved for future use" is set to "0" unless otherwise defined

**START\_FREQUENCY:** index of lowest used OFDM subcarrier of a **C2 system**

NOTE: The value of **START\_FREQUENCY** is a multiple of  $D_X$ .

**Transport Stream:** multiplex signal as defined in ISO/IEC MPEG Systems specification ISO/IEC 13818-1 [i.10]

$x^*$ : Complex conjugate of  $x$

**XFECFrame:** **FECFrame** mapped onto QAM constellations:

- $\lfloor x \rfloor$ : round towards minus infinity: the most positive integer less than or equal to  $x$ .
- $\lceil x \rceil$ : round towards plus infinity: the most negative integer greater than or equal to  $x$ .

## 3.2 Symbols

For the purposes of the present document, the symbols given in ETSI EN 302 769 [i.1] and the following apply:

$\oplus$	exclusive OR / modulo-2 addition operation
$\Delta$	Absolute guard interval duration
$A$	LDPC codeword of size $N_{ldpc}$
$\lambda_i$	LDPC codeword bits
$\lambda^{RM}$	32 output bits of Reed-Muller encoder
$\lambda_i^{RM}$	Bit number of index $i$ of 32 bit long output bits of Reed-Muller encoder
$\eta_{MOD}, \eta_{MOD}(i)$	Number of transmitted bits per constellation symbol (for <b>PLP</b> $i$ )
$\pi_p$	Permutation operator defining parity bit groups to be punctured for L1 signalling
$\pi_s$	Permutation operator defining bit-groups to be padded for L1 signalling
$A_{m,l}$	Output vector of the frequency interleaver of <b>OFDM symbol</b> $l$ and <b>C2 frame</b> $m$
$A_{CP}$	Amplitude of the continual pilot cells
$A_{PP}$	Amplitude of the preamble pilots
$A_{SP}$	Amplitude of the scattered pilot cells
$a_{m,l,q}$	Frequency-Interleaved cell value, cell index $q$ of symbol $l$ of <b>C2 frame</b> $m$
$B(n)$	Location of the first <b>data cell</b> of symbol $l$ allocated to <b>Data Slice</b> $n$ in the frequency interleaver
$b$	16 bit long <b>FECFrame</b> signalling data vector

$b_{e,do}$	Output from the demultiplexer, depending on the demultiplexed bit sub-stream number $e$ and the input bit number $d_i$ of the bit interleaver demultiplexer
$b_i$	Bit number of index $i$ of 16 bit long <b>FECFrame</b> signalling data vector
$C/N$	Carrier-to-noise power ratio
$C/N+I$	Carrier-to-(Noise+Interference) ratio
$C_i$	Column of index $i$ of time interleaver
$c_i$	Column of index $i$ of bit interleaver
$c(x)$	Equivalent BCH codeword polynomial
$c_{m,l,k}$	Cell value for carrier $k$ of symbol $l$ of <b>C2 frame</b> $m$
$C_{RM}$	Correlation function of the received RM code vector
$\text{dB}\mu\text{V}$	relative logarithmic signal level with reference to $1\mu\text{V}$
$DFL$	Data field length
$D_P$	Difference in carrier index between adjacent preamble-pilot-bearing carriers
$D_X$	Difference in carrier index between adjacent scattered-pilot-bearing carriers
$D_y$	Difference in symbol number between successive scattered pilots on a given carrier
$d(x)$	Remainder of dividing message polynomial by the generator polynomial $g(x)$ during BCH encoding
$d_i$	Input bit number $d_i$ of the bit interleaver demultiplexer
$d_o$	Bit number of a given stream at the output of the demultiplexer of the bit interleaver
$e$	Demultiplexed bit sub stream number ( $0 \leq e < N_{\text{substreams}}$ ), depending on input bit number $d_i$ of the bit interleaver demultiplexer
$f_q$	Constellation point normalized to mean energy of 1
$G$	Reed-Muller encoder matrix
$g(x)$	BCH generator polynomial
$g_1(x), g_2(x), \dots, g_{12}(x)$	Polynomials to obtain BCH code generator polynomial
$g_q$	Complex cell of index $q$ of a <b>Data Slice packet</b>
$H(q)$	Frequency interleaver permutation function, element $q$
$I$	Output codeword of BCH encoder
$i_j$	BCH codeword bits which form the LDPC information bits
$j$	$\sqrt{-1}$
$K_{\text{bch}}$	Number of bits of BCH uncoded block
$K_i$	L1 signalling part 2 parameter selected as $N_{L1part\ 2}(K_i) \leq N_{L1part\ 2\_Cells} \times \eta_{\text{MOD}}$
$K_{\text{ldpc}}$	Number of bits of LDPC uncoded block
$K_{L1}$	3408 (number of OFDM carriers per L1 block)
$K_{L1\_PADDING}$	Length of L1_PADDING field
$K_{L1part\ 2}$	Length of <b>L1-part 2</b> signalling field including the padding field
$K_{L1part\ 2\_ex\_pad}$	Number of information bits in <b>L1-part 2</b> signalling excluding the padding field.
$K_{N,min}$	Lowest frequency carrier index of a frequency <b>notch</b>
$K_{N,max}$	Highest frequency carrier index of a frequency <b>notch</b>
$K_{sig}$	Number of signalling bits per FEC block for L1 signalling part 2
$K_{min}$	Lowest frequency carrier index of a C2 signal
$K_{max}$	Highest frequency carrier index of a C2 signal
$K_{total}$	Number of OFDM carriers per OFDM symbol
$k$	Absolute OFDM carrier index
$L_{data}$	Number of data <b>OFDM symbols</b> per <b>C2 frame</b> (excluding Preamble)
$L_F$	Number of OFDM Symbols per <b>C2 frame</b> including excluding <b>preamble</b>
$L_P$	Number of preamble OFDM Symbols within the <b>C2 frame</b>
$l$	Index of <b>OFDM symbol</b> within the <b>C2 frame</b>
$l_d$	Index of data <b>OFDM symbol</b> within the <b>C2 frame</b>
$l_p$	Index of preamble <b>OFDM symbol</b> in the <b>C2 frame</b>
$m$	<b>C2 frame</b> number
$m(x)$	Message polynomial within BCH encoding
$m_i$	Input bit of index $i$ from uncoded bit vector $M$ before BCH encoder



$M$	Uncoded bit vector before BCH encoder
$M_{max}$	Maximum Sequence length for the frequency interleaver
$N_{bch}$	Number of bits of BCH coded block
$N_{bch\_parity}$	Number of BCH parity bits
$N_c$	Number of columns of bit or time interleaver
$N_{CP}$	Number of continual pilots in 3 408 sub-carriers
$N_{data}$	Number of <b>Data cells</b> in a <b>Data Slice</b> in frequency interleaver
$N_{DP}$	Number of complex cells per <b>Data Slice packet</b>
$N_{DS}$	Number of cells between (KDS,max - KDS,min ) minus the number of continual pilots, scattered pilots, reserved tones and cells that are located in notches within the Data Slice Size if the <b>FFT</b>
$N_{FFT}$	Number of bit-groups for BCH shortening
$N_{group}$	Length of punctured and shortened LDPC codeword for <b>L1-part 2</b> signalling
$N_{L1part\ 2}$	Number of available cells for <b>L1 part 2</b> signalling in one <b>OFDM symbol</b>
$N_{L1part\ 2\_Cells}$	Number of LDPC blocks for the L1 signalling part 2
$N_{L1part\ 2\_FEC\_Block}$	Maximum number of L1 information bits for transmitting the encoded <b>L1 part 2</b> signalling through one <b>OFDM symbol</b>
$N_{L1part\ 2\_max\_per\_Symbol}$	Time interleaving depth for L1 signalling part 2
$N_{L1\_TI\_Depth}$	Intermediate value used in L1 puncturing calculation
$N_{L1part\ 2\_temp}$	Number of bits of LDPC coded block
$N_{ldpc}$	Number of modulated cells per FEC block for the <b>L1-part 2</b> signalling
$N_{MOD\_per\_Block}$	Total number of modulated cells for the <b>L1-part 2</b> signalling
$N_{MOD\_Total}$	Number of BCH bit-groups in which all bits will be padded for <b>L1-part 2</b> signalling
$N_{pad}$	Number of <b>PLPs</b> in a <b>C2-system</b>
$N_{plp}$	Number of LDPC parity bits to be punctured
$N_{punc}$	Number of parity groups in which all parity bits are punctured for L1 signalling
$N_{punc\_groups}$	Intermediate value used in L1 puncturing calculation
$N_{punc\_temp}$	Number of bits in Frequency Interleaver sequence
$N_r$	Number of rows of bit or time interleaver
$N_r$	Number of rows of bit or time interleaver
$N_{substreams}$	Number of substreams produced by the bit-to-sub-stream demultiplexer
$N_{RT}$	Number of reserved carriers
$N_{TR}$	Number of reserved carriers
$n$	<b>Data Slice</b> number
$P_k(f)$	Power spectral density
$p_i$	LDPC parity bits
$Q_{ldpc}$	Code-rate dependent LDPC constant
$q$	<b>Data cell</b> index within the <b>OFDM symbol</b> prior to frequency interleaving and pilot insertion
$R_{eff\_16K\_LDPC\_1\_2}$	Effective code rate of 16K LDPC with nominal rate 1/2
$R_{eff\_L1part\ 2}$	Effective code rate of <b>L1-part 2</b> signalling
$R_i$	Row of index $i$ of time interleaver
$R_i$	Value of element $i$ of the frequency interleaver sequence following bit permutations
$R'_i$	Value of element $i$ of the frequency interleaver sequence prior to bit permutations
$r_i$	Row of index $i$ of bit interleaver
$r_k$	DBPSK modulated pilot reference sequence
$R_S$	Recovered modulator sampling rate
$R_{TS}$	Data rate of the received <b>Transport Stream</b>
$S_0$	List of reserved carriers
$T$	<b>Elementary period</b>
$T_{Ci}$	Column-twist value for column $C$ of time interleaver
$T_{CH}$	Component set of carrier indices for reserved carriers
$T_F$	Duration of one <b>C2 frame</b>

$T_P$	Time interleaving period
$T_S$	Total <b>OFDM symbol</b> duration
$T_U$	Useful <b>OFDM symbol</b> duration
$t$	BCH error correction capability
$t_c$	Column-twist value for column $c$ of bit interleaver
$U$	Parity interleaver output
$UPL$	User Packet Length
$u_i$	Parity-interleaver output bits
$u^{RM}$	32 bit output vector of the cyclic delay block in the <b>FECFrame header</b> encoding
$u_{(i+2) \bmod 32}^{RM}$	Output of the cyclic delay block for input bit $i$ in the <b>FECFrame header</b> encoding
$V$	Column-twist interleaver output
$v_i$	Column-twist interleaver output bits
$v_{m,l,i}$	Output vector of frequency interleaver, starting at carrier index $i$ (= <b>Data Slice</b> start carrier) of the current <b>OFDM symbol</b> $l$ and <b>C2 frame</b> $m$
$v^{RM}$	Scrambled output sequence in the lower branch of the <b>FECFrame header</b> encoder
$v_i^{RM}$	Bit $i$ of scrambled output sequence in the lower branch of the <b>FECFrame header</b> encoder
$w_i$	Bit $i$ of the data scrambling sequence
$w^{RM}$	32 bit scrambling sequence in the lower branch of the <b>FECFrame header</b> encoder
$w_i^{RM}$	Bit $i$ of scrambling sequence in the lower branch of the <b>FECFrame header</b> encoder
$w^P$	Pilot synchronization sequence, build out of $w_i$ and $w'$
$w_k^P$	Bit of index $k$ of pilot synchronization sequence
$w'$	L1 block specific pilot synchronization sequence
$w'_i$	Bit of index $k$ of L1 block specific pilot synchronization sequence
$X_j$	The set of bits in group $j$ of BCH information bits for L1 shortening
$X_{m,l}$	Frequency interleaver input <b>data cells</b> of the <b>OFDM symbol</b> $l$ and the <b>C2 frame</b> $m$
$x$	Address of the parity bit accumulator according to $i_{360}$ in LDPC encoder
$y_{i,q}$	Bit $i$ of cell word $q$ from the bit-to-cell-word demultiplexer
$z_q$	Constellation point prior to normalization

The symbols  $s$ ,  $t$ ,  $i$ ,  $j$ ,  $k$  are also used as dummy variables and indices within the context of some clauses or equations.

In general, parameters which have a fixed value for a particular **PLP** for one processing block (e.g. **C2 frame**, Interleaving Frame, TI-block) are denoted by an upper case letter. Simple lower-case letters are used for indices and dummy variables. The individual bits, cells or words processed by the various stages of the system are denoted by lower case letters with one or more subscripts indicating the relevant indices.

### 3.3 Abbreviations

For the purposes of the present document, the abbreviations given in ETSI EN 302 769 [i.1] and the following apply:

1 024-QAM	1 024-ary Quadrature Amplitude Modulation
16-QAM	16-ary Quadrature Amplitude Modulation
256-QAM	256-ary Quadrature Amplitude Modulation
64-QAM	64-ary Quadrature Amplitude Modulation
8PSK	8-ary Phase Shift Keying
ACI	Adjacent Channel Interference
ACM	Adaptive Coding and Modulation
ADC	Analogue to Digital Converter
ASI	Asynchronous Serial Interface
ATM	Asynchronous Transfer Mode
AWGN	Additive White Gaussian Noise
BB	BaseBand
<b>BBFrame</b>	BaseBand Frame
BCH	Bose-Chaudhuri-Hocquenghem multiple error correction binary block code
BER	Bit error rate
BOS	National (Germany) variant of TETRA (Terrestrial Trunked Radio) services
BPSK	Binary Phase Shift Keying

BUFS	Size of the receiver buffer
BUFSTAT	Actual status of the receiver buffer
BW	BandWidth
C/N	Carrier to Noise ratio
C/N+I	Carrier to noise and intermodulation power ratio
C2dsd	C2 delivery system descriptor
C2SF	C2_System centre frequency
CATV	Community Antenna Television
CCM	Constant Coding and Modulation
CINR	Carrier to Intermodulation Noise Ratio
CIR	Carrier to Intermodulation Ratio
CIRCISO	Carrier-to-intermodulation ratio for CSO (Composite Second Order)
CIRCTB	Carrier-to-intermodulation ratio for CTB (Composite Triple Beat)
CNR	Carrier to Noise Ratio
COFDM	Coded Orthogonal Frequency Division Multiplex
CP	Continual Pilot
CPE	Customer Premises Equipment
CR	Code Rate
CRC	Cyclic Redundancy Check
CSO	Composite Second Order
CTB	Composite Triple Beat
CTO	Composite Third Order
CW	Continuous wave
D	Decimal notation
dB	decibel
DBPSK	Differential Binary Phase Shift Keying
DC	Direct current
DFL	Data Field Length
DFT	Discrete Fourier Transformation
DJB	De-jitter buffer
DNP	Deleted Null Packets
DOCSIS®	Data over Cable Service Interface Specification
DSDS	<b>Dependent Static Data Slice</b>
DSMCC	Digital Storage Media Command and Control
DSTF	<b>Data Slice</b> Tuning Frequency
DVB	Digital Video Broadcasting project
DVB-C	DVB System for cable transmission

NOTE: As defined in ETSI EN 300 429 [i.2].

DVB-C2 DVB-C2 system

NOTE: As specified in ETSI EN 302 769 [i.1].

DVB-S DVB System for digital broadcasting via satellites

NOTE: As specified in ETSI EN 300 421 [i.6].

DVB-S2 Second Generation DVB System for satellite broadcasting

NOTE: As specified in ETSI EN 302 307 [i.5].

DVB-SI DVB Service Information

DVB-T DVB System for Terrestrial broadcasting

NOTE: As specified in ETSI EN 300 744 [i.11].

DVB-T2 Second Generation DVB System for Terrestrial broadcasting

NOTE: As specified in ETSI EN 302 755 [i.3].

DVB-X First generation DVB transmission systems

DVB-X2 Generic abbreviation for the group of 2<sup>nd</sup> generation DVB systems: DVB-C2, DVB-S2 and DVB-T2

EBU	European Broadcasting Union
EC	European Commission
EIT	Event Information Table (DVB SI Table)
EMM	Entitlement Management Message
EN	European Norm
EPG	Electronic program guide
EU	European Union
EWS	Early Warning System
FEC	Forward Error Correction
FFT	Fast Fourier Transformation
FIFO	First In First Out
FM	Frequency Modulation
GCS	Generic Continuous Stream
GFPS	Generic Fixed-length Packetized Stream
GI	Guard Interval
GS	Generic Stream
GSE	Generic Stream Encapsulation
HD	High Definition
HDTV	High Definition Television
HEM	High Efficiency Mode
HFC	Hybrid Fibre Coax
ICI	Inter Carrier Interference
IDFT	Inverse Discrete Fourier Transformation
IDTV	Integrated Digital TV
IEC	International Electrotechnical Commission
IF	Intermediate Frequency
IFFT	Inverse Fast Fourier Transform
IM	InterModulation
IP	Internet Protocol
IPTV	Television via the open Internet (IP)
IPv4	Internet Protocol version 4
IPv6	Internet Protocol version 6
IS	Interactive Services
ISCR	Input Stream Time Reference
ISDB-T	Integrated Services Digital Broadcasting Terrestrial
ISI	Input Stream Identifier
ISSY	Input Stream SYNchronizer
Kbit	$2^{10} = 1\,024$ bits
LDPC	Low Density Parity Check (codes)
LFSR	Linear feedback shift register
LSB	Least Significant Bit
LTE	Long Term Evolution
LUT	LookUp-Table
MAC	Media access control
MBaud	1 000 000 Baud = 1 000 000 Symbols per second
Mbit	$2^{20} = 1\,048\,576$ bits
Mbit/s	Mbit per second
Mega	1 000 000 (1 Million)

NOTE: MBaud is used for single carrier systems, such as DVB-S2, only.

MI	Modulator Interface
MPEG	Moving Pictures Experts Group
MPS	Monic Polynomial Sequence
MSB	Most Significant Bit

NOTE: In DVB-C2 the MSB is always transmitted first.

NA	Not Applicable
NIT	Network Information Table
NL	The Netherlands
NM	Normal Mode

NPD	Null Packet Deletion
OFDM	Orthogonal Frequency Division Multiplex
OSI	Open Systems Interconnection Reference Model
PAL	Phase Alternating Line, name of a analogue TV system
PAPR	Peak to Average Power Ratio
PDF	Probability Density Function
PER	(MPEG TS) Packet Error Rate
PHY	Physical Layer modulation scheme
PLL	Phase locked loop
<b>PLP</b>	Physical Layer Pipe
PLP_ID	PLP Identifier
PN	Pseudo noise
PRBS	Pseudo Random Binary Sequence
PSD	Power Spectral Density
PSI/SI	Program Specific Information/Service Information
PSK	Phase Shift Keying
PVR	Personal Video Recorder
QAM	Quadrature Amplitude Modulation
QEF	Quasi error free
QoS	Quality of Service
QPSK	Quaternary Phase Shift Keying
RF	Radio Frequency
RM	Reed Muller
SD	Standard Definition
SDTV	Standard Definition TV
SI	Service Information
SNR	Signal to Noise Ratio
SP	Scattered Pilot
STB	Set Top Box
TDI	Time de-interleaver
TDM	Time Division Multiplex
TF	Time/Frequency
TI	Time Interleaver
TNO	Dutch telecommunication research institute
TS	<b>Transport Stream</b>
TSPSS	<b>Transport Stream</b> Partial Stream Synchronized
TV	Television
UP	User Packet
UPL	User Packet Length
US	United States of America
VBR	Variable Bit Rate
VCM	Variable Coding and Modulation
VHF	Very High Frequency
VOD	Video On Demand
<b>XFECFrame</b>	XFEC Frame

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## 4 Overview of DVB-C2

### 4.1 DVB-C2 commercial requirements

#### 4.1.0 Introduction

In recent years, customer demands for sophisticated multimedia applications have increased significantly. Broadband Internet peers providing even higher speeds of beyond 100 Mbit/s as well as new TV services such as High Definition Television (HDTV), Back-up TV, Video On Demand (VOD) and other interactive TV services have been included in the service portfolios of cable operators. The provision of all these services side by side via CATV networks to a high number of customers requires the availability of modern infrastructures supporting the transmission of both broadcast and broadband services under severe Quality of Service (QoS) requirements. Also, the continuous transmission of traditional services and of analogue TV in particular has been clearly requested by cable customers mainly for the purpose of providing a compatible service which can be received by second and third TV sets per household without

any need for hardware changes. Today, a modern Hybrid Fibre Coax (HFC) network is an infrastructure making sufficient transmission capacity available. Various technologies are in place which can be used to upgrade traditional CATV networks to state of the art HFC networks. However, network upgrades require huge financial investments and can take a significant time, which both urge operators to carefully adopt the optimal upgrade strategy.

The transmission systems currently deployed in HFC networks, namely DVB-C and DOCSIS<sup>®</sup>, have used physical layer techniques which were invented in 1994 and earlier. Their current applications cannot be configured to support digital transmissions with an increased spectral efficiency. Ongoing investigations have shown that technical advancements of the existing systems, e.g. by introducing a 1 024-QAM, would only result in very limited efficiency improvements while providing non-backwards compatible solutions requires the introduction of a new equipment generation. Also, state of the art upgrade technologies do not provide the feature of improving the spectral efficiency to an extent required for future network upgrades. In contrast, European cable operators have ascertained that under certain commercial and technical conditions, the introduction of equipment supporting an advanced physical layer with capabilities going beyond of what is possible with advanced versions of the existing systems would provide the best means for a successful implementation of new business models. Therefore, the operators approached the DVB Project with the request to launch the development of a new physical layer and lower layer signalling technology based on commercial requirements to be specified. The subsequent clauses summarize the requirements compiled for DVB-C2 by the DVB Commercial Module. The requirements are differentiated in 4 categories:

- 1) General requirements;
- 2) Performance and efficiency requirements;
- 3) Backwards compatibility requirements; and
- 4) Interactive systems requirements.

#### 4.1.1 General requirements

- The technologies shall aim at optimizing the use of cable channels in state of the art cable networks. This includes enhanced flexibility and robustness, as well as maximum payload data capacity.
- DVB-C2 should not primarily aim at matching DVB-S2 and/or DVB-T2, but fully exploit its differentiating features to compete in the market of content delivery. Therefore downstream transmission technologies that maximally benefit from the availability of the return channel should be evaluated. However, the specification of DVB-C2 shall not depend on the availability of a return channel.
- A toolkit of system parameters shall be available to address applications across consumer to business applications, taking into account different performance level of the CATV network.
- The specification shall allow service providers on cable networks to have individual quality of service targets, even for services within the same multiplex.
- Suitable techniques already in existence shall be adopted wherever possible.
- Due account shall be taken of anticipated cable network characteristics (e.g. with fibre to the curb, building and home, as far as applicable).
- New technical specifications shall address transmit-end functions only, but shall take account of cost implications for different devices, such as receivers or head-end equipment.
- The DVB-C standard shall not be modified, nor shall it require changes to other specifications (e.g. SI) or cause any existing feature to become invalid.
- The specifications shall be transmission frequency neutral within typical cable frequency bands.
- DVB family approach: DVB-C2 should reuse existing solutions for interfacing, coding and modulation wherever appropriate.

#### 4.1.2 Performance and efficiency requirements

- DVB-C2 should be able to efficiently support the migration from a mixed analogue/digital to full digital network and be able to offer the maximum performance/throughput in both networks.

- DVB-C2 shall give at least 30 % more throughput in existing cable plants and in-house networks compared to 256-QAM (DVB-C).
- DVB-C2 shall allow achieving the maximum benefit from statistical multiplex method. E.g. the current fixed channel raster could be deregulated.
- Cable networks should be characterized and modelled on a global (e.g. US, Asia and Europe) level (including in-house network) and the best modulation/FEC schemes shall be selected taking into account a realistic cable channel model including:
  - Deployment of analogue PAL/SECAM/NTSC TV channels.
  - Deployment of different digital signals (such as DVB, DOCSIS<sup>®</sup>, DVC) and the associated signal back off ratios to analogue signals.
  - Different noise (white, burst, impulse), non-linearities and other interferences present in current and future networks.
- The error performance of the system shall be suitable for all types of services that may be carried.
- The DVB-C2 transmission system should be able to support low power modes to maximally reduce power consumption in receivers according to the EU Code of Conduct on Energy Consumption.
- Seamless retransmission (e.g. from DVB-S2 to DVB-C2, or DVB-T2 to DVB-C2) should be fully supported.
- The DVB-C2 standard [i.1] shall provide a fully transparent link for **Transport Stream**, IP-packets and other relevant protocols between the input of the modulator and the output of the demodulator.
- The Zapping time (time to tune a receiver from one service to another) shall not be significantly increased due to the introduction of DVB-C2 (in relation to today's user experience of digital TV services with DVB-C). For any change in RF channel, the DVB-C2 front-end shall deliver a quasi error free signal within 300 ms.

### 4.1.3 Backward compatibility requirements

- DVB-C2 shall not be backwards-compatible with DVB-C (in a sense that a DVB-C receiver is able to process a DVB-C2 signal). The capability for a DVB-C2 receiver to include DVB-C functionalities should be addressed as an optional requirement in the technical specification, so that:
  - if this is a requirement from the industry players to include DVB-C functionality into DVB-C2 equipment, chipset manufacturers can provide compliant solutions;
  - if in the long term networks will have migrated completely to DVB-C2, these chipsets may be produced as well.
- For DVB-C2 transmissions, there shall be no requirement for any change to existing DVB-C receivers. This assumes continued use of the same cable network architecture and the same cable channel characteristics.
- In order to allow for self install, the DVB-C2 standard [i.1] should be as insensitive as possible to typical characteristics of in-house networks using coaxial cable systems.

### 4.1.4 Interactive systems requirements

- The specification shall be available for consideration as an alternative downstream coding and modulation scheme for the DOCSIS<sup>®</sup> systems currently using DVB-C for the EuroDOCSIS<sup>®</sup> System.
- DVB-C2 shall include techniques for improving the efficiency of carriage of IP data.
- DVB-C2 shall allow cost effective integration of DVB-C2 into Edge QAM solutions for modulation equipment.
- The specification shall provide a low latency mode for those interactive services that require such a mode.

## 4.2 Key features of DVB-C2

The excellent physical channel characteristics of Hybrid Fibre Coaxial (HFC) networks provide an ideal platform for all current and future broadband communication services. In the past, the available spectrum has been mainly used for the transmission of analogue TV signals. For some years, the application of digital transmission standards as DVB-C in addition with the MPEG-2 video compression has offered the means to provide a huge variety of digital TV programs. However, in many cable networks the analogue programs still have not been switched off, resulting in a parallel transmission of analogue and digital broadcast signals. Over the past years, many cable networks have been upgraded to allow bidirectional communications. As a result, cable operators are nowadays able to offer bandwidth demanding "Triple Play" packages, providing telephony and Internet access in addition to the classical TV broadcast services. Especially the IP traffic per customer is expected to increase rapidly in the coming years. Furthermore, an increasing number of services are offered in High Definition (HD) quality, requiring much higher bit-rates in comparison to Standard Definition (SD) resolution services. On the other hand, the usable downlink spectrum in typical cable networks is often limited to frequencies below 800 MHz. Therefore, many cable operators are running out of spectrum in the near future, as a study of the EU funded "ReDeSign" project (visit: [www.ict-redesign.eu](http://www.ict-redesign.eu)) indicates. Solving this problem is either possible by an extension of the used spectrum or by the reduction of the number of subscribers per network segment. Both approaches are very costly to the cable network operator, as they require many new active and passive network components. The third and most promising possibility to face the increasing demand of bandwidth is the application of a transmission system with a more spectrum efficient physical layer. The latter is the target of the DVB-C2 standard [i.1].

Although the MPEG **Transport Stream** (TS) is still the most favourite protocol used in digital broadcasting, DVB-C2 supports TS, any packetized and continuous input formats as well as the so called Generic Stream Encapsulation (GSE). All input streams are multiplexed into a Baseband Frame format. The Forward Error Correction (FEC) scheme is applied to these Baseband Frames. In line with the other DVB-X2 systems, DVB-C2 uses a combination of LDPC and BCH codes, which is a very powerful FEC providing about 5 dB improvement of signal-to-noise ratio (SNR) with reference to DVB-C. Appropriate Bit-Interleaving schemes optimize the overall robustness of the FEC system. Extended by a header, those frames are called Physical Layer Pipes (**PLP**). One or several of such **PLPs** are multiplexed into a **Data Slice**. A two-dimensional interleaving (in time and frequency domain) is applied to each slice enabling the receiver to eliminate impacts of burst impairments and frequency selective interference such as single frequency ingress. One or several **Data Slice** compose the payload of a C2-frame. The Frame Building process includes inter alia the insertion of Continual and Scattered Pilots. The first symbol of a **C2 frame**, the so-called "Preamble", carries the signalling data. A DVB-C2 receiver will find all relevant configuration data about the structure and the technical parameters of the DVB-C2 signal in the signalling data block in the Preamble as well as in the headers of the **PLPs**. In the following step the OFDM symbols are generated by means of an Inverse Fast Fourier Transformation (IFFT). A 4K-IFFT algorithm is applied generating a total of 4 096 sub-carriers, 3 409 of which are actively used for the transmission of data and pilots within a frequency band of 8 MHz. The guard interval used between the OFDM symbols has a relative length of either 1/128 or 1/64 in reference to the symbol length (448  $\mu$ s).

In a nutshell the key technical features of DVB-C2 are the combination of flexibility and efficiency. It is expected that the deployment of DVB-C2 on one hand will increase the downstream capacity of cable networks by 30 % and for optimized networks up to 60 %. On the other hand DVB-C2 will allow network operators to utilize the available frequency resources in a more flexible way and allow the introduction of both enhanced services and bigger pipes, for all kind of service containers, including the support of IP based transport mechanisms.

## 4.3 Benefits of DVB-C2 compared to DVB-C

Table 1 lists the major differences between DVB-C and DVB-C2 in terms of the relevant used technologies. As a result of the technologies introduced in DVB-C2, the potential gain in capacity that could be achieved is about 33 % compared to the DVB-C mode with the same SNR requirements (see table 2). In addition to the increased efficiency due to the chosen coding and modulation schemes, DVB-C2 allowed to further optimize the spectrum efficiency by increasing the bandwidth of the transmitted signal beyond 8 MHz as shown in table 3. It should be noted that DVB-C2 signals with broader bandwidth than 8 MHz generally can be processed by receivers with fixed 8 MHz receiving windows.



Table 1: Comparison between the basic building blocks of DVB-C and DVB-C2

	DVB-C	DVB-C2
Input Interface	Single Transport Stream (TS)	Multiple Transport Stream and Generic Stream Encapsulation (GSE)
Modes	Constant Coding & Modulation	Variable Coding & Modulation and Adaptive Coding & Modulation
FEC	Reed Solomon (RS)	LDPC + BCH
Interleaving	Bit-Interleaving	Bit- Time- and Frequency-Interleaving
Modulation	Single Carrier QAM	COFDM
Pilots	Not Applicable	Scattered and Continual Pilots
Guard Interval	Not Applicable	1/64 or 1/128
Modulation Schemes	16- to 256-QAM	16- to 4096-QAM

Table 2: Potential capacity or robustness increase of DVB-C2 in relation to 256-QAM DVB-C

	DVB-C	DVB-C2	DVB-C2
64-QAM	256-QAM	1 024-QAM	256-QAM
Guard Interval	NA	1/128	1/128
FEC	RS	9/10 LDPC + BCH	5/6 LDPC + BCH
Symbol rate	6,875 MBaud	NA	NA
Bandwidth	8 MHz	8 MHz	8 MHz
SNR	29,7 dB	29,5 Db	22 dB (see note 2)
Capacity	50,87 Mbit/s	66,14 Mbit/s (see note 1)	49,01 Mbit/s
NOTE 1: The relative capacity gain is 30,5 % at a SNR of 29,7 dB (required for 256-QAM DVB-C).			
NOTE 2: The relative robustness gain is 25,9 % at 50 Mbit/s payload (as provided by 256-QAM DVB-C).			

Table 3: Further improvements of spectrum efficiency due to increased transmitter bandwidth

	DVB-C (256-QAM)	DVB-C2 (4 096-QAM)
8 MHz channel	50,9 Mbit/s	79,54 Mbit/s
16 MHz channel	50,9 Mbit/s	81,58 Mbit/s, 2,56 % (see note)
24 MHz channel	50,9 Mbit/s	82,26 Mbit/s, 3,42 % (see note)
32 MHz channel	50,9 Mbit/s	82,60 Mbit/s, 3,85 % (see note)
64 MHz channel	50,9 Mbit/s	83,11 Mbit/s, 4,45 % (see note)
NOTE: Relative gain with reference to 4 096-QAM in 8 MHz bandwidth.		

#### 4.4 General remark on the applicability of DVB-C2 for cable systems using 8 MHz or 6 MHz basic channel raster

The features of DVB-C2 are applicable for cable network of a "European type" (using an 8 MHz channel raster) and of a "US-type" (using a 6 MHz channel raster). This is possible due to the definition of two different figures for carrier spacing of the OFDM system.

Whenever in the present document a DVB-C2 feature or characteristic is discussed in relation to the European type (8 MHz) scenario with 2,232 kHz carrier spacing, or 7,61 MHz preamble bandwidth, this feature is also available for the "US-type" (6 MHz) scenario using 1,674 kHz carrier spacing and 5,71 MHz preamble bandwidth respectively.

The reason for this easy conversion of features is the fact that in both scenarios the number of OFDM carriers per frequency raster block is the same: 3,408 carriers.

## 5 Anatomy of the DVB-C2 signal

### 5.0 Introduction

DVB-C2 employs state-of-the-art algorithms for higher spectral efficiency and flexibility that allow for optimum usage of the cable network resources. The technical details to reach this goal and the resulting advantages are given in the following clauses of the present document.

### 5.1 System Overview

In accordance with the widely deployed first generation DVB physical layer specifications, the second generation of the DVB physical layer standards family (DVB-X2) consists of three transmission systems for the three different transmission media: DVB-S2 (satellite), DVB-T2 (terrestrial) and DVB-C2 (cable). One main idea of DVB-X2 is the maximum possible commonality and reasonable re-usage of building blocks between the different transmission systems, which is also called the 'Family of Standards' approach. As an example, all standards use a common forward error correction (FEC) scheme consisting of a concatenation of an outer Bose Chaudhuri Hocquenghem (BCH) code and an inner Low Density Parity Check (LDPC) code. Furthermore, the system layers of DVB-S2 and DVB-C2 are very similar, which allows a simple conversion of satellite signals into cable networks - an important feature of a cable broadcasting system. From a first glance, it may be surprising that DVB-C2 applies Orthogonal Frequency Division Multiplex (OFDM) as modulation scheme since it is a novelty in cable downstream transmission. However, the benefits of this approach are explained in more detail in the following clauses.

Figure 1 shows a high-level block diagram of signal processing defined for the DVB-C2 transmitting end. In accordance to DVB-T2, the cable system also adopts the Physical Layer Pipe (**PLP**) approach. A **PLP** is a logical channel that may contain one or more regular MPEG-2 **Transport Streams** but also IP data being processed by the so-called Generic Stream Encapsulation (GSE) protocol. Each **PLP** consists of an input processing block followed by a Forward Error Correction (FEC) and a QAM Mapping stage. One or multiple **PLPs** can be combined to so-called **Data Slice** (similar to channels) which are interleaved in time and frequency to mitigate the influence of burst errors or narrow-band interferers. Finally, the frame builder combines the different **Data Slice**, prepends a preamble with the Level 1 signalling information, and forwards the resulting **C2 frame** to the OFDM generation stage.

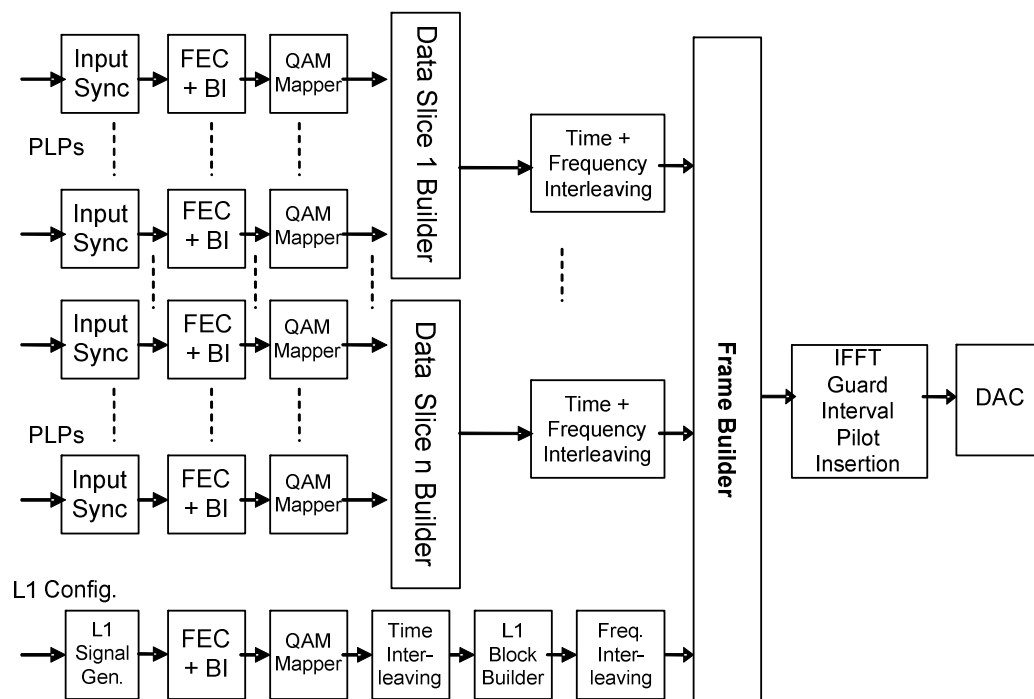


Figure 1: High-level block diagram of the signal processing defined for the DVB-C2 transmitting end

## 5.2 Concept of Physical Layer Pipes (PLP) and Data Slices

The Physical Layer Pipe (**PLP**) concept allows the transmission of several independent logical channels. Each **PLP** constitutes such a logical channel containing data based for instance on the MPEG-2 **Transport Stream** or on the Internet protocol by using the Generic Stream Encapsulation (GSE). The **PLP Identifier (PLP\_ID)**, which allows a unique identification of dedicated **PLPs** at the receiver side, is part of a so-called FECFRAME header being placed in front of each user packet. After decoding this header and evaluating the **PLP\_ID**, the receiver is able to decide whether it has to decode the following data packet. Data packets which do not belong to the desired **PLP** are ignored and do not have to be processed neither by the QAM demapper nor by the forward error correction decoder. Consequently, the effective receiver bit rate as well as the related processing power decrease significantly.

Another advantage of the **PLP** approach is the possibility to assign different robustness levels to different streams: Each **PLP** can adjust its modulation and FEC rate independently from the others. As a result, different 'Quality of Service' levels can be assigned to differentiate services at the physical layer already. While the robustness settings for broadcast services have to be adjusted to guarantee a high service quality at all user outlets of a cable network, the advantage for the provision of interactive data in point-to-point connections is even more evident: Depending on physical cable network characteristics like distance, number of amplifiers, and the quality of in-house installation, the achievable signal quality can vary significantly. If the cable headend could get knowledge about the signal quality and the associated characteristics of the cable channel carrying the signal to a dedicated DVB-C2 user terminal, it would be able to adjust the robustness settings for each individual case thus optimizing the data throughput for each user. An example of a typical application of this technique certainly is High-speed Internet access via cable. DVB-C2 could be used as downlink medium, whereas information about the current downlink quality could be sent to the headend by means of cable modems employing return channels. This application describes how the headend would be enabled to configure the downstream signals to each individual use providing the maximum spectral efficiency offered by the network.

## 5.3 Forward Error Protection and Modulation Constellations

The performance of the forward error correction builds the basis of a powerful transmission system. As part of the DVB-X2 family approach, DVB-C2 employs exactly the same Low Density Parity Check (LDPC) codes that have been already used for DVB-T2 and DVB-S2. This code class has already been known since the sixties, but its utilization in real implementations only became possible in recent years due to the progress in semiconductor manufacturing. The impressive benefits of LDPC can be shown by the following numbers: The 9/10 code rate of DVB-C2 is able to correct bit streams with bit-error rates of several percent measured at the input of the FEC decoder. In contrast, the Reed Solomon code applied for DVB-C, which has a similar effective code rate, only tolerates a maximum bit-error rate of  $2 \cdot 10^{-4}$  to reach the goal of quasi error-free reception corresponding to one erroneous event per hour. This high performance of the LDPC codes is the most significant for large LDPC codeword lengths. In fact DVB-C2 uses a codeword length of 64 800 bits, for instance, which has a multiple length compared with the 1 632 bits and 204 bytes, respectively, in DVB-C. Consequently, DVB-C2 is no longer restricted to the transmission of MPEG-2 **Transport Stream** packets but support also other packetized data streams. Besides the LDPC code, a BCH code is employed by DVB-C2 after LDPC decoding. It adds redundancy to the bit stream which consumes less than 1 percent of the total bit rate. This code with very little error correction capabilities is used to correct the error-floor which typically arises at the output of the LDPC decoder. This error-floor, which occurs in most iterative coding schemes as LDPC or Turbo codes, leads to few remaining bit errors after the decoding process, which cannot be corrected by further iterations of the FEC decoder.

The significantly increased performance of the forward error correction allows for the application of higher constellation schemes being used for modulation of the OFDM subcarriers. While DVB-C maximally employed 256-QAM, DVB-C2 now adds 1 024-QAM and 4 096-QAM. The possible combinations of modulation and coding schemes are given in table 4. The table also indicates the required signal-to-noise-ratios for quasi error-free reception. The range of signal-to-noise ratios varies from approximately 10 dB to 35 dB, while the available modulation and coding schemes allow for a granularity of approximately 2 dB.

**Table 4: Available QAM mappings and code rates (CR) for DVB-C2 and their required signal-to-noise-ratio (SNR) for quasi-error-free reception, "-" means not applicable**

CR	16-QAM	64-QAM	256-QAM	1 024-QAM	4 096-QAM
2/3	-	13,5 dB	-	-	-
3/4	-	-	20,0 dB	24,8 dB	-
4/5	10,7 dB	16,1 dB	-	-	31,4 dB
5/6	-	-	22,0 dB	27,2 dB	32,4 dB
9/10	12,8 dB	18,5 dB	24,0 dB	29,5 dB	35,0 dB

## 5.4 DVB-C2 Framing and OFDM Generation

### 5.4.0 Introduction to DVB-C2 Framing and OFDM Generation

As a major difference to DVB-C, DVB-C2 uses OFDM instead of single-carrier QAM modulation. OFDM is applied in most state-of-the-art broadcast and bidirectional transmission schemes due to its well-known and proven robustness to different types of channel impairments (e.g. multipath effects or narrowband interferers). DVB already deployed OFDM in the first generation terrestrial DVB-T system. For the second generation terrestrial transmission system DVB-T2 the parameters were substantially refined and extended. DVB-C2 reuses a parameter set of DVB-T2 that is well-suited for the cable specific requirements. As a result of using the same OFDM parameters and the large number of other common blocks such as the error correction, it is assumed that combined DVB-T2 and DVB-C2 chips can be realized without severe overhead.

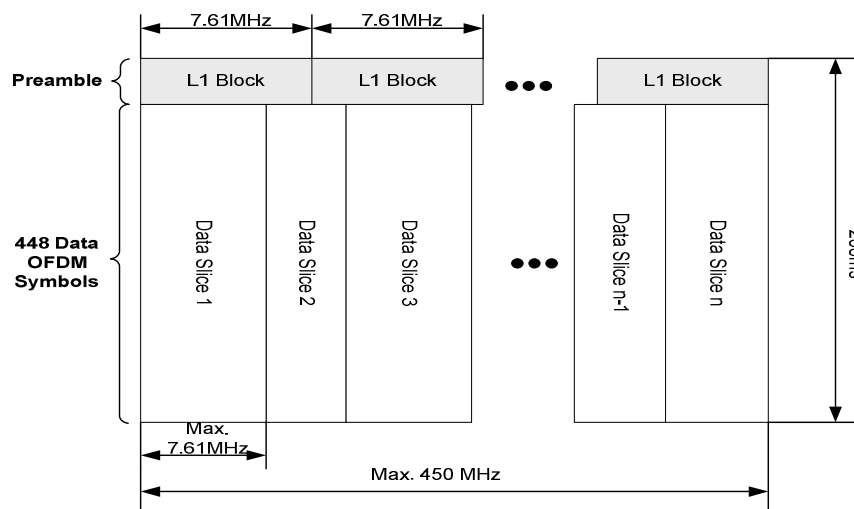
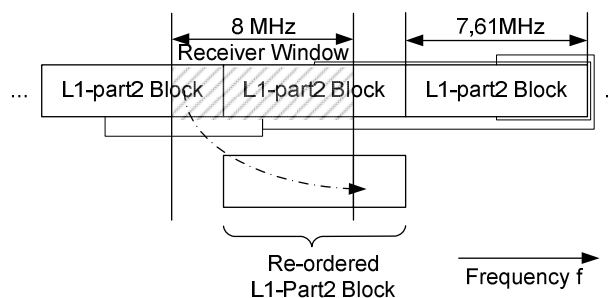


Figure 2: Time frequency diagram of a DVB-C2 frame

DVB-C2 reuses the OFDM subcarrier spacing of the 4K FFT mode from DVB-T2, i.e. the useful **OFDM symbol** duration is 448  $\mu$ s. Two options for the Guard Interval lengths are available, namely a fraction of 1/64 and 1/128 of the total symbol duration. Furthermore, DVB-C2 uses the same scattered pilot patterns which allow the implementation of the same channel estimation block for both systems.

One key feature of DVB-C2 is the capability of generating signals with a variable bandwidth. This is achieved by allocating a specific number of OFDM sub-carriers while keeping the various filter parameters and the system clock unchanged. As a result, the signal bandwidth at the transmitting end can be extended to higher figures for inclusion of a larger number of services. To avoid complex and expensive consumer electronic receivers, segmented OFDM reception - as also used in the Japanese terrestrial broadcasting standard ISDB-T for instance - is applied. The receiver with its traditional 6 MHz or 8 MHz TV tuner bandwidth can extract that part of the broad transmission signal which contains the desired service. This part is constituted by a **Data Slice** which never exceeds the traditional bandwidth of a receiver tuner. The time frequency diagram of the C2 framing enabling such flexible reception behaviour is depicted in figure 2.

Each **C2 frame** starts with a preamble consisting of one or more OFDM symbols. The preamble has two main functions. On the one hand, it allows for reliable time and frequency synchronization to the OFDM signal and the framing structure itself. Therefore, a unique preamble pilot sequence is modulated onto every 6th OFDM sub-carrier of the **preamble symbols**. On the other hand, the preamble carries the Layer 1 signalling required for the decoding of the **Data Slice** and their payloads. The preamble consists of a frequency cyclic repetition of the **L1 part 2** blocks that are repeated every 7,61 MHz. The reason for the fixed allocation of the **L1 part 2** blocks and their repetition is the requirement to access the complete **L1 part 2** signalling in any tuning position of an 8 MHz receiver tuner. As depicted in figure 3, the receiver is able to restore the complete data of a **L1 part 2** block by re-ordering the OFDM carriers after converting them into the frequency domain (i.e. by the FFT at the receiving end). Even the loss of some carriers would not seriously affect the system performance, as the signalling data is transmitted in a very robust mode.



**Figure 3: DVB-C2 preamble structure, re-ordering of the OFDM sub-carriers in the L1-part2 blocks to extract the L1 signalling information**

In contrast to the **L1-part2** blocks, the **Data Slice** do not have to comply with any constant frequency pattern, thus can be allocated in a flexible way. This is the reason for the requirement to access and decode the **L1-part2** signalling data at any possible tuning position. The only requirement for the **Data Slice** is that each **Data Slice** shall not exceed the maximum reception bandwidth of 7,61 MHz. As a result, the bandwidth of a **Data Slice** can be adjusted very precisely to the bit-rate of the source signal. For example, satellite streams with very different bit rates can be inserted into the C2 signal without the need of exhaustive stuffing overhead or re-multiplexing of the MPEG-2 **Transport Streams**. Different **Data Slice** can be accumulated until the overall number of OFDM sub-carriers of the C2 signal is reached. Both position and bandwidth of the **Data Slice** may vary between different **C2 frames** as this does not require any re-tuning of the receiver. The signalling inside the **L1-part2** blocks does not only contain the start and the end frequency of the **Data Slice**, but also the optimal tuning position. Thus, the transmitter may vary the **Data Slice** parameters inside the transmitter defined receiving window.

## 5.4.1 DVB-C2 signalling concept

### 5.4.1.0 Elements of the DVB-C2 signalling

The DVB-C2 signalling scheme consists of two components, the Layer 1 and the Layer 2 signalling.

#### 5.4.1.1 L1 signalling scheme

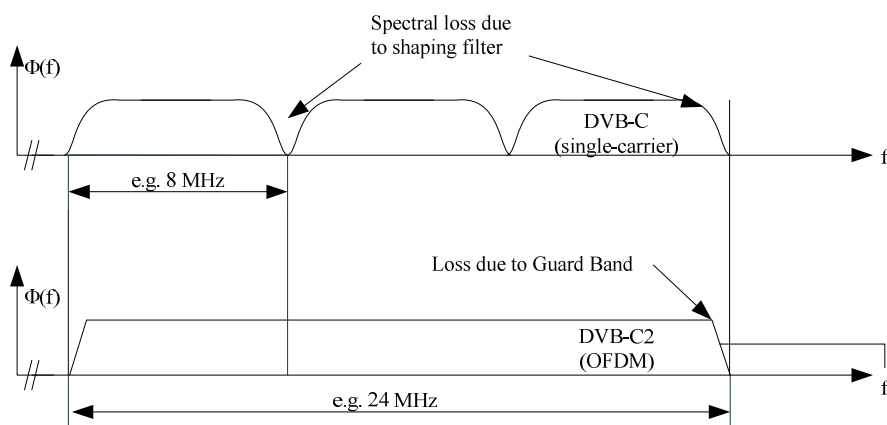
Layer 1 signalling is spread into two parts. The first part (**L1-part1**) is transmitted in the **FECFrame** headers (carrying) modulation scheme and code rate of the individual **PLPs** and the second part of Layer 1 signalling (**L1-part2**) is carried in the preamble payload. There is also a Layer 2 signalling scheme defined in DVB-C2. The so-called DVB-C2 delivery system descriptor is integrated into the DVB-SI system and transmitted in the so called Network Information Table (NIT). More details about L1 signalling and the choice of relevant parameters are given in clauses 6.8 and 8.4 respectively.

#### 5.4.1.2 L2 signalling scheme

Layer 2 signalling delivers the mapping of **PLPs** to the relevant services to the receiver. More details are given in clause 7.8

## 5.5 Spectral Efficiency and Transmission Capacity

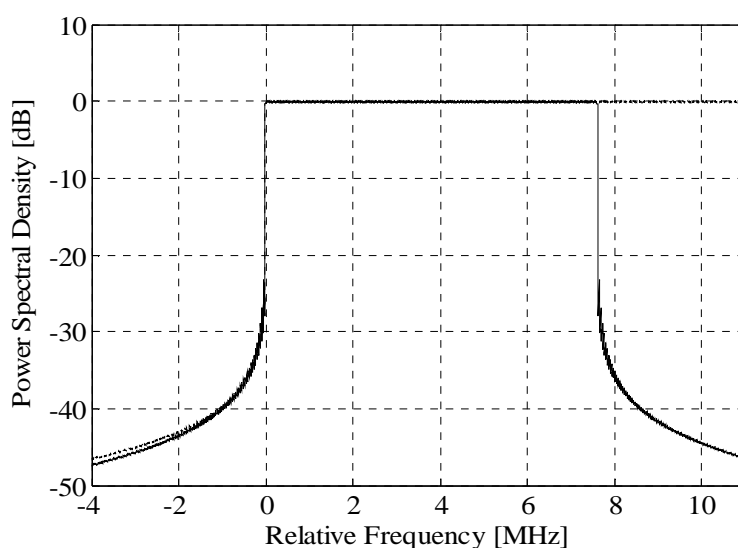
One main goal of the DVB-C2 specification is an increased spectral efficiency. DVB-C2 achieves this goal through the utilization of the LDPC codes in combination with higher QAM mappings and the application of OFDM. The gain caused by OFDM is depicted in figure 4.



**Figure 4: Occupation of frequency spectrum by DVB-C today (upper diagram) and by DVB-C2 in the future (lower diagram)**

Today, DVB-C is based on single-carrier modulation which uses a shaping-filter to form the transmitted signal. The resulting curve progression at the frequency edges of each signal spectrum shows the roll-off effect caused by the Half Nyquist filter used in the transmitter. It has a roll-off factor of 0,15 (corresponding 15 %) for DVB-C and reduces the spectral efficiency of the signal by the same number. Smaller values (e.g. 10 %) are feasible, but require higher accuracy for a number of elements in the transmitter and particularly in the receiver. Furthermore, this factor is independent of the channel bandwidth, i.e. the relative spectral loss of 15 % is the same for 16 MHz channels. In contrast, this is not the case for OFDM as applied in DVB-C2.

The spectral loss for OFDM is caused by the Guard Interval, the frequency domain pilots and the Guard Bands at the edges of the spectrum. In the normal mode with a Guard Interval length of 1/128 and a frequency domain pilot density of 1/96, the loss caused by these two components is equal to approximately 2 %. Furthermore, a Guard Band at the frequency edges of the spectrum is required to avoid disturbance to neighbouring channels. There is no need of any frequency separation between the **Data Slice** of one OFDM signal, i.e. **Data Slice** can be seamlessly concatenated in the frequency direction. The width of the Guard Band is nearly independent of the actual bandwidth of the OFDM signal, as figure 5 shows. The power spectral density of an ideal DVB-C2 signal for 7,61 MHz and 450 MHz bandwidth nearly overlaps at the frequency edge of the signal (left hand side). Thus, for 7,61 MHz wide signals as well as for 450 MHz wide signals a Guard Band of roughly 200 kHz will be sufficient. Consequently, the spectral loss is significantly reduced for wider OFDM signals. The overall spectral loss for a 32 MHz DVB-C2 signal (e.g. 5 **Data Slice** of 6,4 MHz) is only 3,25 %, while it is 15 % for DVB-C.



**Figure 5: Ideal power spectral density of DVB-C2 signal with Guard Interval 1/128 for 7,61 MHz and 450 MHz**

The combination of the reduced modulation overhead and the increased robustness of the LDPC codes provides a system performance that closely reaches the theoretical limit of spectral efficiency, also referred to as Shannon Limit. Figure 6 shows the Shannon Limit in comparison with the performance figures reached by DVB-C2. Also, the spectral efficiency of a 256-QAM DVB-C signal is shown at the related signal-to-noise ratio that is necessary to receive the signal without any erroneous bits. It is apparent that the distance in terms of signal-to-noise ratio between the value achieved by DVB-C and the Shannon Limit is 10 dB, whereas the respective distance between the values produced by DVB-C2 and the Shannon limit is only 2 dB to 3 dB. This difference in signal-to-noise ratio required by the two systems to reach a certain spectral efficiency shows the substantial increase which has been achieved through the development of DVB-C2.

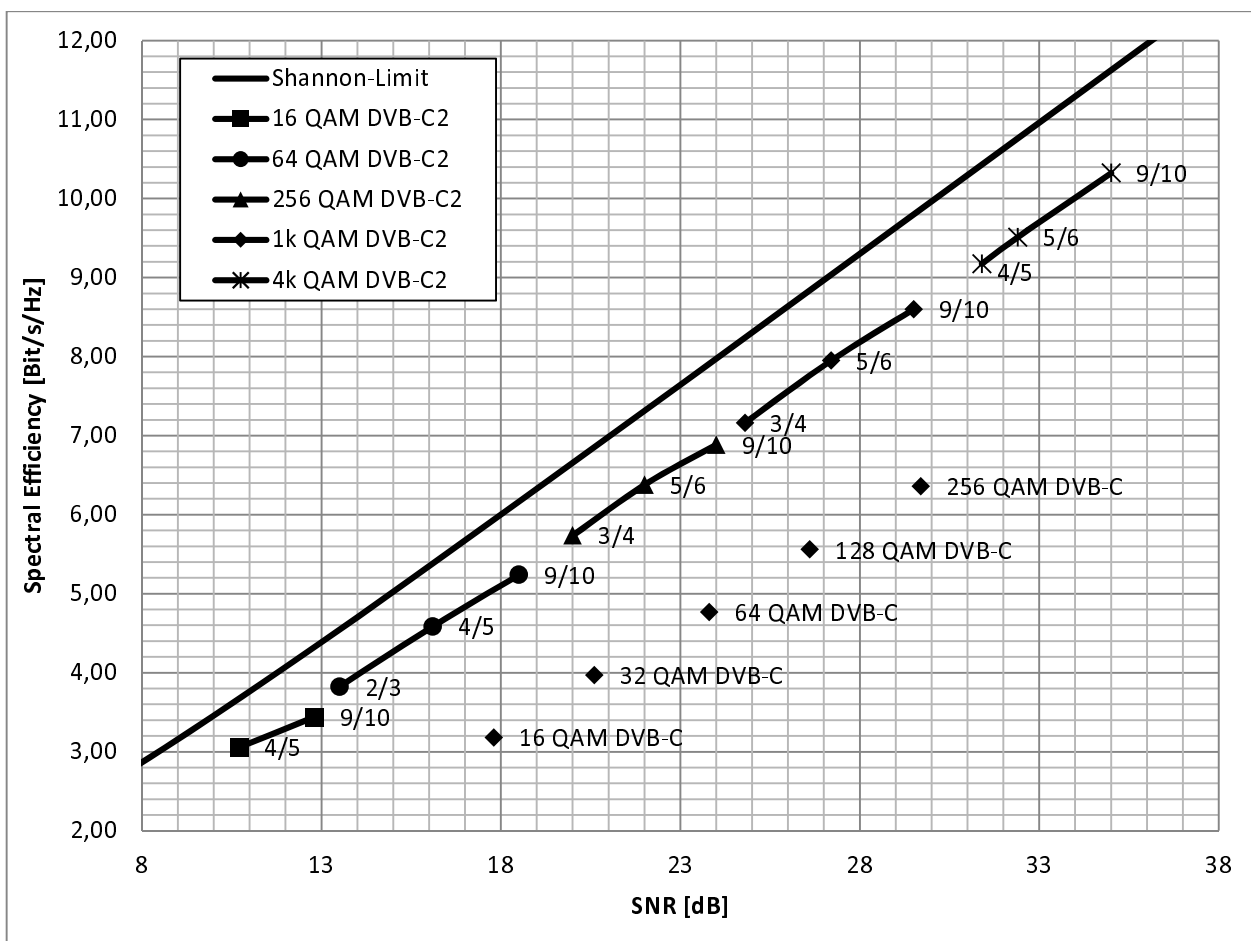


Figure 6: Overall spectral efficiency of DVB-C and DVB-C2 for different code rates and modulation schemes (DVB-C2 parameters: 32 MHz signal bandwidth, Guard Interval 1/128, pilot density 1/96)

Table 5: List of available payload bit-rates for DVB-C (partly) and DVB-C2 per 8 MHz channel (DVB-C2 parameters: 32 MHz signal bandwidth, Guard Interval 1/128, pilot density 1/96)

	16-QAM	64-QAM	256-QAM	1 024-QAM	4 096-QAM
DVB-C	25,6 Mbit/s	38,4 Mbit/s	51,2 Mbit/s	-	-
C2,2/3	-	30,6 Mbit/s	-	-	-
C2,3/4	-	-	45,91 Mbit/s	57,3 Mbit/s	-
C2,4/5	24,5 Mbit/s	36,7 Mbit/s	-	-	73,4 Mbit/s-
C2,5/6	-	-	51,04 Mbit/s	63,6 Mbit/s	76,1 Mbit/s
C2,9/10	27,3 Mbit/s	41,3 Mbit/s	55,1 Mbit/s	68,8 Mbit/s	82,6 Mbit/s

NOTE: "-" means not applicable.

The available payload bit-rates for DVB-C and DVB-C2 are summarized in table 5. For the comparison of both systems, the figures shown refer to a transmission within an 8 MHz channel. In case of DVB-C2, an overall signal bandwidth of 32 MHz has been assumed. The modes of DVB-C2 allow an increase in payload bit-rate of up to 65 %, while the required signal-to-noise-ratio of 35 dB is achievable in most state-of-the-art cable networks.

## 5.6 DVB-C2 multiplexing schemes

### 5.6.0 Introduction

The DVB-C2 multiplexing schemes consist of two parts. The so called "physical frame structures" are the schemes used to multiplex Physical Layer Pipes into a **C2 frame** and the so called logical frame structure are the schemes to convert input signals into Physical Layer Pipes (**PLPs**).

### 5.6.1 Physical frame structure

#### 5.6.1.1 C2-System

The DVB-C2 specification [i.1] uses the phrase "C2 System" in different senses. In some places it is used in a general way to refer to the DVB-C2 standard [i.1] and the various elements which make it up.

However, "C2-System" is also used as the name of a transmitted DVB-C2 signal. In this specific sense a certain multiplexing configuration of **PLPs** and **Data Slice** form together with the preamble **C2 frames** to form a C2-System transmitted via a cable channel.

#### 5.6.1.2 C2-frame

The **C2 frame** is the highest level of the DVB-C2 multiplexing schemes. A Frame starts always with a preamble (one or several symbols duration), followed by 448 **data symbols**. More details about the C2-frame and the choice of relevant parameters are given in clauses 6.5 and 8.6.

#### 5.6.1.3 Data Slices

A **Data Slice** contains one or more **PLPs** and is transmitted with identical time and frequency interleaving parameters. More details about **Data Slice** and the choice of relevant parameters are given in clause 6.9.

#### 5.6.1.4 Physical Layer Pipes

A Physical Layer Pipe (**PLP**) is a data container, carrying data with a certain set of modulation parameters (QAM-modulation scheme and FEC code rate). More details about **PLPs** and the choice of relevant parameters are given in clause 7.

#### 5.6.1.5 FECFrames

##### 5.6.1.5.0 Introduction of FECFrames

The baseband frame, completed with a header, is treated as an information word to which BCH and LDPC coding are applied. The resulting codeword always contains either 64 800 bits or 16 200 bits and is known as a **FECFRAME**. The two different lengths correspond to the choice of long or short FEC blocks respectively. Short FEC blocks allow a finer granularity of bit-rate but incur a greater overhead and slightly worse performance than long blocks.

##### 5.6.1.5.1 FECFrame Headers

**FECFrame** headers include the L1 part 1 signalling information. They are applied for **Data Slice** of Type 2. The **FECFrame** header is prepended at each QAM modulated **FECFrame** and includes information on **PLP\_ID** as well as used QAM modulation scheme and LDPC code rate of the following **FECFrame** packet. Furthermore the **FECFrame** header allows for reliable synchronization to the **XFECFrames**. The structure and functionality of the **FECFrame** header is defined in clause 7.2.2 of ETSI EN 302 769 [i.1].

#### 5.6.1.6 BBFrames

**BBFrames** are the basic unit in the logical framing structure of DVB-C2: allocation and scheduling are performed in whole numbers of **BBFrames**. Where packetized streams are being carried, the packets may be mapped to **BB-Frames** either synchronously or asynchronously, i.e. each **BBFrame** may contain a whole number of packets, or packets may be fragmented across two **BB-Frames**. **BBFrames** contain a header, including the packet length and position of the first packet, which allows the original packets to be reconstructed at a receiver. **BBFrames** may contain padding, in case insufficient data is available for a whole **BBframe**, or in case it is desired not to avoid packet fragmentation. The total size of each **BBFrame**, including any padding is constant for a given **PLP** and depends on the LDPC code rate and whether short or long FEC blocks are used.



### 5.6.1.7 Packets

**PLPs** may carry packets, including MPEG **Transport Stream** packets or other types of packets used by generic streams. They may also carry continuous, non-packetized streams. The **BBFrame** header (see clause 5.1.6 in ETSI EN 302 769 [i.1]) provides a mechanism for packets to be reconstructed at the receiver, but the framing of the DVB-C2 signal itself is generally independent of any packet structure which the input streams may have.

## 5.7 Overview of interleaving

There are three different interleavers in the DVB-C2 specification, occurring in the following order in the modulation chain:

- Bit Interleaver.
- Time Interleaver.
- Frequency Interleaver.

The Bit Interleaver interleaves the code-bits within an LDPC codeword in order to avoid undesirable interactions between the bits carried by the same cell and the structure present in the LDPC code.

The Time Interleaver (clause 10.2) provides protection of the signal against impulsive interference as well as time-varying channels. The Frequency Interleaver (clause 10.3) improves the performance in multipath channels as well as in case of narrowband interference.

## 5.8 Payload Capacity

This clause lists the important technical parameters influencing the available payload capacity of a C2\_System.

- Modulation and code rate used for **PLPs**

Obviously the choice of modulation parameters and code rate has the strongest impact on the payload capacity. In annex A examples for the calculation of the payload capacity for different modulation schemes and code rates are given.
- Transmission channel bandwidth
 

European cable networks are currently using an 8 MHz channel raster, in some countries also a 7 MHz raster for analogue services. In case of the 2,232 kHz carrier spacing, defined for European DVB-C2 applications, a DVB-C2 signal may have any bandwidth from 8 MHz up to 450 MHz. It should be noted that a DVB-C2 transmitter signal with  $n \times 8$  MHz bandwidth provides a higher payload than  $n$  DVB-C2 transmitter signals with 8 MHz bandwidth (see also annex A).
- Carrier spacing
 

DVB-C2 has defined two different figures for carrier spacing: 1,674 kHz and 2,232 kHz. The 1,674 kHz spacing is defined for cable networks with basic 6 MHz frequency raster, whereas the 2,232 kHz spacing fits for cable network with 8 MHz frequency raster.
- Guard interval/Pilot pattern
 

DVB-C2 has defined two possible lengths for the guard interval: 1/64 and 1/128. The pilot density is different for those two options and therefore the choice of the guard interval has an impact on the payload capacity of a given configuration.
- Mode adaptation/**PLP/Data Slice** parameters
 

The DVB-C2 system does provide means to further optimize the payload capacity in simple and static configurations. Details are discussed in clauses 6.7, 6.8 and 6.9.

- Usage of reserved tones

In case the Peak-to-Average-Power-Ratio of a DVB-C2 signal needs to be reduced, a set of symbols per C2-frame has been defined (reserved tones) to carry a special modulation for that purpose. Those symbols are no longer available for payload purposes and the payload capacity is reduced accordingly.

- Number of **preamble symbols**

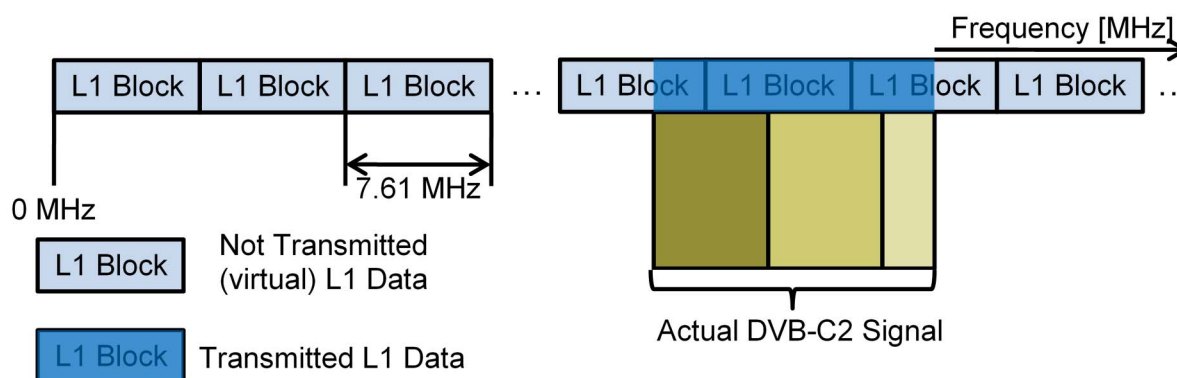
For DVB-C2 signals with complex **PLP** and **Data Slice** configuration the number of Preamble symbols required to carry the **L1-part2** signalling may exceed one symbol and therefore the payload capacity is reduced accordingly.

- **notches**

In case **notches** are required in order to cope with certain interference scenarios, the payload symbols falling in the notch bands are switched off and the payload capacity is reduced accordingly.

Annex A gives examples for payload capacity calculations for different scenarios.

## 5.9 New concept of absolute OFDM



**Figure 7: Absolute OFDM concept in DVB-C2**

The concept of 'absolute OFDM' (see figure 7) is unique to DVB-C2. The first **L1-part2** signalling block begins at the absolute frequency of 0 MHz and the further blocks are partitioned in steps of 7,61 MHz towards higher frequencies. In contrast to other DVB standards it is not possible to shift a C2 baseband signal to any RF carrier frequency rather than being defined in a unique way for the whole cable spectrum: Especially the pilot sequences of the OFDM signal are different for all different frequencies. The reason for that behaviour is to avoid unwanted repetitions in the frequency domain which may cause unwanted high peak values of the OFDM signal in time domain. Furthermore, the unambiguous pilot sequences allow for easy and reliable synchronization and offset compensation.

Although the **L1-part2** block partitioning and the related pilot sequences are defined for the whole cable spectrum - of course L1 blocks are only transmitted in those frequencies where **Data Slice** are present.

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## 6 Choice of Basic Parameters

### 6.0 Introduction

There are a large number of parameters defined in DVB-C2 for configuring a cable transmission system. This clause will discuss the choice of each of the main parameters in turn.

### 6.1 Choice of code rate, block length and constellation

A choice of five rectangular QAM constellations is available in DVB-C2: 16-QAM, 64-QAM, 256-QAM, 1 024-QAM and 4 096-QAM. Selection of the constellations required can be related back to the commercial requirements of a minimum 30 % increased throughput over DVB-C and reuse of all existing cable network architectures and related channel characteristics. These requirements effectively extend the range of channel SNR to be covered by DVB-C2 over DVB-C.

NOTE: During the standardization process non-rectangular QAM constellations were also considered but the performance benefits were not considered to be sufficient to justify inclusion in ETSI EN 302 769 [i.1].

Use of the LDPC codec first used in DVB-S2 and more recently DVB-T2 maximizes error correction capabilities with performance close to the theoretical Shannon limit across the range of channel SNRs required for DVB-C2. The LDPC codec also brings the flexibility of variable coding rates and the option of short/long codes to cable transmission systems while maintaining the 'family of standards' approach desired for DVB-C2/T2/S2. The performance of the LDPC short codes is some tenths of a dB worse than normal codes and will typically be used for low-bit-rate applications requiring shorter latency.

Higher code-rates and higher-order constellations both give greater bit rates but require higher signal-to-noise ratios and improved phase noise performance from the frequency conversion components in a cable transmission system.

Use of 1 024-QAM or 4 096-QAM modulation schemes are required to meet the minimum requirement of a 30 % increased maximum throughput over the 50 Mbit/s achievable using 256-QAM in DVB-C. Use of the LDPC codec enables that same throughput to be achieved at a channel SNR similar to 256-QAM DVB-C when used with a high code rate (9/10) thus bringing improved performance to existing cable networks. 4 096-QAM was added to the modulation options to provide extended throughput (> 50 % increase) over the minimum requirement in high quality cable networks where the available channel SNR at the receiver is 40 dB or greater. The expectation is that over time more networks will use 4 096-QAM as they become upgraded with increased use of fibre. For channels with low SNR 16-QAM can be used with the LPDC codec improving reception performance from the 16 dB SNR from DVB-C down to 10 dB SNR.

Knowledge of the current use of cable networks suggested that a 2 dB granularity in SNR between the limits of 10 dB (16-QAM) and 35 dB (4 096-QAM) would provide sufficient options (table 6) for configuration and optimization of a DVB-C2 -based cable network. The LDPC code rates of 2/3, 3/4, 5/6 and 9/10 combined with the constellations listed above provide the desired granularity (figure 8). The number of options is kept to a minimum to simplify the implementation and testing requirements for related cable equipment. For example, not all code rates are used with all QAM constellations as this can result in some overlap between higher order QAM at a low code rate and a lower order QAM at a high code rate - e.g. 64-QAM at 9/10 code rate has similar spectral efficiency to 256-QAM at 2/3 code rate. In these cases it is always preferable to select the lower order QAM.

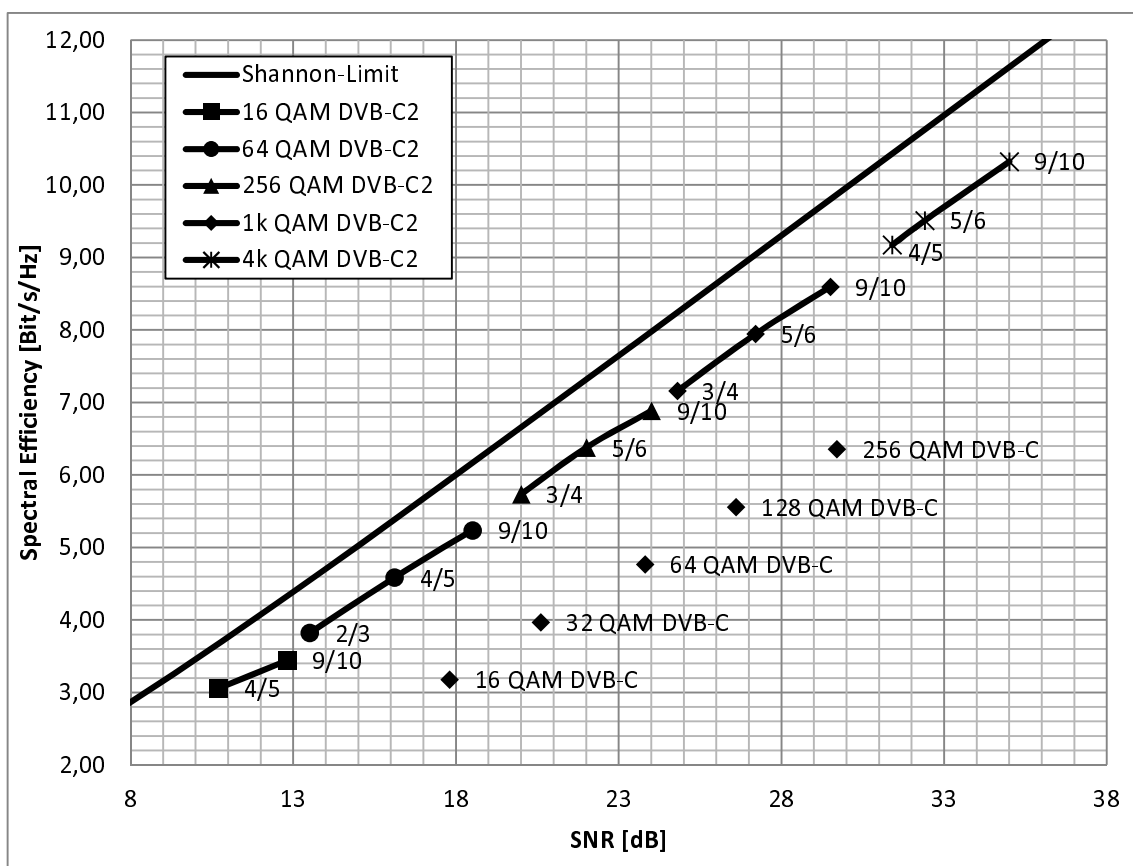


Figure 8: DVB-C2 Modulation Levels, Code Rates and Spectral Efficiency compared with Theoretical Shannon Limit and DVB-C

Table 6: Granularity of Modulation and Code Rates in DVB-C2 ( $N_{ldpc} = 64\ 800$ )

Modulation	FEC Code Rate	Spectrum Efficiency (Bit/s/Hz)	SNR @ bit error rate $10^{-6}$ (dB)	Difference from Upper ModCod (dB)
4 096-QAM	9/10	10,8	34,97	-
4 096-QAM	5/6	10,0	32,36	2,6
4 096-QAM	4/5	9,6	31,40	1,0
1 024-QAM	9/10	9,0	29,50	1,9
1 024-QAM	5/6	8,33	27,15	2,3
1 024-QAM	3/4	7,5	24,81	2,3
256-QAM	9/10	7,2	24,02	0,8
256-QAM	5/6	6,67	21,96	2,0
256-QAM	3/4	6,0	19,97	2,0
64-QAM	9/10	5,4	18,40	1,5
64-QAM	4/5	4,8	16,05	2,4
64-QAM	2/3	4,0	13,47	2,6
16-QAM	9/10	3,6	12,80	0,7
16-QAM	4/5	3,2	10,72	2,1

## 6.2 Choice of FFT size and Carrier Spacing

Table 7 lists some of the tradeoffs to be considered when selecting an appropriate FFT size in OFDM-based transmission systems.

Table 7: Tradeoffs in FFT size

Increasing FFT size - Benefits	Decreasing FFT size - Benefits
Smaller guard interval fraction for required maximum echo delay tolerance	Reduced complexity (although impact on receiver chip is relatively small)
Reuse of DVB-T2 parameters - well understood and helps maintain desired 'family of standards' approach to C2	Improved performance in time-varying channels (not relevant to cable as channel is static)
Improved spectral efficiency as pilot density can be reduced	
Improved spectrum characteristic (side lobe steepness)	
Improved flexibility for notching of subcarriers	

The combination of OFDM modulation used with high order QAM constellations per carrier requires that phase noise effects need to be carefully considered to build a viable system. The effects of phase noise in OFDM are well documented and can be elaborated down to two essential components: Common phase error and Inter-Carrier Interference (ICI). The effects of CPE can be removed in the receiver using e.g. continuous pilots in the transmitted data. On the other hand the ICI cannot be removed which results in an AWGN like SNR degradation. In principle increasing the distance between carriers by using a smaller **FFT size** in an 8 MHz bandwidth would reduce the effects of ICI. During development of the DVB-C2 standard [i.1] a comparison in measured performance using 2 K and 8 K **FFT sizes** (from DVB-T) with different phase noise profiles has been made. No major dependency on OFDM carrier spacing in the frequency domain have been found despite no mechanism existing in the measurement setup to remove the effects of ICI.

In DVB-C2 the OFDM subcarrier spacing is adopted from the 4 k mode used in DVB-T2/H terrestrial and mobile systems. The useful **OFDM symbol** duration is the inverse of the subcarrier spacing and has a value of 448  $\mu$ s. It is independent from the channel bandwidth since the carrier spacing remains constant.

Different channel bandwidths are therefore realized by the related number of OFDM subcarriers. For example, a basic 8 MHz C2 signal would consist of typically 3 409 subcarriers and could be realized by a 4 k IFFT on transmitter side. A broader C2 signal could cover a bandwidth of e.g. 32 MHz, consisting typically of  $3 \times 3\,584 + 3\,409 = 14\,161$  subcarriers. In this case a 16 k IFFT could be used for signal generation on transmitter side.

The two defined figures for carrier spacing 2,232 kHz and 1,674 kHz (see table 38 in [i.23]) are allowing a minimum transmitter and receiver bandwidth of 8 MHz for 2,232 kHz carrier spacing and 6 MHz for 1,674 kHz carrier spacing. Depending on the existing basic frequency raster (corresponding with the receiver bandwidth of the deployed CPEs), the OFDM carrier spacing should be chosen by a cable operator.

A mixed operation utilizing the two defined carrier spacing within one network is not foreseen in the DVB-C2 system due to the concept of absolute OFDM (see clause 5.9). Taking into account that DVB-C2 receivers may have a fixed tuning window (e.g. 8 MHz for European networks) and that the bandwidth of the preamble is linked to the choice of the carrier spacing, a mixed operation would not be optimal from a receiver's point of view. However, the DVB-C2 system does allow implementations of broadband services, where the receiver tuning window exceeds the 8 MHz for 2,232 kHz spacing or 6 MHz for 1,674 kHz spacing and where the receiver recovers the service by multiplexing the payload of different **PLPs** of different **Data Slice**.

There are no alternative carrier spacings used in DVB-C2. Fundamental features of C2 are the provision of **Data Slice** with their flexible number of carriers providing sufficient flexibility for configuring networks in practical applications of ETSI EN 302 769 [i.1]. The continued use of the existing channel raster in cable networks is expected to be the most commonly used operational mode of C2 in the medium term, i.e. the overall C2 channel bandwidth will be in most cases to a multiple of the existing cable channel raster (e.g. multiple of 8 MHz). In the long term DVB-C2 provides the features to overcome any channel raster in the cable network. The maximum theoretical bandwidth of a C2 signal is approximately 450 MHz, being limited from the parameter bitwidth in the **L1-part2** signalling.

## 6.3 Choice of Guard interval and impact of **OFDM symbol** Duration

DVB-C2 offers two options for guard interval fraction  $\Delta / T_U$ : 1/64 and 1/128, where  $T_U$  is the useful **OFDM symbol** period and  $\Delta$  is the duration of the guard interval. Selecting a guard interval longer than the channel impulse response enables inter-symbol interference to be removed in the DVB-C2 receiver. The normal approach in choosing the guard interval is therefore to determine the expected maximum echo delay in the cable network and then choose a  $\Delta$ , that matches or exceeds this.

In a cable environment, the channel conditions are static and therefore the normal considerations given to Doppler performance in satellite and terrestrial networks in selecting a guard interval are not relevant here.

Recent work into the study of cable network characteristics in modern cable network architectures has been undertaken by the European 'ReDeSign' project. This work identified that cable networks in the most common usage scenarios exhibit echo delays  $< 2,5 \mu\text{s}$  in length. Therefore, the expectation is the 1/128 option for the guard interval fraction will be the most commonly used. Professional cable networks with extended fibre reach have the potential to exhibit longer echo delays and hence the 1/64 option is provided for this purpose.

With a fixed maximum echo length then the guard interval fraction is inversely proportional to the **OFDM symbol** period which in turn is dependent on **FFT size**/carrier spacing. The symbol period used in DVB-C2 enables the guard interval fraction to be kept very small. This minimizes impact on spectral efficiency and thereby maximizes the potential data throughput. Due to the relation between guard interval size and required pilot density the overhead of scattered pilot patterns also decreases with shorter guard intervals.

Although 8 MHz is considered to be the primary transmission bandwidth, 6 MHz and 7 MHz are other examples of bandwidths that can be used in the application of DVB-C2. Echo tolerance improves in proportion to the reduction in bandwidth for the same **FFT size** as shown in table 8.

**Table 8: Relationship between Bandwidth, Echo Tolerance and Carrier Spacing**

Guard Interval Fraction	8 MHz Bandwidth		6 MHz Bandwidth	
	Echo Tolerance ( $\mu\text{s}$ )	Carrier Spacing (kHz)	Echo Tolerance ( $\mu\text{s}$ )	Carrier Spacing (kHz)
1/64	7,0	2,232	9,3	1,674
1/128	3,5	2,232	4,6	1,674
NOTE: The figures for echo tolerances in the above table are the maximum delays to be tolerated in a cable network for the available guard intervals.				

## 6.4 Choice of Pilot Pattern

DVB-C2 uses three types of pilots to aid the processes of synchronization and channel estimation in the DVB-C2 receiver. This clause covers the choice of pattern for scattered pilots.

A DVB-C2 receiver typically makes measurements of the channel using scattered pilots (SPs) and then interpolates between these measurements to construct estimates of the channel response for every **OFDM cell**. The measurements shall be sufficiently dense that they can follow channel variations as a function of both frequency (carrier index) and time.

The scattered pilot pattern density shall fulfil the sampling theorem in time and frequency direction. This can be summarized as:

- The maximum channel impulse response length to determine the repetition rate of the SPs in the frequency direction.
- The maximum Doppler frequency of the channel to determine the repetition rate of the SPs in the time direction.

The delay length can be determined from the guard interval period as described above in clause 5.3.

The cable channel is a static environment and therefore a consideration of the Doppler frequency is not required.

The reuse of many of the OFDM parameters from DVB-T2 in the DVB-C2 system led to some consideration of the pilot patterns already defined in T2 - could some be reused for C2 or were some new patterns required? The following aspects were considered:

- Maximum echo length is a small fraction of the **OFDM symbol** duration. This suggests a low pilot density would be sufficient for channel equalization in a **C2 system**.
- Use of frequency interpolation only is possible in a **C2 system**. Use of time interpolation should not be necessary but may improve the quality of channel estimation in cable networks with poor echo characteristics. No frame closing symbol is required as used in DVB-T2.

- Time repetition rate ( $D_Y$ ) should be as low as possible to avoid increasing memory requirements in the receiver.  $D_Y$  can be directly related to time interleaver depth.
- Use of a different pilot pattern for each guard interval.
- **Data Slice** granularity.

Patterns PP5 and PP7 were selected from DVB-T2 for DVB-C2 guard intervals of 1/64 and 1/128 respectively. The characteristics are summarized in table 9.

**Table 9: Comparison of scattered-pilot patterns**

GI	1/64	1/128	Interpretation
$D_X$	12	24	Separation of pilot-bearing carriers
$D_Y$	4	4	Length of sequence in symbols
$1/D_X \times D_Y$	2,08 %	1,04 %	SP overhead
$1/D_X$	1/12	1/24	$T_{\text{Nyquist}}/T_U$ , for f-&-t interpolation
$1/D_X \times D_Y$	1/48	1/96	$T_{\text{Nyquist}}/T_U$ , for f-only interpolation
$1/(2D_Y)$	0,125	0,125	$f_{\text{NYQUIST}}/f_S$ , for f-&-t interpolation ( $\pm$ )
	0,5	0,5	$f_{\text{NYQUIST}}/f_S$ , for f-only interpolation ( $\pm$ )

The operator should choose a guard interval size and the related pilot pattern considering the expected channel for the type of usage that it is desired to support, and being aware of the trade-off between capacity and performance.

Capacity is clearly reduced as the density of inserting scattered pilots increases; the SP overhead can be expressed as the fraction  $1/(D_X D_Y)$  where  $D_X$  is the separation of pilot bearing carriers and  $D_Y$  is the number of symbols forming one scattered pilot sequence. In DVB-T2 many of the SP patterns are optimized for performance in time varying channels with a correspondingly high overhead and can be disregarded for DVB-C2.

Note that there may be advantages to choosing a pilot pattern that supports a greater Nyquist limit for channel echoes than appears to be needed to support the chosen guard interval, despite the increase in overhead that such a choice implies. Circumstances supporting such an approach include:

- when the maximum channel echo delay is not well defined - while low-level signal components whose delays lie outside the nominal guard-interval window may not cause excessive ISI, they can cause aliasing in the channel measurement if they exceed the Nyquist limit;
- receiver interpolation performance will degrade as the Nyquist limit is approached;
- where the design time-width of the receiver's frequency interpolator is substantially less than the Nyquist limit, the interpolator can be designed to reduce estimation noise;
- in DVB-C2 the SP overhead is only 2 % in the worst case.

## 6.5 Choice of C2 frame and FECFrame length

Factors affecting choice of the **C2 frame** length include:

- A longer value for the frame length generally decreases the percentage overhead associated with the **preamble symbols** thus increasing the total bit-rate.
- A shorter value for the frame length enables the L1 signalling to occur more frequently, allowing faster lock-up and service acquisition.
- Longer frames may be used to support longer time-interleaving depths.

- Shorter frames may be used to achieve faster acquisition time.

The conclusion from considering the above factors was to fix the number of **data symbols** to 448 to enable a frame length of approximately 200 ms with some additional dependency on the guard interval length and the number of **preamble symbols** required.

The number of **preamble symbols** in a **C2 frame** is dependent upon the following:

- a minimum of one **preamble symbol** is required in a **C2 frame**. This simplifies the process of synchronization and channel estimation as described in clause 10.5.2;
- the maximum number of **L1-part2** signalling bits permitted in a **C2 frame** is 32 766, ETSI EN 302 769 [i.1], clause 8.3;
- the maximum number of L1 signalling bits in an **OFDM symbol** is 4 759, ETSI EN 302 769 [i.1], clause 8.4.2;
- the time interleaver mode. If the time interleaver is enabled then the number of OFDM symbols required for the preamble is predefined. If the time interleaver is off then the number of OFDM symbols is dependent on the parameters defined in the signalling subject to the above minimum and maximum constraints;
- the number of **Data Slice**;
- the number of **PLPs**.

The frame lengths in table 10 can be derived from this information.

**Table 10: Frame Durations in DVB-C2**

Ndata Symbols	Guard Interval	Ndata Duration (ms)	TI MODE	Npreamble symbols	% preamble overhead	Nsymbols in Frame	Frame Duration (ms)
448	1/128	202,3	00, 01	1 (minimum)	0,22	449	202,7
448	1/128	202,3	00, 01	7 (max.)	0,54	455	205,4
448	1/128	202,3	10	4	0,88	452	204,1
448	1/128	202,3	11	8	1,75	456	205,9
448	1/64	203,8	00, 01	1 (minimum)	0,22	449	204,3
448	1/64	203,8	00, 01	7 (maximum)	0,54	455	207,0
448	1/64	203,8	10	4	0,88	452	205,7
448	1/64	203,8	11	8	1,75	456	207,5

Table 10 shows the frame duration to vary by 2,3 % over all possible combinations of the C2 parameters selected.

Note various combinations of numbers of **Data Slice** and **PLPs** will cause the number of **preamble symbols** to be in the range of 1 to 7 when the time interleaver is either switched off or in 'best fit' mode (where the number of **preamble symbols** is minimized with the time interleaver enabled).

Two options are available for **FECFrame** length - 64 800 bits or 16 200 bits. The two different lengths correspond to the choice of long or short FEC blocks respectively. Originating from DVB-S2, the long FEC block was designated for broadcast operation and the short FEC block for non-broadcast operation. Short FEC blocks allow a finer granularity of bit-rate but incur a greater overhead and slightly degraded performance compared to long FEC blocks.

## 6.6 Choice of Time Interleaving Parameters

The time interleaver provides a variety of different interleaving depths and can also be switched off. Enabling the time interleaver for **L1-part2** may affect the data throughput, which depends on the selected mode (see table 11). The time interleaver may be enabled or disabled depending on the following factors:

- Use of some time interleaving modes for **L1-part2** (L1\_TI\_MODE='10' or '11') may reduce the data throughput, which is decided by the amount of signalling data.



- Cable channels with significant impulsive or burst noise will benefit from the time diversity provided by a time interleaver.
- Services with high QoS (e.g. VOD) can benefit from time interleaving.
- Interactive services requiring low latency (e.g. gaming) are not suited to the additional delays in processing incurred by using time interleaving.

TI parameters can vary between different **Data Slice**.

Time interleaving can be applied differently to **Data Slice** and to **L1-part2** signalling in the preamble. If time interleaving is applied to **Data Slice** it should also be applied to the **L1-part2** signalling to ensure the robustness of the signalling exceeds that of the **Data Slice**. The time interleaving depths available for the **Data Slice** are summarized in table 11.

**Table 11: Data Slice/L1-part2 Time Interleaving Parameters**

<b>DSLICE_TI_DEPTH</b>	<b>DSLICE TI Depth (OFDM symbols)</b>	<b>L1_TI_MODE</b>	<b>L1-part2 TI Depth (OFDM symbols)</b>	<b>Overhead (OFDM symbols)</b>
00	Off	-	-	0
01	4	-	-	0
10	8	-	-	0
11	16	-	-	0
-	-	00	Off	0
-	-	01	1(min) - 7(max)	0
-	-	10	4	0(min) - 3(max)
-	-	11	8	0(min) - 7(max)

## 6.7 Choice of Mode Adaptation

### 6.7.0 Introduction

There are a number of options for mode adaptation, described in clause 5.1 of ETSI EN 302 769 [i.1]:

- Normal Mode (NM) or High Efficiency Mode (HEM).
- Use of Input Stream Synchronization (ISSY).
- Use of Null Packet Deletion (NPD).

It is recommended that the options are used as follows:

- Where compatibility with DVB-S2 [i.5] is required, NM should be used with the same combination of ISSY and NPD as in the original DVB-S2 system. Otherwise:
  - For single **PLP**:
    - HEM should be used.
    - ISSY and NPD need not be used.
  - For multiple **PLP**:
    - HEM should be used.
    - ISSY should be used, especially for **PLPs** with variable bit-rates using null-packet deletion. Without ISSY, the receiver will have to attempt to manage its own de-jitter buffer and this might result in under-or overflow or excessive jitter in the output **Transport Stream**.
    - NPD should be used if statistical multiplexing is performed between **PLPs**, since null packets will be used to carry the variable bit-rate services in a constant bit-rate **Transport Stream**.
- For bundled **PLP**:
  - HEM should be used.

- ISSY should be used to synchronize packets from different **Data Slice** and as described above for multiple **PLPs**.
- NPD should be used as described above for multiple **PLPs**.

Available input formats for each **PLP** includes the new Generic Stream Encapsulation format (GSE) [i.7] + GSE implementation Guidelines [i.8] amongst other modes, for efficient encapsulation of IP and other network layer packets. This format was originally introduced to provide network layer packet encapsulation and fragmentation functions over the Generic Stream (GS) format in DVB-S2 [i.5] and has been recently adopted as an input format in DVB-T2 [i.3].

Some benefits/features of GSE as an alternative to carriage of IP over **Transport Streams** are as follows:

- Packets can be fixed or variable length.
- Reduces overhead of IP datagram transport from 10 % using (Multiple Protocol Encapsulation over MPEG2-TS to 2 % to 3 %.
- Provides more efficient system operation when the C2 physical layer uses Adaptive Coding and Modulation (ACM) for interactive services. GSE provides a mechanism for fragmenting IP datagrams or other network layer packets over baseband frames to support ACM/VCM.
- A smart scheduler can be used to take advantage of the flexible fragmentation and encapsulation methods available in GSE to optimize system performance. This can be by increasing the total throughput or minimizing average packet end-to-end delay. GSE flexibility allows the scheduler at the transmitter to dynamically change transmission parameters (e.g. modulation format, coding rate) for a particular network layer packet under channel fading variations for example.
- Support of multi-protocol encapsulation (e.g. IPv4, IPv6, MPEG, ATM, Ethernet, etc.).
- Transparency to network layer functions, including IP encryption and IP header compression.
- Support of several addressing modes.
- Low complexity.

### 6.7.1 Usage of the optional insertion of additional Null packet into TSPSs (Transport Streams Partial Streams)

Certain signal components in Digital TV signals have 'bursty' characteristics. This is especially the case for EMM messages, where the multiplexer allocates a fixed bit rate, but the CA-System itself tries to send entitlement messages as soon as possible, which e.g. results in a significant variation of EMM packets within certain time slots.

The option of inserting additional Null packets applies only in case that certain signal components of a **Transport Stream** are transmitted in a common **PLP**. In this case the DVB-C2 system does not guarantee that the overall propagation time due to different signal processing of a common **PLP** and the related data **PLPs** is identical. It may happen that a burst of common **PLP** packet does not fully match with null packets in the related data **PLPs**. This may be especially true for EMM-type of data.

The DVB-C2 specification [i.1] states in clause D.2.3: The number of inserted null packets shall be chosen such that a receiver with a 2 Mbit buffer is able to perform the multiplexing of Data **PLP** and Common **PLP** properly.

A DVB-C2 modulator therefore shall allow insertion of null packets and to re-stamp relevant TSPSs. The amount of inserted null packets shall be chosen with reference to the burst characteristics of the input signal.

The receiver does not know about those additional null packets. It recombines the relevant data **PLPs** and the common **PLP** to its output **Transport Stream**. In case of inserted null packets the bit rate of the reconstructed **Transport Stream** will be higher than the input **Transport Stream** of the modulator. The difference is equivalent to the bit rate of the inserted null packets.

## 6.8 Choice of Signalling Schemes

Selection of appropriate signalling schemes for C2 are defined by consideration of the following aspects:

- DVB C2 system architecture.
- The required rate of change of each signalling parameter.
- Overhead on data throughput.
- Compatibility with DVB family of standards.
- Compatibility with the C2 commercial requirements.
- Avoidance of duplication in signalling, e.g. between OSI Layer 1 and Layer 2.
- Future-proofing.

A key requirement is to differentiate between the signalling required for the physical layer (L1) and that the signalling required for the data link layer (L2). L2 signalling in DVB provides the mapping between the **PLP** of the physical layer and the **Transport Stream** so that the desired service can be selected in the receiver. This functionality is encapsulated in the C2 cable delivery system descriptor, which is defined in the DVB-SI specification ETSI EN 300 468 [i.9], clause 6.4.4. The L2 signalling is considered to be static as updates only occur via NIT version number.

L2 signalling is required to include the following:

- TS to **PLP** mapping data.
- Unique **C2 system** identifier.
- Tuning Parameters.

The standard tuning window envisaged for C2 is 8 MHz but the C2 standard uses the concept of variable **Data Slice** and notch frequencies. This requires the selection of the tuning frequency to cover all frequencies in the 8 MHz window, not only the **C2 system** centre frequency. In addition, tuning parameters can be common to several **PLPs** in the case where several **PLPs** share the same **Data Slice** as each **Data Slice** is associated with one tuning frequency.

The C2 system architecture requires the physical Layer 1 signalling to be split into two parts. The highest level of signalling (Layer 1 part 2) covers parameters at the **C2 frame** level and is encapsulated in the Preamble. This needs to include parameters defining **Data Slice**, **PLPs**, **notches**, guard intervals, etc. These parameters can change frame-by-frame. The lowest level of signalling (Layer 1 part 1) is defined at the C2 **FECFrame** level and includes parameters associated with QAM modulation order, FEC code rate, FEC block size, Early Warning System (EWS) etc. These parameters can change per FEC frame and therefore have the most significant impact on overall spectrum efficiency.

The number of available bits assignable to the signalling parameters is limited by the number of OFDM symbols assigned to the **FECFrame** header and the Preamble. Signalling parameters are required to be more robust in transmission than the data and therefore the FEC Frame header and Preamble use lower order modulation schemes and code rates to maximize robustness. This defines upper limits on the number of bits available for signalling.

## 6.9 Number of Data Slices versus PLPs

The cable operator needs to consider how to segment the required downstream services into and across **Data Slice** and **PLPs** to maximize the efficiency and flexibility offered by the C2 physical layer in the HFC network.

The following characteristics and features of DVB-C2 have been considered especially:

- The most common use case is 1 **PLP** and 1 **Data Slice** in an 8 MHz receiver tuning window. This is considered to be an important basic mode to re-multiplex complete streams into a single **Data Slice**. In this case the **Data Slice** can also be of the most efficient Type 1, i.e. the **Data Slice packets** only transmit the **FECFrame** data and rely on a pointer within the Level 1 Signalling part 2 to signal the first appearance of a new **Data Slice packet** within a new **C2 frame**.
- C2 allows for 1 **Data Slice** carrying 1 to 255 **PLPs** or 1 **PLP** spread across 1 to 255 **Data Slice**.

- A **PLP** may carry one or multiple services.
- **PLP** bundling enables a single **PLP** to be carried over several **Data Slice** - to enable transmission of a 200 Mbit/s service for example.
- Multiple groups of **PLPs** (up to 255) can be defined for a **C2 system**. Each group of **PLPs** has its own identifier in the L1 signalling (**PLP\_GROUP\_ID**). This enables a common **PLP** to be matched with a group of data **PLPs** for example. A group of data **PLPs** can exist without a linking common **PLP**.
- The number of **PLPs** and **Data Slice** for a **C2 system** have an upper limit determined by maximum number of bits (32 766) available in **L1-part2** signalling.
- Time and frequency interleaving are provided over each **Data Slice**, providing robustness against channel impairments. Multiple **PLPs** within a **Data Slice** will share the same time and frequency interleaving parameters. In general reducing the width of a **Data Slice** will improve the time diversity and degrade the frequency diversity and conversely increasing the width will degrade the time diversity but improve the frequency diversity.
- Frequency width and notch position can be signalled for a **Data Slice** but not for a **PLP**.
- Each **PLP** can have its own modulation, coding and FEC type signalled.
- If more than one **PLP** are multiplexed in a **Data Slice** the application of **Data Slice** Type 2 is mandatory (i.e. usage of the **FECFrame** Header).
- In case of using a Common **PLP** it has to be ensured that the overall bandwidth of the Common **PLP** and all other **PLPs** of the related **PLP** group are located within 7,61 MHz (or 3 408 subcarriers).
- The choice of the number of **PLPs** and the number of **Data Slice** may influence the performance of the **C2 system**, especially if narrow bandwidth **Data Slice** are chosen. Hence, for scenarios with more than one **PLP** in a **Data Slice** it should be the goal to maximize frequency diversity within the 7,61 MHz tuner reception window: Generally it is beneficial to multiplex as many **PLPs** within a **Data Slice** to achieve a **Data Slice** bandwidth close below or even equal to 7,61 MHz (or 3 408 subcarriers).
- Furthermore the DVB C2 System allows spreading data of a single **PLP** connection over different **Data Slice** (**PLP** Bundling, see annex F in ETSI EN 302 769 [i.1]). In this operation mode the throughput rate for a single **PLP** connection can be increased up to the overall throughput rate of the **C2 system**. However, this mode is not intended for regular broadcasting operation rather than for advanced services that require throughput rates above the capacity of a single **Data Slice**.

Clauses 7.3 and 12 provide further details and examples.

## 6.10 Notches

### 6.10.0 Discussion of the 'Narrowband' and 'Broadband' notch feature of the DVB-C2 system

Spectrum **notches** can be signalled in DVB-C2 to reduce the effect of emissions from cable systems into terrestrial frequency bands (e.g. aircraft radio or conversely ingress from high power interfering signals into the cable network). The notching features in ETSI EN 302 769 [i.1] are intended to enable C2 to coexist with other transmission systems whilst maintaining the maximum possible payload for C2 transmissions.

Notching reduces spectrum efficiency for C2 transmissions by removing selected carriers from the OFDM signal. The notching parameters available in the L1 signalling enable the reduction of spectrum efficiency to be kept at a minimum by precisely defining the start position and width of the spectrum notch to the nearest 12 or 24 subcarriers, depending on guard interval depth.

There are two types of **notches** defined - 'Narrowband' and 'Broadband' **notches**. Notches are considered to be static in terms of their position in the frequency spectrum. Notch characteristics are summarized in table 12.

Table 12: Notching Characteristics in DVB-C2

	Narrowband notches	Broadband notches
<b>Bandwidth</b>	11, 23, 35 or 47 carriers (1/64GI) 23 or 47 carriers (1/128GI)	11 carrier (1/64GI and 23 carriers (1/128 GI) minimum bandwidth no maximum, but one preamble adjacent to a Broadband notch
<b>Notch position within a Data Slice</b>	Yes	No
<b>Signalled in L1</b>	Yes	Yes
<b>Receiver handling</b>	Notch corrected by L1 FEC	Notch not inside tuning bandwidth

NOTE: Receiver performance in the presence of **notches** will be degraded due to imperfect equalization, and this should be taken into account in the choice of modulation and coding parameters of **Data Slice** close to **notches**. The use of particularly robust modulation and coding is recommended for **Dependent Static Data Slice** of less than 3 408 carriers.

Narrowband **notches** are treated differently for preamble and **data symbols**: To allow initial decoding of the **preamble symbols** without any knowledge of the C2 signal, the overall structure and location of the L1 signalling blocks shall not be changed. Preamble subcarriers within narrowband **notches** are therefore simply blanked but can be recovered by the robust FEC on the receiver side. In contrast, notched subcarriers in **data symbols** are not used for payload transmission.

Up to 15 discrete **notches** can be signalled in a **C2 frame**, enabling a mix of non C2-related interfering signals and sensitive spectrum areas to coexist with C2 transmissions. Note that not more than one narrowband notch is allowed at any tuning position of 3 408 OFDM subcarrier bandwidth (see clause 9.4.5.1 of ETSI EN 302 769 [i.1]).

Broadband **notches** (**notches** with bandwidth above 105 kHz) cannot exist within a **Data Slice** as they interfere with the process of the receiver tuning to a **Data Slice**. Practically the bandwidth of broadband **notches** is limited by the bit width of the related **L1-part2** signalling field

- GI = 1/64:  $511 \times 12 - 1 = 6\ 131$  subcarriers
- GI = 1/128:  $255 \times 24 - 1 = 6\ 119$  subcarriers

Notches of any type shall not overlap with another notch.

In general the notching features available in C2 should only be used to mitigate the effects of possible interference scenarios and maximize the C2 payload. However, these features could be used to enable other types of new transmissions to co-exist in the RF spectrum occupied by C2 in the knowledge that C2 services can still operate either side of defined notch frequency bands. Additionally where non-C2 interfering signals are known to exist but at a medium/low level then notching may not be necessary. In this case, the comprehensive error correction provided by the C2 system architecture can minimize any possible performance degradation without the necessity to switch off carriers in the transmitter.

Clause 12.1.6 gives some examples of the application of notching in C2 cable systems.

## 6.10.1 OFDM sub-carrier bandwidth and Receiver Considerations with Broadband **notches**

### 6.10.1.0 Introductory remarks

The following two clauses 6.10.1.1 and 6.10.1.2 relate to clause 9.3.5.2 of ETSI EN 302 769 [i.1].

### 6.10.1.1 Minimum Data Slice Width

The minimum width of a **Data Slice** is constrained by the presence of an adjacent broadband notch. The C2 receiver requires a minimum number of Continual Pilots (CPs) for signal processing (e.g. for Common phase error correction). Table 13 shows the relationship between numbers of CPs and the **Data Slice**/OFDM bandwidth. Columns for numbers of CPs on non-SP bearing carriers are also shown as additional information.

Table 13: Number of Continuous Pilots in Different C2 sub-carrier bandwidths

Bandwidth (MHz)	D <sub>X</sub>	NUM_CARRIERS	MAX_CP_COUNT	MIN_CP_COUNT	MAX CP_non_SP COUNT	MIN CP_non_SP COUNT
7,61 (full b/w)	12	3 408	30	30	14	14
7,61 (full b/w)	24	3 408	30	30	22	22
4,00	12	1 788	18	13	10	5
4,00	24	1 776	18	13	14	9
3,80 (1/2 b/w)	12	1 704	17	13	9	5
3,80 (1/2 b/w)	24	1 704	17	13	13	9
3,00	12	1 332	15	9	8	3
3,00	24	1 320	15	9	13	6
2,00	12	888	12	5	7	1
2,00	24	888	12	5	9	2
1,90 (1/4 b/w)	12	852	12	5	7	1
1,90 (1/4 b/w)	24	840	11	5	9	2
1,00	12	444	6	1	4	0
1,00	24	432	6	1	5	0
0,95 (1/8 b/w)	12	420	6	1	4	0
0,95 (1/8 b/w)	12	408	6	1	5	0

Table 13 shows how at bandwidths of 1 MHz and lower the minimum number of CPs available for signal processing can be as low as 1 - this is insufficient. Reducing the number of CPs for processing increases the susceptibility to noisy carriers, particularly outside the normal SNR environment (e.g. CW interference). This is discussed further in clause 10.5.1. Five CPs has been defined as the minimum acceptable number of CPs for the receiver to work reliably and hence the minimum bandwidth of the received OFDM signal is around 1,9 MHz from the above table.

The number of available CPs could be artificially increased by treating the edge carriers for **Data Slice** neighboured by **notches** as CPs. However, in the receiver the edge pilots suffer from the "edge effect" in frequency interpolation. Since there are no pilots after the edge pilots, the filter coefficients have to be interpolated based on holding the edge pilot value. This introduces errors for all pilots and data carriers until the filter coefficients have been correctly filled. Additionally the broadband **notches** can also contain noise/interference which will introduce further errors into the pilots close to the border of the broadband notch and C2 OFDM signal.

### 6.10.1.2 Dependent Static Data Slices

Decoding of the **L1-part2** signalling data from the preamble cannot be reliable in the case where a **Data Slice** with less than 3 408 carriers exists between broadband **notches**. This is a problem for a **Data Slice** Type 1 as a pointer, which points to the start of the first **FECFrame** packet, is normally decoded from the **L1-part2** signalling. In this instance **L1-part2** signalling data cannot be decoded and so the pointer is unknown. Therefore it is necessary to use **Data Slice** Type 2 with **FECFrame** headers as the starting position of the **FECFrame** packet can be obtained from the **FECFrame** header. An additional constraint imposed by the absence of **L1-part2** signalling is that the structure and parameters of this **Data Slice** need to remain fixed by setting **DSLICE\_CONST\_CONF** flag to 1.

## 6.11 Choice of Layer 1 signalling parameters

### 6.11.1 Choice of the "Data Slice Tuning Position" parameter

To avoid the ingress of spectral artefacts which are not part of the signal spectrum of the DVB-C2 signal itself the value of the parameter **DSLICE\_TUNE\_POS** of each **Data Slice** of the DVB-C2 system might be selected such that the amount of out-of-band interference within the spectrum preselected by the tuner window is minimized.

The most common cases of out-of-band interference are the spectrum to lower and higher frequencies of the DVB-C2 signal, i.e. the spectrum exceeding the range defined by  $K_{\min}$  and  $K_{\max}$ , as well as **notches** within the signal spectrum itself.

For the first case the reduction of spectral artefacts might easily be achieved if the following inequations are applied:

$$\text{START\_FREQUENCY} + (\text{DSLICE\_TUNE\_POS} - \text{DSLICE\_OFFSET\_LEFT}) \times D_X \geq K_{\min}$$

$$\text{START\_FREQUENCY} + (\text{DSLICE\_TUNE\_POS} + \text{DSLICE\_OFFSET\_RIGHT}) \times D_X \leq K_{\max}$$

The influence of the spectral artefacts within the frequency range of a narrow band notch might be negligible and thus be ignored for the selection of the `DSLICE_TUNE_POS` parameter, i.e. referring to clause 9.3.5.1 of the DVB-C2 specification "Only one narrowband **notch** shall be allowed at any tuning position of 3 408 OFDM subcarriers [...] bandwidth."

In case of broadband **notches** equivalent equations are applied as for the reduction of out-of-band influences. Here, the `DSLICE_TUNE_POS` parameter shall be selected according to clause 9.3.5.2 of the specification ("For **Data Slice** decoding the receiver tuner shall not tune to a region with a broadband **notch** [...]") such that the tuner window at the lower or higher frequency edge, respectively, aligns with the ending or beginning, respectively, of the broadband notch.

If a part of the DVB-C2 signal spectrum which is to be decoded is bounded by either two broadband **notches** or by a broadband notch and the signal spectrum itself, the `DSLICE_TUNE_POS` parameter is for simplicity reasons selected to be located in the mid of the contiguous DVB-C2 signal spectrum. Alternatively, `DSLICE_TUNE_POS` might be chosen such that only the spectrum of the broadband notch is covered excluding any ingress from a neighbouring channel on the opposite spectrum edge of the DVB-C2 signal.

## 6.11.2 Usage of 'Early Warning System' signalling

The 'Early Warning System' signalling bit is embedded in the L1 signalling. This allows for low latency detection of an alarm event as it is available at the beginning of every **C2 frame**. Furthermore receivers in sleep mode might be able to detect emergency events by processing the **preamble symbol** only.

In case the DVB-C2 modulator gets its signal via a satellite or terrestrial feed (transmodulation application), the 'Early Warning System' signal of the satellite feed can be automatically transferred to the DVB-C2 modulator.

DVB has not defined physical interfaces, both on the modulator and on the demodulator side for the delivery or the reception of the 'Early Warning System' signalling.

In order to establish an 'Early Warning System' further standardization of functionalities and interfaces is required. This is out of scope of the DVB-C2 modulation system specification.

# 7 Input Processing / Multiplexing

## 7.1 Generation of the **FECFrame** Header

Two encoding schemes of the **FECFrame** Header are shown in the figures 9(a) and 9(b). Initially the 16 bits of the L1 signalling part 1 are FEC encoded by a Reed-Muller (32,16) encoder. Subsequently each bit of the 32 bit Reed-Muller codeword is split to form an upper and a lower branch. The lower branch applies a cyclic shift within each Reed-Muller codeword and scrambles the resulting data using a specific PN sequence. The difference of two encoding schemes is that a QPSK constellation is used for the robust **FECFrame** header and a 16-QAM constellation is used for the high efficiency **FECFrame** header.

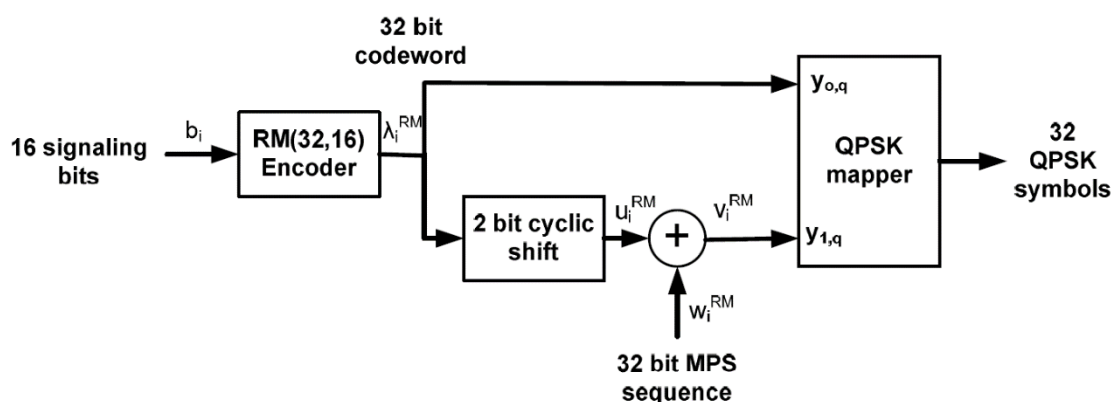


Figure 9(a): Robust **FECFrame** header

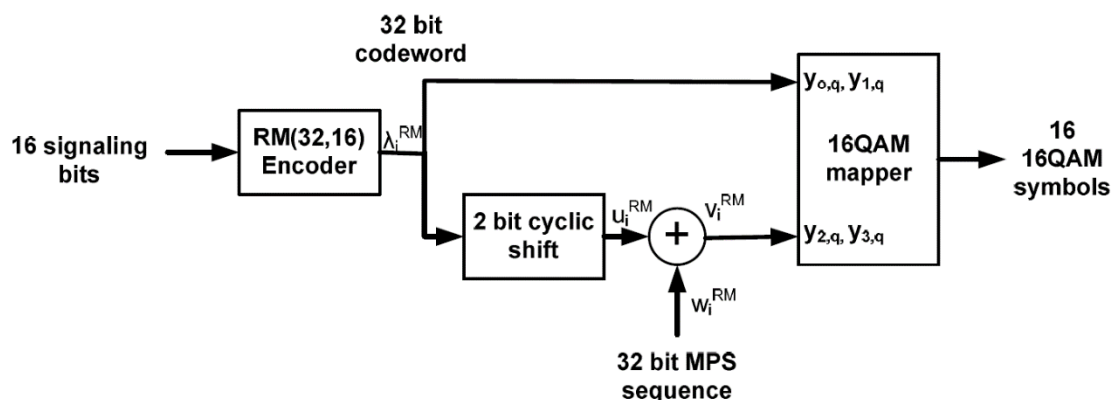


Figure 9(b): High efficiency FECFrame header

The 32 Reed-Muller [i.14] encoded data bits  $\lambda_i^{RM}$  of the lower branch are cyclically delayed by two values within each Reed-Muller codeword. The output of the cyclic delay block is:

$$u_{(i+2)_{32}}^{RM} = \lambda_i^{RM} \quad i = 0, 1, \dots, 31.$$

## 7.2 Use of common PLPs

A common **PLP** is only required when multiple TSs are transmitted, and these TSs are called as 'group of **PLPs**'. The common **PLP** of a group of **PLPs** is transmitted in such a way that a receiver can receive simultaneously any data **PLP** of the group as well as the common **PLP**.

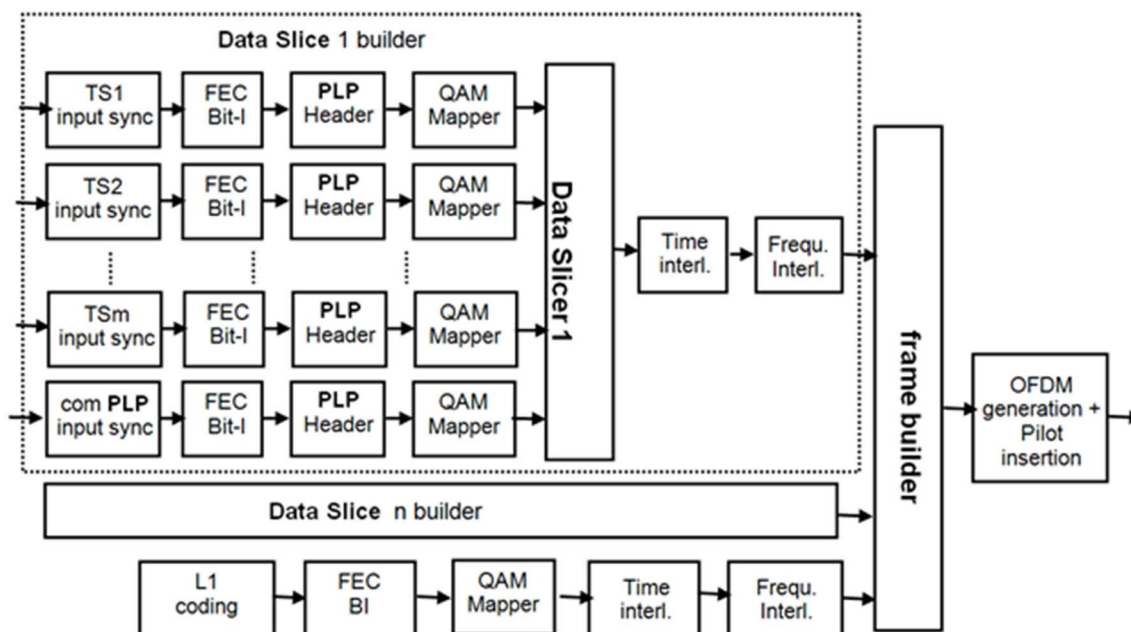


Figure 10: DVB-C2 modulation including common PLP concept

A cable playout centre generating DVB TSs uses statistical multiplexing in order to optimize the overall spectrum efficiency. The capacity for CA-messages data and SI data has to be fixed in order to calculate the overall payload capacity available for video and audio streams within the statistical multiplex. The relevant application is also possible if several TSs have to be transmitted in one cable channel or if packages of services require different levels of robustness.



Also, common **PLP** can be used to decentralize cable head end (e.g. several Astra-type DTH-signals are locally multiplexed into one cable channel), or to centralize head end where premium tier HDTV services and basis tier SDTV service are required to be transmitted in the same cable channel. This ensures that different levels of robustness are available for QoS reasons.

So the TS packets that may be transmitted in the Common **PLP** are:

- 1) TS packets carrying any type of data, which does not require an exact time synchronization with other TS packets carried in other **PLPs** of the related group of **PLPs**.
- 2) EPG data, e.g. based on Event Information Table (EIT) format, but not using the 'actual' and 'other' mechanisms as specified in the DVB-SI specification ETSI EN 300 468 [i.9].
- 3) Conditional Access control data, e.g. Entitlement Management Messages (EMMs).

When using common **PLP** in a group of **PLPs**, the existence of common **PLP** should be signalled in the L1 signalling. The **PLP\_TYPE** of the common **PLP** should be set to '00', and the group of **PLPs** should have the same **PLP\_GROUP\_ID** with the common **PLP**. See clause 8.3 in ETSI EN 302 769 [i.1].

To get the information about splitting of input MPEG-2 **Transport Streams** into Data **PLPs**, generation of a Common **PLP** of a group of **PLPs** and insertion of Null Packets into **Transport Streams**. See annex D in ETSI EN 302 769 [i.1].

### 7.3 PLP bundling and Statistical Multiplexing

In case of input signals that exceed the capacity of a single **Data Slice** (e.g. 200 Mbit/s), the **XFECFrames** of the input signal has to be distributed into different **Data Slices** with one **PLP** each. If the input signal consists of HDTV services encoded in statistical multiplexing, the size of those three **Data Slices** will vary over time. The modulator can adapt the size of **Data Slices** per C2-frame.

As the total input stream in the upper example is 200 Mbit (constant bit rate) the sum of sizes of the three **Data Slices** is constant. According to the characteristics of the statistical multiplexing, the number of cells of payload allocated to the three **Data Slices** slightly varies. If a Common **PLP** is used, it has to be transmitted twice in order to allow every relevant receiving window to access the common **PLP** data.

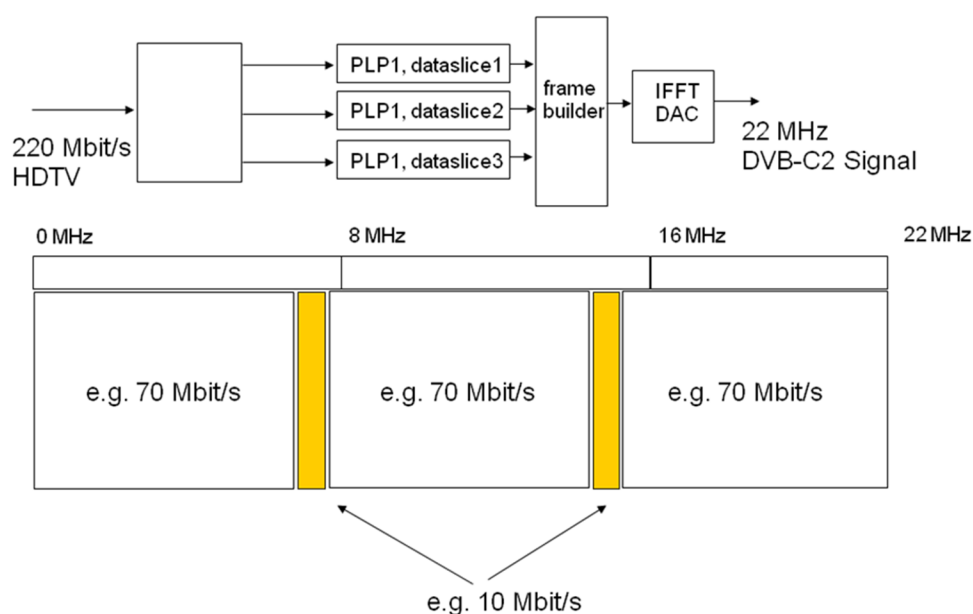


Figure 11: PLP bundling use case example

**PLP** bundling is signalled by '**PLP\_BUNDLED**' field in the layer-1 signalling. If one TS is bundled over multiple **PLPs**, all the **PLPs** will have the same **PLP\_ID**, even if all the **PLPs** are in different **Data Slices**. Also all the **PLP\_BUNDLED** field for bundled **PLPs** will be set to '1'. The layer-2 signalling will not reflect the **PLP** bundling usage, because layer-2 signalling will only inform the mapping relation between specific transport\_stream\_id to specific **PLP\_id** - that is why all the bundled **PLPs** share the same **PLP\_ID**.

To guarantee the order of each packet spread over multiple **PLPs**, ISSY timestamp should be used in the mode adaptation block. This field will be used in the receiver side to re-order packets received from multiple **PLPs** bundled.

See annex F in ETSI EN 302 769 [i.1].

**PLP** bundling within a 7,61 MHz **Data Slice** is not allowed. It is always more effective to carry a data stream within one **PLP** in a **Data Slice**.

Bundling of **PLPs** therefore only makes sense if it not possible to carry a data stream within a single **Data Slice**. For receivers with a fixed receiving window (e.g. 8 MHz for European networks) the capability to recombine bundled **PLPs** is not mandated. However, those receivers normal operation shall be unaffected by the presence of bundled **PLPs**.

Any operator, who wants to use the **PLP** bundling concept for dedicated services, e.g. for transmitting big pipes of data, therefore has to ensure that the receivers used either are equipped with multiple fixed bandwidth tuners or with an appropriate wideband tuner.

The use of **PLP** bundling is discussed further in annex C.

## 7.4 Use of the Null Packet Deletion mechanism

### 7.4.0 Introduction

Null Packet Deletion is an option offered by DVB-C2 that removes the Null Packets (stuffing data) at the transmitter side from the MPEG-2 **Transport Stream**. However, the receiver is able to re-insert the deleted Null Packets at their original positions, as these positions are signalled within the transmitted data (see clause 5.1.4 of ETSI EN 302 769 [i.1]). Different application for this mechanism may exist. The text below lists three examples for the application of the Null Packet Deletion mechanism.

#### 7.4.1 Use of Null Packet Deletion with Common **PLP**

The Common **PLP** allows for the transmission of **Transport Stream** packets, which are common for multiple **Transport Streams**. At the receiver side the Common **PLP** and the Data **PLP** are merged in order to obtain the fully syntactically correct **Transport Stream**. Therefore this requires sufficient Null Packets within the individual Data **PLPs**, as the Common **PLP** data can only be reinserted at existing Null Packet positions within the Data **PLPs** for the reestablishment of the targeted **Transport Stream**. The Common **PLP** would bring no benefit if the transmitter has to actually transmit these Null Packets required for the insertion of the Common **PLP**. Therefore, the transmitter can add sufficient Null Packets, but it does not actually have to transmit these due to the Null Packet Deletion.

#### 7.4.2 Use of Null Packet Deletion for Efficient Statistical Multiplexing

Statistical Multiplexing can improve the source coding gain. Normally, different services that are jointly encoded are transmitted within a single MPEG-2 **Transport Stream**. However, for Video on Demand services it is also possible to transmit these services using multiple **PLPs** (**Data Slice** Type 2 only), which e.g. allows for the application of independent modulation and coding parameters for each user. Naturally, this hinders the application of Statistical Multiplexing that is normally used within a single **Transport Stream**, only. Though, using Null Packet Deletion, Statistical Multiplexing can also be applied on multiple **PLPs**. Therefore, the **Transport Stream** within each **PLP** is stuffed to a specific bit-rate using Null Packets while the actual payload bit-rate is varying. Certainly, the bit-rate of the stuffed **Transport Stream** has to be higher than the maximal payload bit-rate. The inserted Null Packets are then deleted using the Null Packet Deletion mechanism, leading to variable bit-rate streams for each **PLP**, which can be statistically multiplexed. At the receiver side, the removed Null Packets are inserted again, leading to a fixed bit-rate **Transport Stream**.

### 7.4.3 Null Packet Deletion for Reduced Power Consumption

The Null Packet Deletion mechanism can be applied for reduced power consumption within the receiver (**Data Slice** Type 2 only). Instead of using the Null Packets within the **Transport Stream** to obtain a specific bit-rate, these Null Packets can be removed using the Null Packet Deletion. Then Stuffing **Data Slice** Packets are inserted to fill the remaining **Data Slice** capacity. Hence, the resulting bit-rate of the **PLP** is reduced, which also reduces the amount of data that has to pass the Forward Error Correction decoder within the receiver. As the FEC decoder is one major contributor to the overall power consumption of the DVB-C2 decoder, the power consumption can be decreased significantly.

## 7.5 Use of Issy Time Stamping

The use of the Input Stream Synchronizer subsystem is mandatory for bundled **PLPs** (see annex F of ETSI EN 302 769 [i.1]). For non-bundled **PLPs** the use of the Input Stream Synchronizer subsystem is optional for **PLPs** carrying GSE, GCS or GFPS streams, but in the case of **PLPs** carrying **Transport Streams** (TS), it shall always be used, except when all of the following conditions are met:

- 1) FECFRAME header is not used (i.e. DSLICE\_TYPE=0); this requires that there is only one **PLP** (i.e. DSLICE\_NUM\_PLP=1) and modulation and coding parameters are fixed, and Constant Coding and Modulation (CCM) is signalled).
- 2) DFL is constant (DFL=K\_bch-80 in every **BBFrame**, BB padding is not used).
- 3) The number of **data cells** per **C2 frame** is constant (**Data Slice** width is constant).
- 4) The **C2 frame** length is constant (the number of **preamble symbols** and GI length does not change).
- 5) Null Packet Deletion is not used (i.e. NPD=0) in order that the TS bit rate can be calculated from the transmitted L1 signalling.

In line with annex C of ETSI EN 302 769 [i.1] the usage of the ISSY field shall comply with the following characteristics:

ISSY shall be coded according to table 14, sending the following variables:

- ISCR (short: 15 bits; long: 22 bits) (ISCR=Input Stream Time Reference), loaded with the LSBs of the counter content at the instant the relevant input packet is processed (at constant rate R\_IN), and specifically the instant the MSB of the relevant packet arrives at the modulator input stream interface. In case of continuous streams, the content of the counter is loaded when the MSB of the Data Field is processed. ISCR shall be transmitted in the ISSY field whenever BUFS and BUFSTAT are not transmitted. In a given **PLP**, either ISCR\_short or ISCR\_long shall be used, together with the short or long versions of BUFS. A **PLP** shall not change from short to long ISSY except at reconfiguration. In HEM, ISCR\_long shall always be used.
- BUFS (2+10 bits) (BUFS= maximum size of the requested receiver buffer to compensate delay variations). This variable indicates the size of the receiver buffer assumed by the modulator for the relevant **PLP**. It shall have a maximum value of 2 Mbit. When a group of Data **PLPs** share a Common **PLP**, the sum of the buffer size for any Data **PLP** in the group plus the buffer size for the Common **PLP** shall not exceed 2 Mbit. BUFS shall be transmitted in the ISSY field of the Baseband Header of the first non-fragmented **BBFrame** in a **C2 frame**. In NM, BUFS shall be transmitted in the ISSY field of the first complete packet of the first non-fragmented **BBFrame**. In HEM, BUFS shall also be transmitted for **BBFrames** for which no UP begins in the data field. This field shall not be transmitted in case of **PLP** bundling as the ISCR field is required in every **BBFrame** for sorting the input stream.
- BUFSTAT (2+10 bits) (BUFSTAT = actual status to reset the receiver buffer = number of filled bits). In HEM, BUFSTAT shall be transmitted in the ISSY field following the transmission of BUFS. BUFSTAT can be used to set the receiver buffer status during reception start-up procedure, and to verify normal functioning in steady state. This field shall not be transmitted in case of **PLP** bundling as the ISCR field is required in every **BBFrame** for sorting the input stream.

NOTE 1: This requires that there are always at least three ISSY fields in every **C2 frame**. It might be necessary to use short FEC blocks and/or Normal Mode in order to ensure that this is the case. Furthermore, ISSY applies to a transmitted User Packet and so it might be necessary to transmit a null packet which would otherwise be deleted to provide a packet for the ISSY field to refer to.

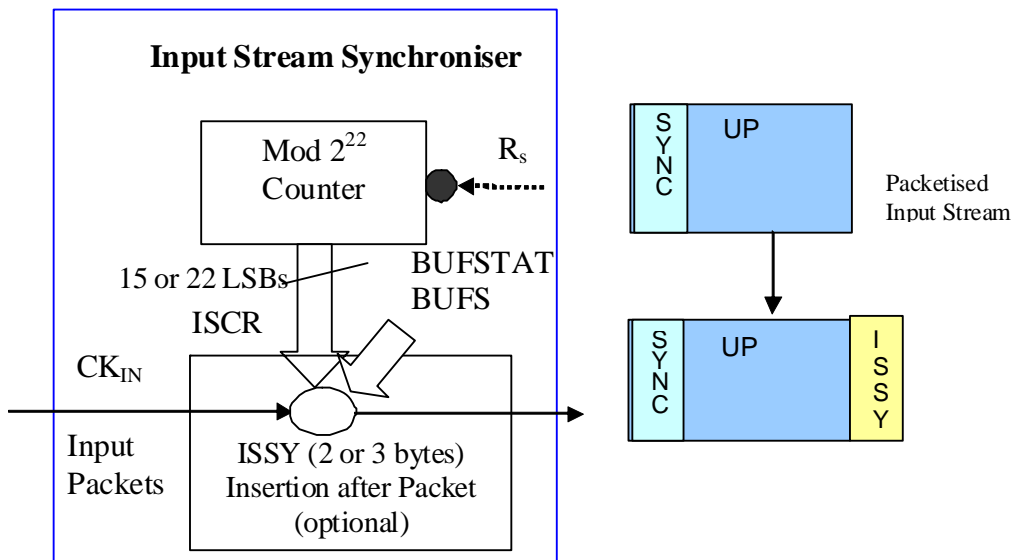


Figure 12: Input stream synchronizer block diagram

Table 14: ISSY field coding (2 bytes or 3 bytes)

First Byte					Second Byte	Third Byte
Bit-7 (MSB)	bit-6	Bit-5 and bit-4	bit-3 and bit-2	bit-1 and bit-0	Bit-7 to bit-0	bit-7 bit-0
0 = ISCR <sub>short</sub>	MSB of ISCR <sub>short</sub>	next 6 bits of ISCR <sub>short</sub>			Next 8 bits of ISCR <sub>short</sub>	Not present
1	0 = ISCR <sub>long</sub>	6 MSBs of ISCR <sub>long</sub>			Next 8 bits of ISCR <sub>long</sub>	Next 8 bits of ISCR <sub>long</sub>
1	1	00 = BUFS	BUFS unit 00 = bits 01 = Kbits 10 = Mbits 11 = 8 Kbits	2 MSBs of BUFS	Next 8 bits of BUFS	Not present when ISCR <sub>short</sub> is used; else reserved for future use
1	1	10 = BUFSTAT	BUFSTAT unit 00 = bits 01 = Kbits 10 = Mbits 11 = BUFS/1 02 4	2 MSBs of BUFSTAT	Next 8 bits of BUFSTAT	Not present when ISCR <sub>short</sub> is used; else reserved for future use
1	1	Others = reserved for future use	Reserved for future use	Reserved for future use	Reserved for future use	Not present when ISCR <sub>short</sub> is used; else reserved for future use

NOTE 2: The BUFSTAT unit "BUFS/1 024" should only be used when the value of BUFS unit is not "00", i.e. BUFS unit is represented by Kbits, Mbits or 8 Kbits. The most accurate BUFSTAT unit should always be used.

In line with clause 10.7 the **Transport Stream** regeneration and clock recovery using ISCR (annex G of ETSI EN 302 769 [i.1]) is described below.

When the modulator operates in a mode that employs Null Packet Deletion, the receiver may regenerate the **Transport Stream** by inserting DNP into the useful packets, which are read out from the FIFO buffer. As shown in figure 13, the **Transport Stream** clock  $R'_{IN}$  may be recovered by means of a Phase Locked Loop (PLL). The recovered modulator sampling rate  $R_s$  may be used to clock a local counter (which by definition runs synchronously with the input stream synchronization counter of figure 13). The PLL compares the local counter content with the transmitted ISCR of each TS packet, and the phase difference may be used to adjust the  $R'_{IN}$  clock. In this way  $R'_{IN}$  remains constant, and the reception FIFO buffer automatically compensates the chain delay variations. Since the reception FIFO buffer is not self-balancing, the BUFSTAT and the BUFS information may be used to set its initial state.

As an alternative, when dynamic variations of the end-to-end delay and bit-rate may be acceptable by the source decoders, the receiver buffer filling condition may be used to drive the PLL. In this case the reception buffer is self-balancing (in steady state half the cells are filled), and the ISSY field may be omitted at the transmitting side.

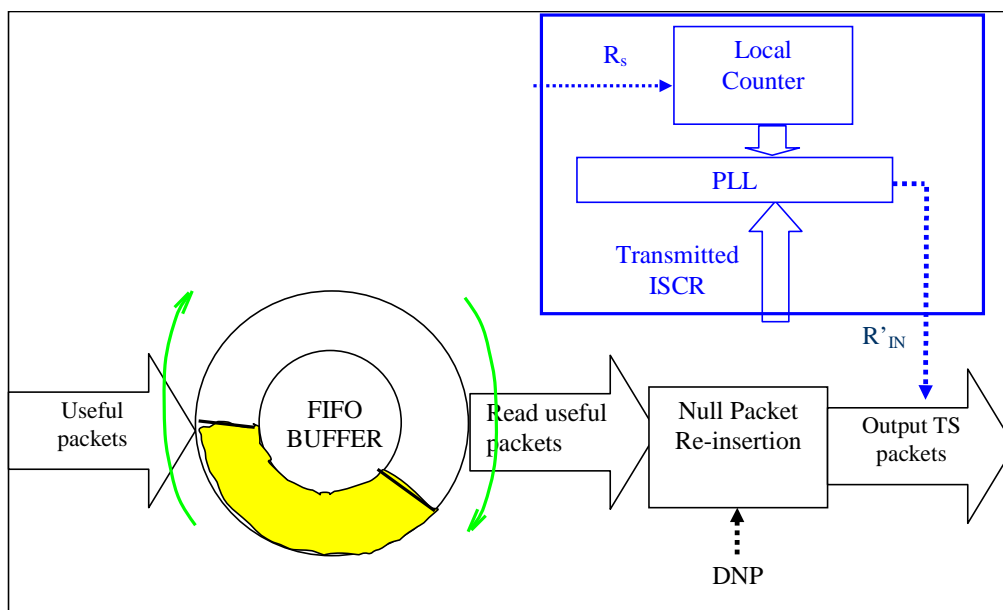


Figure 13: Example receiver block diagram for Null Packet re-insertion and  $R_{TS}$  clock recovery

## 7.6 Stuffing Mechanism

### 7.6.0 Introduction

The DVB-C2 system offers multiple stuffing means. This stuffing is mainly required to adapt the bit rate of the incoming signal to the bit rate of the current **C2 system** settings. The different mechanisms are described in the following clauses.

#### 7.6.1 Transport Stream Stuffing

The stuffing on **Transport Stream** level by means of Null Packets is a well-known technique to adapt the bit rate of an input stream to a fixed output bit rate. For details see ISO/IEC 13818-1 [i.10].

#### 7.6.2 Base Band Frame Stuffing

The Base Band Frame Stuffing is the recommended stuffing mode for the lower layer stuffing if **Data Slice** type 1 is used. It may be necessary if the **Data Slice** bit rate is not adjustable to the payload bit rate ideally (e.g. due to the granularity of **Data Slice** bandwidth). Hence, the Base Band Frame Stuffing is a means to finally adjust the bit rate to the input stream bit rate. Such a technique is of special interest if the network provider is e.g. not allowed to re-multiplex the MPEG-2 **Transport Stream** due to political regulations.

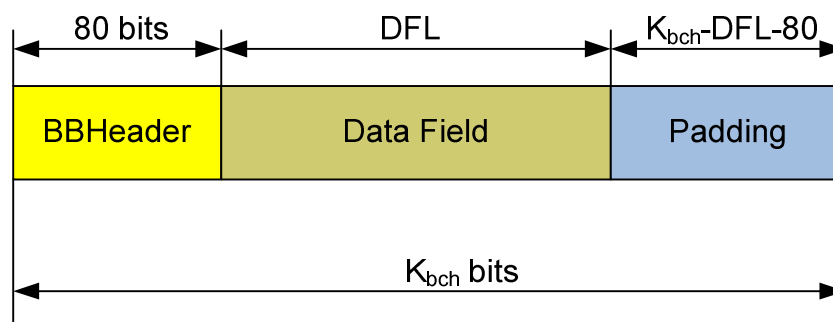


Figure 14: Structure of BBFrame with padding field

Figure 14 shows the structure of a Base Band Frame. The **BBHeader** is followed by the Data Field, which is followed by optional padding. Hereby, the length of the Data Field is signalled within the **BBHeader**. Thus, if the amount of data is not sufficient to fill the Base Band Frame stuffing is added and the actual number of payload bits in the **BBFrame** is reduced to the requirements. It is even possible to transmit **BBFrames** that consist of stuffing data only.

### 7.6.3 Data Slice Packet Stuffing (only Data Slice Type 2)

**Data Slice** Packet Stuffing is the recommended stuffing mechanism for lower layer stuffing if **Data Slices** type 2 is used, because it reduces the amount of data the LDPC decoder has to process. If the transmitter does not have any payload **Data Slice** Packet to transmit, it simply transmits a Stuffing **Data Slice** Packet (see clause 7.2.6 of ETSI EN 302 769 [i.1]). The length of this Stuffing **Data Slice** Packet is fixed to 900-QAM cells plus the **Data Slice** Packet Header (32-QAM or 16-QAM cells).

### 7.6.4 Data Slice Stuffing

DVB-C2 offers the possibility to adapt the bandwidth of the **Data Slices** between consecutive frames without service interruption, which is also a stuffing mechanism.

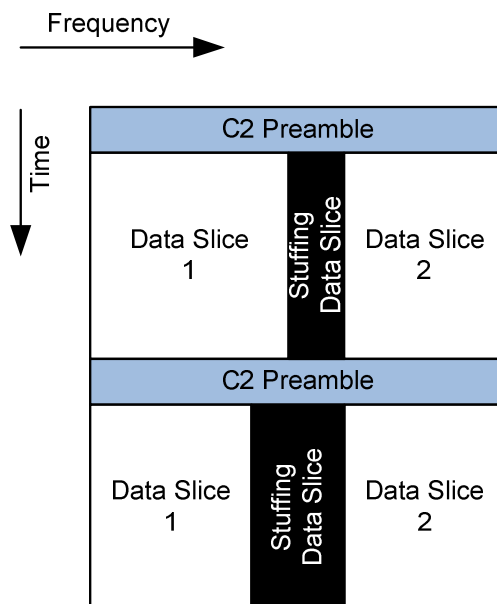


Figure 15: Principle of Data Slice Stuffing

The principle of **Data Slice** Stuffing is depicted in figure 15. The two (payload) **Data Slices** do not fill the complete bandwidth of the **C2 system**. Hence, a Stuffing **Data Slice** is included between the two **Data Slices**. However, this Stuffing **Data Slice** is only signalled implicitly, as all remaining OFDM subcarriers that are not occupied by **notches** or **Data Slices** are Stuffing **Data Slices**. The bandwidth and the positions of Stuffing **Data Slices** may vary between **C2 frames**, as the positions and the bandwidth of **Data Slices** may vary as well.

Within Stuffing **Data Slices** the normal pilot scheme (i.e. scattered, continual and edge pilots) and the tone reservation are continued. Additionally, also **notches** may be placed within Stuffing **Data Slices**. The modulation of the cells not mapped to pilots or reserved tones can be chosen freely. There are only the limitations that these cells shall carry the mean power level of 1 and that repetitions of the same sequence shall be avoided in order to prevent high peaks of the OFDM time domain signal.

## 7.7 Multiplexing, Dimensioning of PLPs and Data Slices

### 7.7.0 Introduction

The following clauses list possible scheduling algorithms. Those are based on the MPEG-2 **Transport Stream**. However, extensions to other protocol schemes are possible. For simplicity, it is assumed that the available cell rate within the **Data Slice** is always higher than the required cell rate of the multiplexed payload data.

#### 7.7.1 Single PLP per Data Slice

##### 7.7.1.0 The Single PLP per Data Slice application

When using a single **PLP** per **Data Slice** only, no complex multiplexing is required. However, as the available bit-rate of the **Data Slice** will not match the bit-rate of the input stream, stuffing is required. As already explained in clause 7.3, different possibilities exist. Their application depends on the used **Data Slice** type.

For simplicity, it is assumed that the bandwidth of the **Data Slice** does not change over time, the input is an MPEG-2 **Transport Stream**, and a Common **PLP** is not used.

##### 7.7.1.1 Data Slice Type 1

Figure 16 depicts the scheduler for **Data Slice** Type 1. It is assumed that the available **Data Slice** bit-rate is sufficient to transmit the input data stream. As the output stream (i.e. **Data Slice** Packets) has to have a fixed output bit-rate, the application of Null Packet Deletion is not useful for this configuration. The MPEG-2 **Transport Stream** packets are stored in a FIFO buffer. The scheduler requests the data output of the FIFO buffer in order to create an output stream that completely fills the available **Data Slice** resources. If the data within the FIFO buffer is not sufficient to fill a **BBFrame** completely, the transmitter shall use **BBFrame** Stuffing as described in clause 7.3.2.

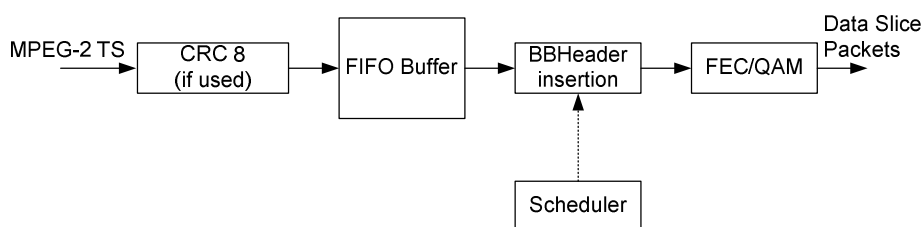


Figure 16: Block Diagram for Data Slice Type 1 Scheduler

##### 7.7.1.2 Data Slice Type 2

The transmission within the **Data Slice** Type 2 can follow the same principle as for **Data Slice** Type 1. However, the stuffing does not take place within the **BBFrames**, but uses Stuffing **Data Slice** Packets. The application of Null Packet deletion has not been considered for simplicity reasons.

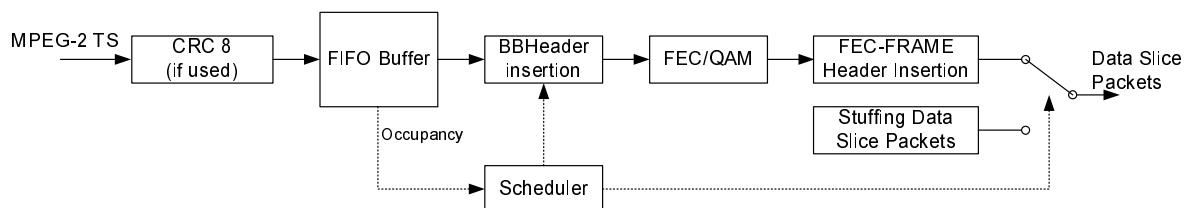


Figure 17: Block Diagram for Data Slice Type 2 Scheduler

The scheduler checks the occupancy of the FIFO buffer. If the FIFO buffer contains sufficient data to fill one **BBFrame** completely, this data is transmitted. Furthermore, if the FIFO buffer contains sufficient information to fill more than one **BBFrame**, it can also use the option to transmit one **FECFRAME** Header for two **BBFrames**.

If the FIFO does not contain enough data to fill one **BBFrame**, but the transmitter has to transmit data within the **Data Slice**, the transmitter sends a Stuffing **Data Slice Packet** instead.

NOTE: When Null Packet Deletion is used and the input **Transport Stream** contains high number of Stuffing Packets, the Scheduler may also be forced to transmit **BBFrames** not completely filled with data in order to avoid buffer under runs in the receiver.

## 7.7.2 Multiple PLP

The transmission of multiple **PLPs** within one **Data Slice** is only possible using the **Data Slice Type 2**, because **Data Slice Type 1** does not allow for the transmission of multiple **PLPs**. Figure 18 shows the block diagram of a possible scheduling algorithm. The associated pseudo code for the scheduling algorithm is listed in figure 19.

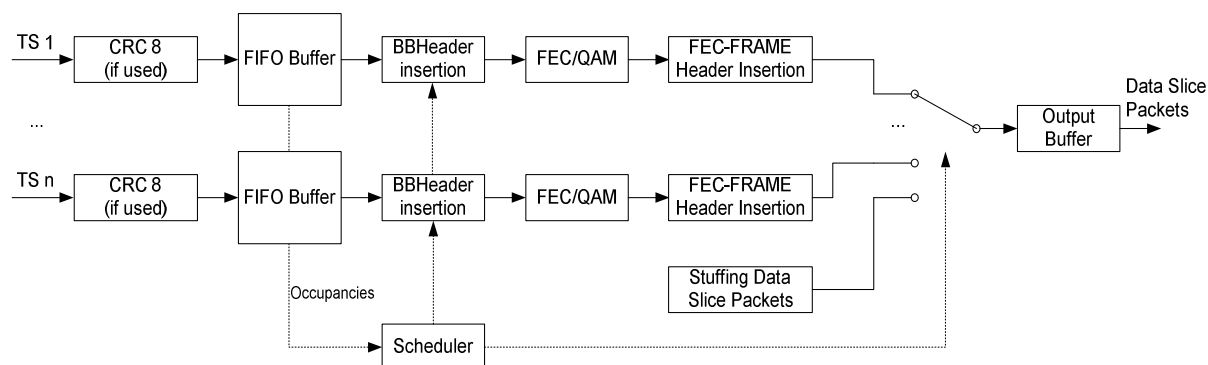


Figure 18: Block Diagram for Multiple PLP per Data Slice Scheduling

```

while (true) {
    // Loop forever
    for (inputFIFO = 1; inputFIFO <= numberOfPLPs; inputFIFO++) {
        if (hasBBFrame(inputFIFO)) { // Transmit one BBFRAME of
            // this PLP if available in FIFO
            transmitBBFrame(inputFIFO);
        }
    }
    if (outputBufferEmpty()) // Transmit Stuffing if buffer empty
        transmitStuffing();
    while (!outputBufferEmpty()); // Wait until data transmitted
}

```

Figure 19: C-Pseudo code for multiple PLP per Data Slice scheduling

Firstly, it is assumed that the execution time of the code is zero. The scheduling algorithm checks the FIFO buffers of all input **PLPs** each loop. If the corresponding FIFO buffer contains at least one complete **BBFrame**, one **BBFrame** of this **PLP** is transmitted, i.e. copied to the output buffer that ensures the synchronous transmission at the required cell rate within the **Data Slice**. If no **BBFrame** was available for all **PLPs**, a Stuffing **Data Slice Packet** is transmitted instead. Finally, the loop waits until the output buffer is empty.

## 7.8 Layer-2 signalling

### 7.8.0 Introduction

Layer 2 signalling is provided differently for MPEG-2 **Transport Streams** and Generic Streams.

DVB-C2 introduces the layer-1 concept of the "C2 System" (see clause 8 in ETSI EN 302 769 [i.1]). The mapping of **Transport Streams** to **PLPs** and the C2-system is signalled in the C2\_delivery\_system\_descriptor transmitted in the NIT.

Furthermore for the transport of **Transport Stream (TS)** two different delivery system descriptor have been defined for the application delivery of the TS in a single **PLP** and for the delivery of the TS in multiple **PLPs** (so called **PLP bundling**).



### 7.8.1 Transport Streams transported in a single PLP

For the Transport-Stream case a single new PSI/SI signalling element is used: the C2 delivery system descriptor (C2dsd). It is a mandatory element of the Network Information Table (NIT) that is provided for all **Transport Streams** of a **C2 system**. For each **Transport Stream** a single instance of the descriptor will be provided.

The C2dsd maps a **Transport Stream** being signalled with the NIT, and heading the TS descriptor loop, to the corresponding **Data Slice** (data\_slice\_id) and the **PLP** (PLP\_id) that carries this **Transport Stream** within the **C2 system**. This mapping is handled by the C2dsd which is illustrated in Table 15. Each TS carried requires a separate instance of the C2 delivery-system descriptor, i.e. another loop cycle of the NIT's TS loop.

Table 15: C2 delivery system descriptor

Syntax	Number of bits	Identifier
C2_delivery_system_descriptor() {		
descriptor_tag	8	uimsbf
descriptor_length	8	uimsbf
descriptor_tag_extension	8	uimsbf
plp_id	8	uimsbf
data_slice_id	8	uimsbf
C2_tuning_frequency	32	bslbf
C2_tuning_frequency_type	2	uimsbf
active_OFDM_symbol_duration	3	bslbf
guard_interval	3	bslbf
}		

A parameter 'C2\_tuning\_frequency\_type' is introduced in the C2dsd to signal the different tuning possibilities of the C2\_tuning\_frequency within the C2 specific framing structure and give information about its underlying **Data Slices** and broadband **notches**. The C2\_tuning\_frequency\_type field informs the receiver whether the frequency signalled in C2\_tuning\_frequency is the final **Data Slice** specific tuning position or whether the receiver needs to retune after initial evaluation of the preamble with the embedded L1 signalling part 2. This enables the receiver to skip unnecessary re-tunes and shorten demodulation time. It is particularly helpful for Video on Demand (VoD) systems where the final **Data Slice** specific tuning frequency is decided dynamically.

The following three cases exist:

#### Case 1: '8 MHz only' C2 system

In case the **C2 system** covers the bandwidth of only a single **L1 block** (i.e. 3 409 subcarriers, including edge pilot), the both the C2\_System centre Frequency (C2SF) and the **Data Slice** specific frequency, the **Data Slice** Tuning Frequency (DSTF) are equivalent (see figure 20). In this case the C2\_tuning\_frequency\_type should be set to 00, i.e. the C2\_tuning\_frequency is already the final **Data Slice** specific tuning position.

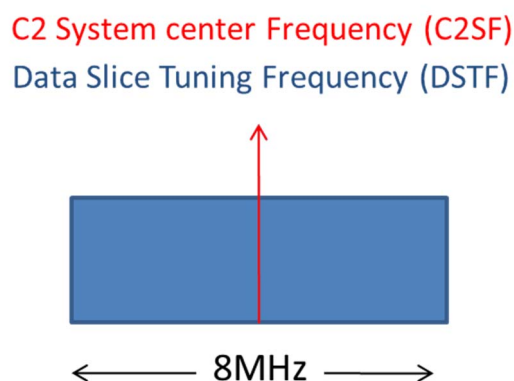
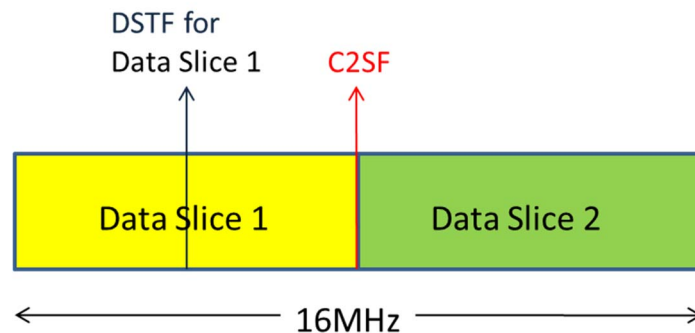


Figure 20: Tuning position of an '8 MHz only' C2 system

### Case 2: C2 system with more than 8 MHz bandwidth

For **C2 systems** covering more than 3 409 subcarriers (i.e. greater than the bandwidth of one L1 signalling block and including multiple **Data Slices**), the **C2 system** centre Frequency (C2SF) is usually at the centre of the band and is different from the **Data Slice** tuning frequency (DSTF). Figure 21 shows a possible **Data Slice** allocation example.



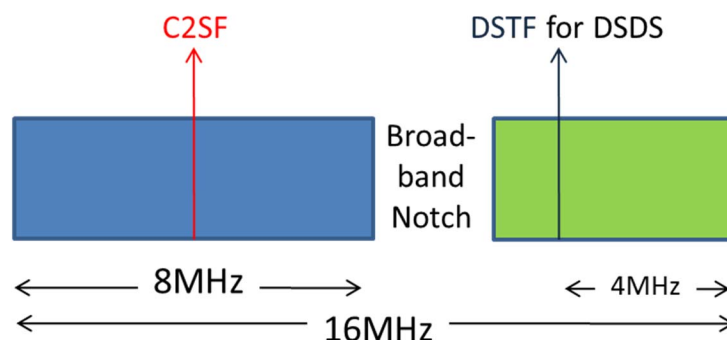
**Figure 21: C2 systems above 8MHz with the different tuning position possibilities**

For rather static C2 framing structures above 8 MHz it is beneficial if the C2\_tuning\_frequency is the DSTF, since this allows the receiver to avoid a 2-step tuning process and therefore reduces the demodulation time to the desired **PLP** inside a **Data Slice**. In this case the C2\_tuning\_frequency\_type should be set to 00, indicating that the C2\_tuning\_frequency is the final **Data Slice** specific tuning position.

There are also cases where it is beneficial to signal the **C2 system** centre frequency in the C2 dsd, even though this can sometimes require a 2-step tuning approach to tune to the final **Data Slice** specific position and has an associated demodulation time penalty. One example is a C2 head end that is not able to update the L1 signalling and the C2 dsd in a synchronous way. In this case the **C2 system** centre frequency should be used and the C2\_tuning\_frequency\_type field should be set accordingly to 01. The same setting should be used for **C2 systems** that have a very dynamic change in behaviour of the **Data Slice** and **PLP** allocation, in order to prevent too many updates of the NIT.

### Case 3: Dependent Static Data Slices

Another case occurs if **Dependent Static Data Slices** (DSDS) are used. Since DSDS are standalone **Data Slices** below 3 408 subcarriers it cannot be guaranteed that the receiver is able to decode the L1 signalling at its final tuning position. Therefore the receiver shall first tune to the signalled initial C2\_tuning\_frequency where a complete preamble is transmitted. This frequency will usually be the **C2 system** (centre) frequency, but can be any tuning position where the receiver can reliably decode the L1 signal. The receiver needs to evaluate the L1 signalling in the preamble in order to determine additional parameters (particularly notch positions) as well as the final tuning frequency of the (dependent) Static **Data Slice**. For this case the C2\_tuning\_frequency\_type should be set to 10.



**Figure 22: Tuning frequency positions for Dependent Static Data Slices**

Another new parameter, which was not present in the corresponding DVB-C descriptor, is the active **OFDM symbol** duration field (see clause 6.4.4.1 in ETSI EN 300 468 [i.9]).

All **PLP**-related information (modulation/constellation, code rate, start address, etc.) is provided via and derived from layer-1 signalling, see clause 8 in ETSI EN 302 769 [i.1].

In case the 'PSI/SI reprocessing' field is set to '1' in layer-1 signalling, it means that the **Transport Stream** in corresponding **PLP** is not reprocessed with valid PSI/SI information and just forwarded from another medium. In this case, layer-2 signalling is not valid anymore and all the information in the NIT should be ignored by a receiver. Instead, valid transport\_stream\_id and original\_network\_id will be delivered by layer-1 signalling in this case.

## 7.8.2 Transport Streams transported in bundled PLPs (PLP bundling)

Table 16 gives the definition of the C2 bundle delivery descriptor. The only difference to the mechanisms described in clause 7.8.1 is that the relevant **Transport Stream** is not transported in a single **PLP** but in multiple **PLPs** as defined in the loop in table 16.

**Table 16: C2 bundle delivery system descriptor**

Syntax	Number of bits	Identifier
C2_bundle_delivery_system_descriptor() {		
descriptor_tag	8	Uimsbf
descriptor_length	8	Uimsbf
descriptor_tag_extension	8	Uimsbf
for(i=0;i<N;i++){		
plp_id	8	Uimsbf
data_slice_id	8	Uimsbf
C2_System_tuning_frequency	32	Bslbf
C2_System_tuning_frequency_type	2	Uimsbf
active_OFDM_symbol_duration	3	Bslbf
guard_interval	3	Bslbf
master_channel	1	uimsbf
reserved_0_future_use	7	bslbf
}		
}		

The detailed description of the C2 bundle delivery system descriptor is given in "Specification for Service Information (SI) in the DVB-System", ETSI EN 300 468 [i.9].

## 7.8.3 Generic Streams

Generic Streams might occur in one of the following three variants:

- Generic Fixed Packet size Stream (GFPS).
- Generic Continuous Stream (GCS), also covering the case of a Generic Variable Packet size Stream.
- Generic Stream Encapsulated (GSE) according to ETSI EN 300 421 [i.6].

For cases a) and b) there is no layer 2 signalling definition available, for c) refer to clause 8 in ETSI TS 102 771 [i.8].

# 8 Modulator

## 8.1 Preamble Generation

### 8.1.0 Introduction

Figure 23 depicts the block diagram of the preamble generation. Firstly, the preamble data is encoded by means of the forward error correction and the QAM mapping. This is followed by the time interleaving. Then, the **preamble header** is inserted. The preamble data and the Preamble Header are cyclically repeated until all available payload OFDM sub-carriers within the preamble are used. Afterwards, the data is interleaved in the frequency interleaver. Finally, the preamble data and the preamble pilots are mapped onto the specific OFDM sub-carriers.

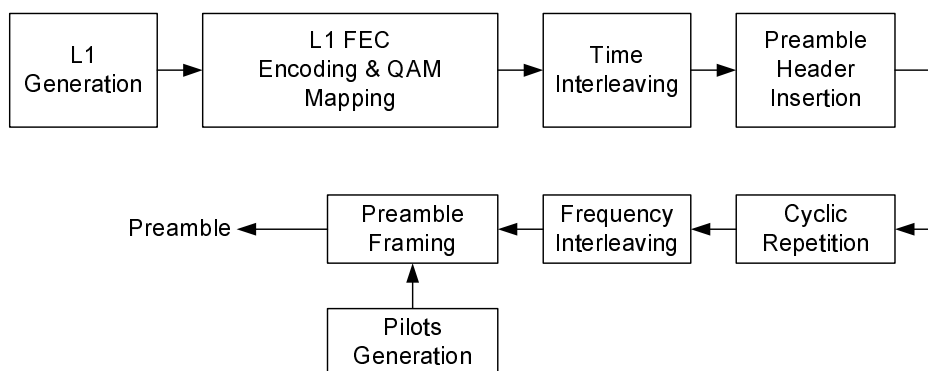


Figure 23: DVB-C2 preamble generation block diagram

## 8.1.1 Preamble Payload Data Processing

### 8.1.1.1 Preamble Time Interleaving

Time interleaving (TI) for the preamble operates at constellation-mapped **Layer 1 (L1) part 2** data level (see clause 8.5 of ETSI EN 302 769 [i.1]). A L1 TI block is the set of **OFDM cells** within which L1 time interleaving is performed; there is no time interleaving between L1 TI blocks within one or multiple preamble(s). No time interleaving is allowed between **Data Slice** and preamble. The L1 header is not included in L1 time interleaving.

The total number of cells within the L1 TI block is decided by the number of total **L1-part2** data bits and constellation size. Each L1 TI block contains integer number of shortened/punctured FEC blocks of **L1-part2** data. Each L1 TI block should contain all cells for encoded and modulated **L1-part2** data; the cells for **L1-part2** data cannot be split into different TI blocks.

The L1 TI block may be copied and repeated to fill entire L1 signalling block bandwidth (3 408 carriers). For the last copy of L1 TI block, the part of L1 TI block may not be repeated into L1 signalling block if all carriers are used thus no carrier is left. In this case, the last part of L1 TI block in order is not transmitted (see figure 28 of ETSI EN 302 769 [i.1]).

The address generation equation in clause 8.5 of ETSI EN 302 769 [i.1] assumes ( $C$  columns  $\times$   $R$  rows) rectangular memory size for L1 time interleaving. Each cell in the memory maps 'one-to-one' to the **OFDM cell** of the L1 TI block in the preamble. Preamble pilots and reserved dummy carriers for PAPR reduction should not be included in address generation and interleaving process.

The trade-off between time interleaving gain and preamble overhead is made through four L1 TI modes (see table 17 of ETSI EN 302 769 [i.1]). The overheads of "no time interleaving" and "best-fit" mode are the same as well as minimum. If "best-fit" mode is chosen, the L1 TI depth is automatically set to the number of OFDM symbols required for "no time interleaving" case. However, if the amount of **L1-part2** data is relatively small and the number of OFDM symbols for preamble is just one or two for example, the robustness of **L1-part2** data may be greatly damaged by a time-domain impulsive noise channel environment. To overcome those cases, the L1 TI depth can be expanded and explicitly set to "4 or 8 OFDM symbols" which significantly improves robustness of **L1-part2** data transmission.

Note that L1 TI mode of "4 or 8 OFDM symbols" depth should be carefully chosen only for the case when the number of OFDM symbols for preamble in the "no time interleaving" case is equal or less than 4 or 8 respectively. Otherwise, it causes the L1 time interleaver to work incorrectly.

EXAMPLE: If preamble is made up of 5 OFDM symbols with no time interleaving, either "best-fit" (L1\_TI\_MODE=01) or "8 OFDM symbols" (L1\_TI\_MODE=11) can be chosen for L1 TI mode. But "4 OFDM symbols" (L1\_TI\_MODE=10) cannot be used.

### 8.1.1.2 Addition of Preamble Header

In addition to the LDPC encoded preamble data a **preamble header** is added to the **preamble symbol**. This **preamble header** does not change within a C2 preamble if it consists of multiple OFDM symbols. It carries information that is required for the decoding of the L1 signalling data. Details are given in clause 8.2 of ETSI EN 302 769 [i.1].

### 8.1.1.3 Cyclic Repetition

After the addition of the **preamble header** there may be still some unused **OFDM cells** within the preamble. Therefore, the remaining parts are filled with a cyclic repetition of the signalling data (including the **preamble header**). Details are given in clause 8.4.1 of ETSI EN 302 769 [i.1].

### 8.1.1.4 Preamble Frequency Interleaving

The purpose of the preamble frequency interleaver is the separation of neighbouring **data cells** and to avoid error bursts caused by narrow band interferers or frequency selectivity. Therefore, the same frequency interleaver as for the **Data Slices** is used **Data Slices** (see clause 9.4.5), which works on the  $N_{L1}=2\ 840$  **data cells** of each **L1 block**.

The interleaved vector  $A_{l_p}^P = (a_{l_p,0}^P, a_{l_p,1}^P, \dots, a_{l_p,N_{L1}-1}^P)$  is defined by:

$$a_{l_p,q}^P = x_{l_p,H_0(q)} \text{ for even symbols in the preamble } (l_p \bmod 2 = 0) \text{ for } q = 0, 1, \dots, N_{L1} - 1$$

$$a_{l_p,q}^P = x_{l_p,H_1(q)} \text{ for odd symbols in the preamble } (l_p \bmod 2 = 1) \text{ for } q = 0, 1, \dots, N_{L1} - 1$$

with  $N_{L1}=2\ 840$ .

In contrast to the frequency interleaver used for the **Data Slices**, the frequency interleaver for the preamble always uses a fixed number of interleaved cells. Furthermore, the frequency interleaver does not interleave between different OFDM symbols, but only within OFDM symbols.

The frequency interleaver uses the odd-only approach. The interleaving for odd and even OFDM symbols is different, which will give additional robustness if the preamble time interleaver is used. The first **preamble symbol** is considered as the first symbol of the frame i.e. with a symbol number of 0. The functioning of the frequency interleaver starting from symbol 0 is described in clause 8.5.3.

The frequency interleaver address generation for the following **data symbols** shall be reset at the end of the **preamble symbols**.

## 8.1.2 Preamble Pilot Generation

### 8.1.2.0 Introduction to the Preamble Pilot Generation

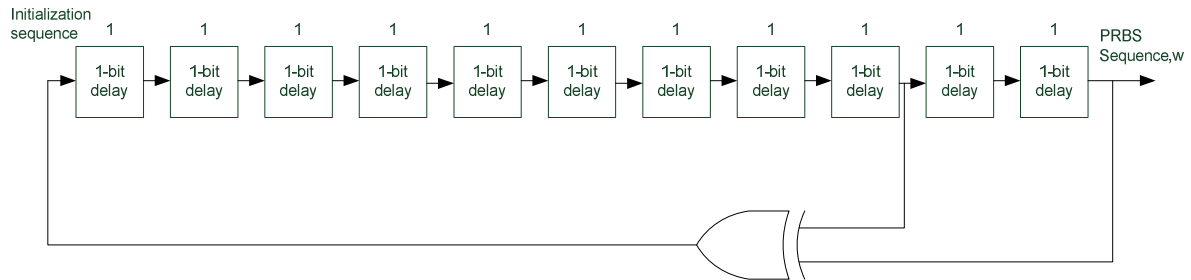
The DVB-C2 preamble repeats itself cyclically in the frequency domain. As repetitions within the frequency domain lead to high peak-to-average power ratios it is desirable to avoid this direct repetition. For the DVB-C2 preamble this aim is reached by means of scrambling the preamble with two scrambling sequences. The data scrambling sequence scrambles the data and the pilots. Furthermore, an additional pilot scrambling sequence is employed that works on the pilots only. It is required as the data scrambling sequence in addition to the preamble pilot separation would lead to a repetition of the pilot modulation every 27 MHz (in 8 MHz operation). Using the additional sequence, the modulation should be unique within the complete cable bandwidth.

#### 8.1.2.1 Data Scrambling Sequence

Figure 24 shows the block diagram for the generation of the data scrambling sequence. The sequence is obtained by using a linear feedback shift register with the generator polynomial:

$$X^{11} + X^2 + 1.$$

The output of the generator is the value of the most right shift register in figure 24. Before the calculation all shift registers have to be initialized with '1' s.



**Figure 24: Data Scrambling Sequence**

The first 100 output values  $w_0, w_1, \dots, w_{99}$  are:

1, 1, 1, 1, 1, 1, 1, 1, 1, 1, 1, 0, 0, 0, 0, 0, 0, 0, 0, 0, 1, 1, 0, 0, 0, 0, 0, 0, 0, 1, 1, 1, 1, 0, 0, 0, 0, 0, 1, 1, 0, 0, 1, 1, 0, 0, 0, 1, 1, 1, 1, 1, 1, 0, 1, 1, 0, 0, 0, 0, 0, 0, 1, 0, 1, 1, 1, 0, 0, 0, 0, 1, 0, 0, 1, 0, 1, 1, 0, 0, 1, 0, 1, 1, 0, 0, 1, 1, 1, 1, 0, 0, 1, 1, 1, 1, 0

In order to obtain the output  $w_k$  of the sequence,  $k$  shift operations have to be performed. However, due to its polynomial the generator is a maximum-length LFSR and the sequence repeats itself every  $2^{11} - 1 = 2047$  outputs:

$$w_k = w_{(k \bmod 2047)}.$$

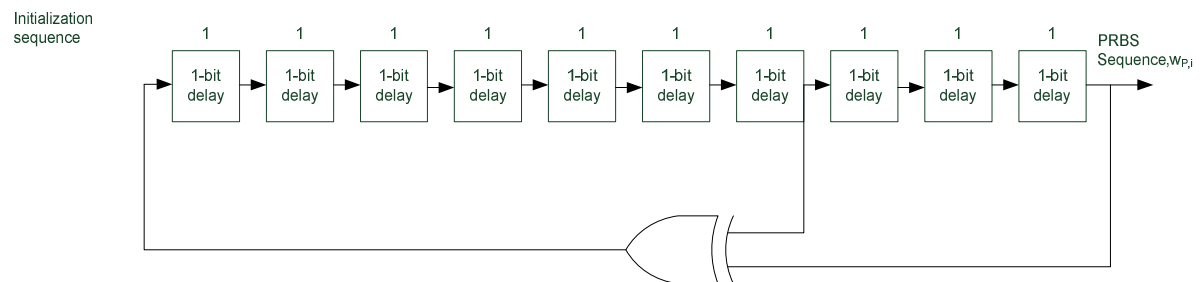
Hence, it is also possible to pre-calculate the sequence and use look-up tables instead of real-time calculation.

### 8.1.2.2 Pilot Scrambling Sequence

Figure 25 depicts the block diagram of the pilot scrambling sequence generator. In contrast to the data scrambling sequence it employs only 10 shift registers. The polynomial of this linear feedback shift register is:

$$X^{11} + X^2 + 1.$$

Similar to the data sequence it is also initialized with all '1's at the beginning.



**Figure 25: Pilot Scrambling Sequence**

In order to obtain the output  $w'_i$   $i$  shift operations have to be performed. The first 100 output values  $w'_0, w'_1, \dots, w'_{99}$  are:

1, 1, 1, 1, 1, 1, 1, 1, 1, 1, 0, 0, 0, 0, 0, 0, 0, 0, 1, 1, 1, 0, 0, 0, 0, 1, 1, 1, 1, 1, 0, 1, 1, 1, 0, 0, 0, 1, 0, 0, 1, 1, 1, 1, 1, 0, 0, 0, 1, 1, 0, 0, 1, 1, 1, 1, 0, 1, 0, 1, 0, 1, 0, 0, 1, 0, 1, 1, 0, 0, 1, 0, 0, 1, 0, 0, 1, 0, 0, 0, 0, 0, 0, 0, 0, 1, 0, 0, 0, 0, 0, 0, 1, 0, 0, 1, 0, 0, 1, 0, 0, 0

Due to the chosen polynomial, the sequence repeats itself every  $2^{10} - 1 = 1023$  outputs:

$$w'_i = w'_{(i \bmod 1023)}.$$

### 8.1.2.3 Impact of long "0" or "1" sequences in the Preamble on the PAPR

Generally the presence of long "0" or "1" sequences within the Preamble leads to a bias on the OFDM subcarriers. This causes a high Peak to Average Power Ratio (PAPR) within the resulting time domain signal, i.e. the presence of very high signal peaks. Therefore, any repetition of sequences is avoided by scrambling of the OFDM subcarriers in DVB-C2.

In contrast to the payload data, DVB-C2 does not apply any scrambling of the preamble data before the QAM mapping. Instead, the scrambling is applied on the individual OFDM subcarriers by means of the Data Scrambling Sequence. For the Preamble Pilots the Pilot Scrambling Sequence is used instead. Due to their special structure of the scrambling sequences no repetition within the frequency domain will occur for the complete DVB-C2 bandwidth. Hence, the occurrence of high peaks within the time domain signal is minimized.

### 8.1.2.4 Preamble Pilot Modulation Sequence

The modulation of the pilots is obtained by combining the pilot and the data scrambling sequence:

$$w_k^P = w_k \oplus w_i' \text{ with } i = (k \bmod K_{L1}) / D_p$$

while  $k$  is the absolute OFDM sub-carrier index,  $K_{L1}=3\,408$  is the number of OFDM sub-carriers per **L1 block**, and  $D_p=6$  is the preamble pilot spacing.

The value  $i$  is the required output index of the pilot scrambling sequence. Due to the modulo operation it is initialized every **L1 block**, and the division ensures that the pilot scrambling sequence gives an output value for each pilot, but not for each OFDM sub-carrier. The calculation of  $i$  without remainder is thus only possible if  $k$  is multiple of 6. However, this is ensured as all pilot positions are located on values of  $k$ , which are dividable by 6.

As an example the values of  $w_k^P$  for **L1 block** 100 (starts at 760,714 MHz in 8 MHz operation) are calculated. Firstly, the number of OFDM sub-carrier of this **L1 block**  $k$  is calculated:

$$k = K_{L1} \cdot 100 = 3\,408 \cdot 100 = 340\,800$$

The following sequence is obtained:

$$\begin{aligned} w_{340\,800}^P &= w_{340\,800} \oplus w'_{(340\,800 \bmod 3\,408) / 6} \\ &= w_{340\,800} \oplus w'_0 \end{aligned}$$

Considering the cyclic repetition of the sequences, the following result is obtained:

$$\begin{aligned} w_{340\,800}^P &= w_{(340\,800 \bmod 2\,047)} \oplus w'_0 \\ &= w_{998} \oplus w'_0 \\ &= 0 \oplus 1 = 1 \end{aligned}$$

Similar calculations can be made for the other pilot positions. Hence:

$$\begin{aligned} w_{340\,806}^P &= w_{(340\,806 \bmod 2\,047)} \oplus w'_1 = w_{1\,004} \oplus w'_1 = 0 \oplus 1 = 1 \\ w_{340\,812}^P &= w_{1\,010} \oplus w'_2 = 1 \oplus 1 = 0, \quad w_{340\,818}^P = 1 \oplus 1 = 0, \quad w_{340\,824}^P = 1 \oplus 1 = 0 \end{aligned}$$

The final pilot modulation  $r_k$  is then generated by differentially modulating the sequence  $w_k^P$ . However, in order to obtain an absolute reference, the modulation of the first pilot within each **L1 block** is absolute. Thus, the pilot modulation is obtained by:

$$r_k = \begin{cases} w_k^P & \text{if } k \bmod K_{L1} = 0 \\ r_{k-6} \oplus w_k^P & \text{otherwise} \end{cases}$$

while the modulo term ensures the absolute modulation of the first pilot within each **L1 block**. Coming back to the example, the values  $r_k$  are:

$$r_{340\ 800} = w_{340\ 800}^P = 1 \text{ (absolute modulation),}$$

$$r_{340\ 806} = r_{340\ 800} \oplus w_{340\ 806}^P = 1 \oplus 1 = 0,$$

$$r_{340\ 812} = r_{340\ 806} \oplus w_{340\ 812}^P = 0 \oplus 0 = 0, \dots$$

while  $k$  is again only defined if it divisible by 6, as only these positions are preamble pilots.

It has to be mentioned that the same pilot modulation sequence  $r_k$  is used for the pilots within the payload OFDM symbols, too.

## 8.1.3 Mapping of the Preamble Pilots and Data

### 8.1.3.1 Mapping of the Pilots

The pilots are mapped onto the preamble by using the equation:

$$\begin{aligned} \operatorname{Re}\{c_{m,l_p,k}^P\} &= A_{PP} \cdot 2(1/2 - r_k) \\ \operatorname{Im}\{c_{m,l_p,k}^P\} &= 0 \end{aligned}$$

while  $c_{c,l_p,k}^P$  is the complex QAM cell of frame  $m$ , preamble **OFDM symbol**  $l_p$  and absolute OFDM sub-carrier  $k$ ,  $A_{PP}=1$  is the amplitude of the preamble pilots, and  $r_k$  is the preamble pilot modulation sequence. Please note that this equation only holds for preamble pilot positions, i.e. where  $k$  is multiple of 6.

The modulation of the pilots is thus BPSK based, with the mapping  $r_k = 1 \Rightarrow c_{m,l_p,k}^P = -1$  and  $r_k = 0 \Rightarrow c_{m,l_p,k}^P = 1$ .

### 8.1.3.2 Mapping and Scrambling of the Preamble Data

The preamble data repeats itself cyclically within the frequency domain, i.e. every  $K_{L1}$  OFDM sub-carriers (7,61 MHz in 8 MHz operation). Additionally, the data is scrambled using the data scrambling sequence  $w_k$ . This is described by:

$$\begin{aligned} \operatorname{Re}\{c_{m,l_p,k}^P\} &= \operatorname{Re}\{a_{l_p,q}^P\} \cdot (-1)^{w_k} \\ \operatorname{Im}\{c_{m,l_p,k}^P\} &= \operatorname{Im}\{a_{l_p,q}^P\} \cdot (-1)^{w_k} \end{aligned} \quad \text{with } q = (k \bmod K_{L1}) - \lceil (k \bmod K_{L1}) / 6 \rceil$$

while  $w_k$  is the  $k$ -th value of the data scrambling sequence,  $a_{l_p,q}^P$  is the  $q$ -th output of the preamble frequency interleaver for the preamble **OFDM symbol**  $l_p$ , and  $c_{c,l_p,k}^P$  is the complex QAM cell of frame  $m$ , preamble **OFDM symbol**  $l_p$  and absolute OFDM sub-carrier  $k$ . Please note that this equation is not valid for pilot positions, i.e.  $k$  is divisible by 6.

The right hand formula calculates the value of  $q$  out of the absolute OFDM sub-carrier index  $k$ . In an example the number of  $q$  for a given  $k$  can be calculated as follows:

$$k \rightarrow q: 0 \rightarrow \text{Pilot}, 1 \rightarrow 0, 2 \rightarrow 1, 3 \rightarrow 2, 4 \rightarrow 3, 5 \rightarrow 4, 6 \rightarrow \text{Pilot}, 7 \rightarrow 5, 8 \rightarrow 6, \dots$$

By means of the modulo part of the equation the mapping of the data gets cyclical every **L1 block**.



## 8.2 Pilots (Scattered-, Continual- pilots)

### 8.2.1 Purpose of pilot insertion

Pilots are inserted for several reasons; the scattered pilots are inserted mainly for channel estimation and subsequently equalization. The pilot patterns for DVB-C2 are chosen so that channel estimation potentially can be done without any time-axis interpolation, however to get maximum performance from the given pilot pattern the receiver should exploit the fact that there is a strong correlation among pilots over long time interval.

The edge pilots are also inserted mainly to assist the channel estimation. These help to improve the frequency axis interpolation around the upper and lower extent of the active spectrum. Edge carriers are continual, hence they can be treated as continual pilots to improve some of parameter estimations that are derived from continual pilots.

The continual pilots can be used for Common-Phase-Error correction, coarse and fine carrier offset and sample rate offset detection.

### 8.2.2 Pilot locations

Regarding pilot locations, particular care should be taken to the following points:

- Preamble symbols do not have any of scattered, continual or edge pilots, instead they have preamble pilot. The preamble pilot locations are a superset of all scattered-, continual- and edge-pilots locations.
- The scattered pilot pattern begins on the first **data symbol** of a **C2 frame**.
- The carrier index used to define scattered pilot locations has its origin at the 0 RF carrier. (Not the left most carrier as DVB-T/H/T2 standards.)
- The carrier index used to define continual pilot locations also has its origin at the 0 RF carrier. The continual pilot locations are repeated every 3 408 carriers, and the 3 408 carrier block is aligned with L1 signalling block. This provides receivers a method to detect L1 signalling block location.

### 8.2.3 Number of pilot cells

Due to the unique definition of the pilot locations across the whole cable medium bandwidth the absolute number of pilot cells depend on the C2 bandwidth as well as its frequency location. The number of pilot cells is furthermore different for the 2 possible guard interval fractions as they result in different scattered pilot densities (scattered pilots every 48 subcarriers for  $GI = 1/64$ , accordingly 96 subcarrier for  $GI = 1/128$ ).

The calculation of the number of pilot cells should also include that there are some scattered pilot positions that coincide with continual pilot positions. Depending on the chosen guard interval fraction these are:

- $GI = 1/64$ : 4 coinciding scattered and continuous pilot locations within a L1 block bandwidth (3 408 subcarriers).
- $GI = 1/128$ : 2 coinciding scattered and continuous pilot locations within a L1 block bandwidth.

For the L1 block bandwidth these numbers are the same for all 4 possible scattered pilot phases.

### 8.2.4 Pilot boosting

All scattered-, continual- and edge-pilots of **data symbols** have the same boosting value of  $7/3$ . The amplitude of preamble pilots ( $A_{pp}$ ) is fixed to  $A_{pp}=6/5$  in case of  $1/128$  Guard Interval and fixed to  $A_{pp}=4/3$  in case of  $1/64$  Guard Interval. The given boosting values ensure that the average energy of the preamble and the **data symbols** is the same.

It is important that the receiver signal processing chain allows adequate headroom to cope with such boosted pilots.

### 8.2.5 Use of reference sequence

All pilots (scattered-, continual- and edge-pilots) are modulated with the same reference sequence as the preambles. The reference sequence is defined for all carrier indices  $k$ , and has its origin at 0 RF. (Not left most carrier as DVB-T/H/T2 system.)

## 8.3 PAPR and Possible Implementation

### 8.3.0 Introduction

When reserved carriers are activated by a relevant L1 signalling part 2, 'RESERVED\_TONES', these reserved carriers defined in clause 9.7 of ETSI EN 302 769 [i.1] can be used for PAPR reduction.

Figure 26 shows the structure of OFDM transmitter with PAPR reduction using reserved carriers. Reserved carriers are allocated according to predetermined carrier locations which are reserved carrier indices. These reserved carriers do not carry any data information and are instead filled with a peak-reduction signal in order to reduce PAPR. Because data and reserved carriers are allocated in disjoint subsets of subcarriers, this PAPR reduction algorithm needs no side information at the receiver. After the IFFT, peak cancellation is operated to reduce PAPR by using a predetermined signal. The predetermined signal, or kernel, is generated by the reserved carriers.

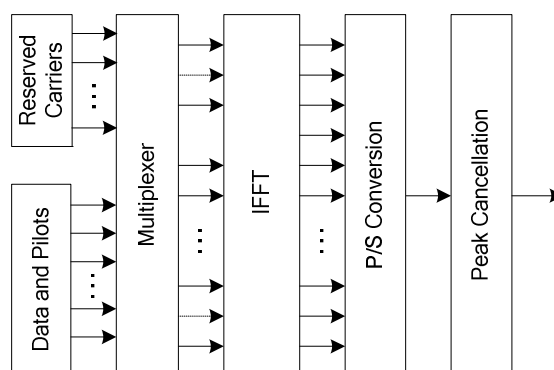


Figure 26: The structure of the OFDM transmitter with PAPR reduction using reserved carriers

### 8.3.1 Reserved carriers

In the **Data symbols** excluding Preamble Symbols, the set of carriers corresponding to carrier indices defined in table 36 of ETSI EN 302 769 [i.1] or their circularly shifted set of carriers shall be reserved depending on the **OFDM symbol** index of the **Data symbol**. Positions of reserved carriers within **notches** shall be excluded from the set of reserved carriers. Reserved carriers are described in more detail in clause 9.7 of ETSI EN 302 769 [i.1].

### 8.3.2 Reference kernel

Signal peaks in the time domain are iteratively cancelled out by a set of impulse-like kernels made using the reserved carriers. A reference kernel signal is defined as:

$$\mathbf{p} = \frac{\sqrt{N_{FFT}}}{N_{TR}} \text{IFFT}(\mathbf{1}_{TR}) = [p_0, p_1, \dots, p_{N_{FFT}-1}]$$

where  $N_{FFT}$  and  $N_{TR}$  indicate the **FFT size** and the number of reserved carriers, respectively. The  $(N_{FFT}, 1)$  vector  $\mathbf{1}_{TR}$  has  $N_{TR}$  elements of ones at the positions corresponding to the reserved carrier indices and has  $(N_{FFT} - N_{TR})$  elements of zeros at the others.

The characteristic of the kernel is similar to an impulse. For example, the primary peak ( $p_0$ ) of the kernel is one and the secondary peaks ( $p_1$  to  $p_{N_{FFT}-1}$ ) of the kernel have a value considerably smaller than  $p_0$ .

### 8.3.3 Algorithm of PAPR reduction using reserved carriers

#### 8.3.3.0 Introduction

Figure 27 shows the detailed block diagram of the peak-cancellation algorithm. The IFFT output ( $\mathbf{x}$ ) is fed into the peak-cancellation block, and the peak position and value of  $\mathbf{x}$  are detected. Then the reference kernel, generated by the reserved carriers corresponding to the current OFDM symbol, is circularly shifted to the peak position, scaled and phase rotated. The resulting kernel is subtracted from  $\mathbf{x}$  and the new PAPR is calculated. The principle is shown in figure 28.

If the PAPR of the resulting signal satisfies the target PAPR level, this signal is transmitted. If not, the cancellation operation is repeated iteratively, until the number of iterations reaches the predetermined maximum iteration number.

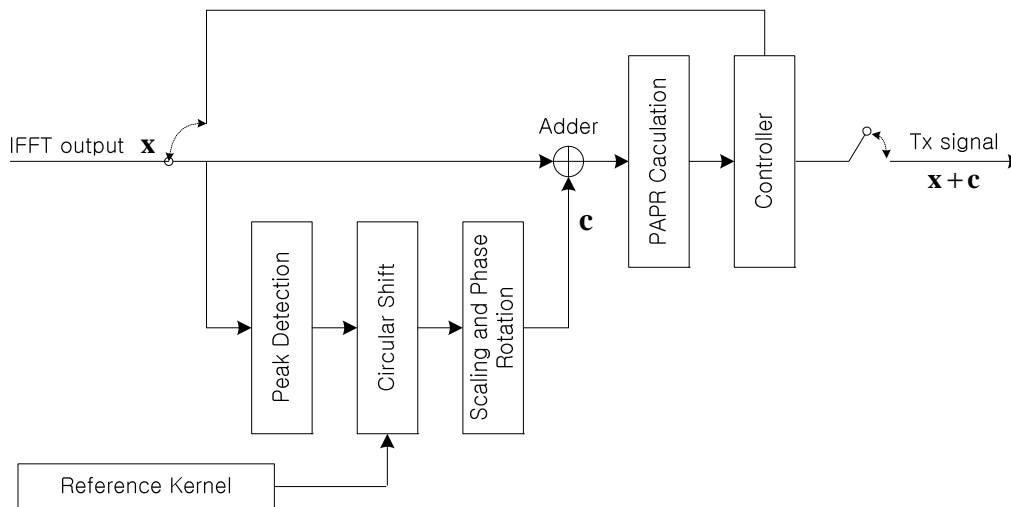


Figure 27: Block diagram of the peak-cancellation algorithm

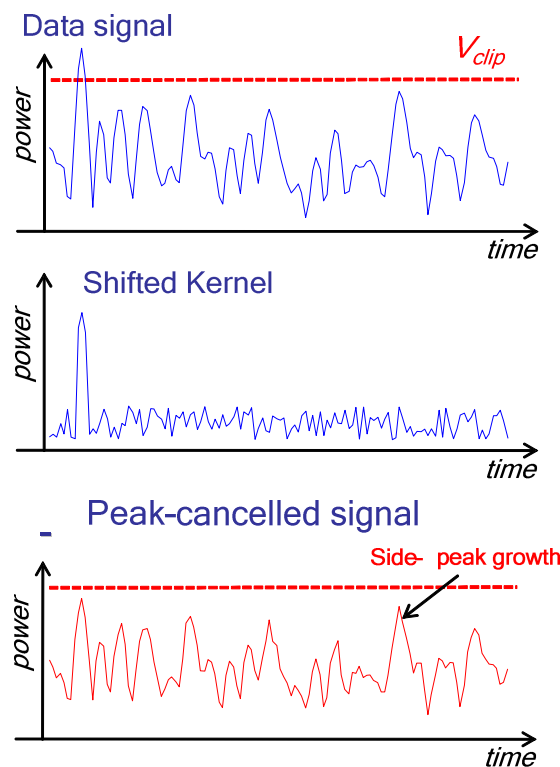


Figure 28: Principle of peak-cancellation algorithm

### 8.3.3.1 PAPR cancellation algorithm

Denote the vector of peak reduction signal by  $\mathbf{c}$ , and the vector of time domain data signal by  $\mathbf{x}$ , then the procedures of the PAPR reduction algorithm are as follows:

#### Initialization:

The initial values for peak reduction signal are set to zeros:

$$\mathbf{c}^{(0)} = [0 \dots 0]^T$$

where  $\mathbf{c}^{(i)}$  means the vector of the peak reduction signal computed in  $i$ -th iteration.

**Iteration:**

- 1)  $i$  starts from 1.
- 2) Find the maximum magnitude of  $(\mathbf{x} + \mathbf{c}^{(i-1)})$ ,  $y_i$  and the corresponding sample index,  $m_i$  in the  $i$ th iteration.

$$\begin{cases} y_i = \max_n |x_n + c_n^{(i-1)}| \\ m_i = \arg \max_n |x_n + c_n^{(i-1)}| \end{cases}, \quad \text{for } n = 0, 1, \dots, N_{FFT} - 1,$$

where  $x_n$  and  $c_n^{(i-1)}$  represent the  $n$ th element of vector  $\mathbf{x}$  and  $\mathbf{c}^{(i-1)}$ , respectively. If  $y_i$  is less than or equal to a desired clipping magnitude level,  $V_{clip}$  then decrease  $i$  by 1 and go to step 5.

- 3) Update the vector of peak reduction signal  $\mathbf{c}^{(i)}$  as

$$\mathbf{c}^{(i)} = \mathbf{c}^{(i-1)} - \alpha_i \mathbf{p}(m_i), \quad \text{where } \alpha_i = \frac{x_{m_i} + c_{m_i}^{(i-1)}}{y_i} (y_i - V_{clip}),$$

where  $\mathbf{p}(m_i)$  denotes the vector circularly shifted by  $m_i$ , of which  $k$ -th element is  $p_k(m_i) = p_{(k-m_i) \bmod N_{FFT}}$

- 4) If  $i$  is less than a maximum allowed number of iterations, increase  $i$  by 1 and return to step 2. Otherwise, go to step 5.
- 5) Terminate the iterations. Transmitted signal,  $\mathbf{x}'$  is obtained by adding the peak reduction signal to the data signal:

$$\mathbf{x}' = \mathbf{x} + \mathbf{c}^{(i)}$$

During each iteration, the peak-cancellation method described above removes only the maximum remaining peak in the time-domain. This method is simple and efficient in terms of peak re-growth control for the following iterations, at the expense of requiring a relatively large number of iterations.

Alternatively, multiple peaks can be removed in a single iteration in order to reduce the computational complexity. The transmitted signal  $\mathbf{x}'$  after the  $j$ th iteration of the simple method is given as:

$$\mathbf{x}' = \mathbf{x} + \alpha_1 \mathbf{p}(m_1) + \alpha_2 \mathbf{p}(m_2) + \dots + \alpha_{j-1} \mathbf{p}(m_{j-1}) + \alpha_j \mathbf{p}(m_j) = \mathbf{x} + \sum_{n=1}^j \alpha_n \mathbf{p}(m_n).$$

Any number of peaks can be cancelled in a single iteration because the kernels can be linearly combined. For example, if two peaks are cancelled per iteration, the total number of iterations is reduced by half and the equation is as follows:

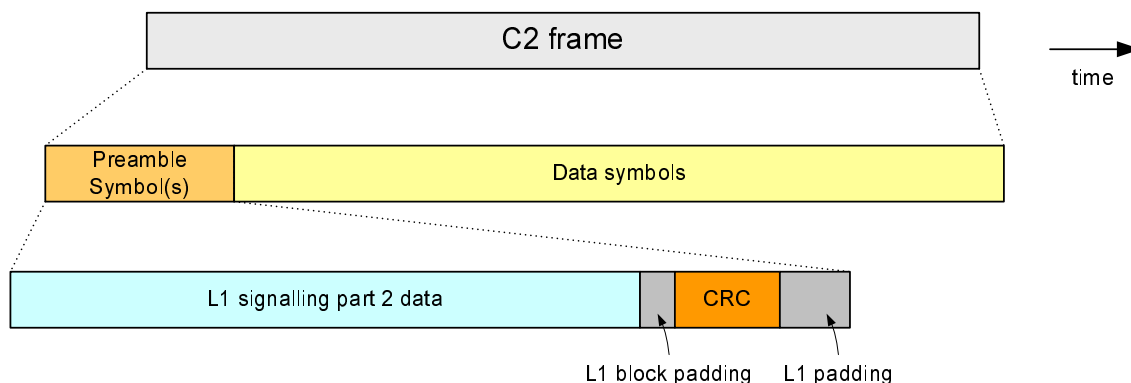
$$\mathbf{x}' = \mathbf{x} + \underbrace{\alpha_1 \mathbf{p}(m_1) + \alpha_2 \mathbf{p}(m_2)}_1 + \dots + \underbrace{\alpha_{j-1} \mathbf{p}(m_{j-1}) + \alpha_j \mathbf{p}(m_j)}_{j/2} = \mathbf{x} + \sum_{n=1}^{j/2} (\alpha_{2n-1} \mathbf{p}(m_{2n-1}) + \alpha_{2n} \mathbf{p}(m_{2n}))$$

Therefore, the number of iterations can be adjusted according to the number of cancelled peaks per iteration.

## 8.4 Signalling (L1-part2, including FEC)

### 8.4.1 Overview

The Layer-1 (L1) part 2 provides all information that receiver needs to access the Layer-2 signalling and the **PLPs** within the current **C2 frame**. The **L1-part2** signalling structure is illustrated in figure 29. All information is conveyed in L1 signalling part 2 data part and it can be changed frame by frame. **L1-part2** data has the **Data Slice** loop and the **notch** band loop and each **Data Slice** loop has the **PLP** loop.



**Figure 29: The L1-part2 signalling structure**

For the **Data Slice** type1 which has only one **PLP** and its code-rate and modulation are constant, all information to access the services is conveyed in **L1-part2** signalling. But for the Data slice type2, some transmission parameters, such as FEC type, modulation and code-rate, are conveyed in a **FECFrame** header which is transmitted through the **Data symbols**.

### 8.4.2 L1 change-indication mechanism

The L1 signalling offers a change-indication mechanism, by means of the **L1\_PART 2\_CHANGE\_COUNTER**-field. It is used to indicate when the configuration (i.e. all fields in L1 signalling part 2 data except for **PLP\_START** and **L1\_PART 2\_CHANGE\_COUNTER** itself) will change. The change is indicated as the number of **C2 frame** after which the change will occur. If this field is set to the value of '0', it means that no change is foreseen during the next 255 **C2 frames**.

The **L1\_PART 2\_CHANGE\_COUNTER** does not take the **PLP\_START** into account because the **PLP\_START** is predictable. If once receiver knows the **PLP\_START**, a **PLP\_START** of next **C2 frame** can be calculated with its coding and modulation information.

### 8.4.3 CRC insertion

The 32-bit CRC is applied only to the L1 signalling part 2 data if the length of L1 signalling part 2 data is a multiple of 2. Otherwise 1-bit L1 block padding is inserted following the L1 signalling part 2 data and CRC is applied to the L1 signalling part 2 data and L1 block padding. The length of the L1 signalling part 2 data can be calculated using **L1\_INFO\_SIZE** in the L1 header.

### 8.4.4 Example of L1 signalling part 2 data

The details of the use of the **L1-part2** signalling are specified in ETSI EN 302 769 [i.1]. This clause gives a full worked example of the construction of the fields of the L1 signalling for the simple case of a single **PLP**, together with comments to explain the chosen meaning in each case. Table 17 gives a worked example of the signalling fields of the **L1-part2** signalling with **Data Slice** type 1.

Table 17: An example of the contents of the L1-part2 signalling fields with Data Slice Type 1

Field	Bits	Contents	Explanation
NETWORK_ID	16	0000000000000000	The network ID is 0x0000
C2_SYSTEM_ID	16	0000000000000000	The <b>C2 system</b> ID is 0x0000
START_FREQUENCY	24	0352E0h	The start frequency of this <b>C2 system</b> is 486,2 MHz (in case of 7,61 MHz BW)
C2_BANDWIDTH	16	0000000010001110	Total number of carriers of this <b>C2 system</b> is 3 409 = 142 x 24 +1 (i.e. including edge pilots)
GUARD_INTERVAL	2	00	Guard interval fraction is 1/128
C2_FRAME_LENGTH	10	0111000000	There are 448 <b>Data symbols</b> in one <b>C2 frame</b>
L1_PART2_CHANGE_COUNTER	8	00000000	There are no changes foreseen in L1 configurable parameters
NUM_DSLSICE	8	00000001	There is 1 <b>Data Slice</b>
NUM_NOTCH	4	0001	There is 1 <b>notch</b> Band
DSLSICE_ID	8	00000000	The <b>Data Slice</b> ID is 0x00
DSLSICE_TUNE_POS	13	0000001000111	The tuning position of this <b>Data Slice</b> is 1 704 <sup>th</sup> carrier frequency of this <b>C2 system</b>
DSLSICE_OFFSET_LEFT	8	10111001	The left edge of this <b>Data Slice</b> is start frequency (apart from tuning position as much as 1 704 carrier spacing)
DSLSICE_OFFSET_RIGHT	8	01000111	The right edge of this <b>Data Slice</b> is apart from start frequency as much as 3 407 carrier spacing
DSLSICE_TI_DEPTH	2	01	TI depth is 4 OFDM Symbols
DSLSICE_TYPE	1	0	This <b>Data Slice</b> is Type 1
DSLSICE_CONST_CONF	1	0	The configuration of this <b>Data Slice</b> can be changed every <b>C2 frame</b>
DSLSICE_LEFT_NOTCH	1	0	There is no left neighboured <b>notch</b> band of this <b>Data Slice</b>
DSLSICE_NUM_PLP	8	00000001	There is 1 <b>PLP</b> in this <b>Data Slice</b>
PLP_ID	8	00000000	The <b>PLP</b> ID is 0x00
PLP_BUNDLED	1	0	This <b>PLP</b> is not bundled with other <b>PLP</b>
PLP_TYPE	2	10	This <b>PLP</b> is normal <b>Data PLP</b>
PLP_PAYLOAD_TYPE	5	00011	This <b>PLP</b> carries a TS
PLP_START	14	00000000000010	First complete <b>XFECFrame</b> starts from 2 <sup>nd</sup> <b>data cell</b> of this <b>Data Slice</b>
PLP_FEC_TYPE	1	1	This <b>PLP</b> uses 64 K LDPC
PLP_MOD	3	011	This <b>PLP</b> uses 256-QAM modulation
PLP_COD	3	100	This <b>PLP</b> uses code rate 5/6
PSI/SI_REPROCESSING	1	1	PSI/SI of this <b>PLPs</b> TS is reprocessed
RESERVED_1	8	00000000	Reserved for future use
RESERVED_2	8	00000000	Reserved for future use
NOTCH_START	13	0000000000010	The <b>notch</b> band starts from 49 <sup>th</sup> carrier of this <b>C2 system</b>
NOTCH_WIDTH	8	00000001	The <b>notch</b> band ends at 71 <sup>th</sup> carrier of this <b>C2 system</b>
RESERVED_3	8	00000000	Reserved for future use
RESERVED_TONE	1	0	There is no reserved tone in this <b>C2 frame</b>
EARLY WARNING SYSTEM (EWS)	1	0	There is no emergency event
C2_VERSION	4	0000	The <b>C2 system</b> ID is 0x0
RESERVED_4	11	00000000000	Reserved for future use

Since all of the bits of the **L1-part2** signalling data are 254-bits, there is a 1-bit L1 block padding following the **L1-part2** signalling data.

Table 18 gives an example of the signalling fields of the **L1-part2** signalling with **Data Slice** type 2.

**Table 18: An example of the contents of the L1-part2 signalling fields with Data Slice Type 2**

Field	Bits	Contents	Explanation
NETWORK_ID	16	0000000000000000	The network ID is 0x0000
C2_SYSTEM_ID	16	0000000000000000	The <b>C2 system</b> ID is 0x0000
START_FREQUENCY	24	0352E0h	The start frequency of this <b>C2 system</b> is 486,2 MHz (in case of 7,61 MHz BW)
C2_BANDWIDTH	16	0000000010001110	Total number of carriers of this <b>C2 system</b> is $3\ 409 = 142 \times 24 + 1$ (i.e. including edge pilots)
GUARD_INTERVAL	2	00	Guard interval fraction is 1/128
C2_FRAME_LENGTH	10	0111000000	There are 448 <b>Data symbols</b> in one <b>C2 frame</b>
L1_PART_2_CHANGE_COUNTER	8	00000000	There are no changes foreseen in L1 configurable parameters
NUM_DSlice	8	00000001	There is 1 <b>Data Slice</b>
NUM_NOTCH	4	0001	There is 1 <b>notch</b> Band
DSlice_ID	8	00000000	The <b>Data Slice</b> ID is 0x00
DSlice_TUNE_POS	13	0000001000111	The tuning position of this <b>Data Slice</b> is 1704 <sup>th</sup> carrier frequency of this <b>C2 system</b>
DSlice_OFFSET_LEFT	8	10111001	The left edge of this <b>Data Slice</b> is start frequency (apart from tuning position as much as 1 704 carrier spacing)
DSlice_OFFSET_RIGHT	8	01000111	The right edge of this <b>Data Slice</b> is apart from start frequency as much as 3 407 carrier spacing
DSlice_TI_DEPTH	2	01	TI depth is 4 OFDM Symbols
DSlice_TYPE	1	1	This <b>Data Slice</b> is Type 2
FEC_HEADER_TYPE	1	0	The type of the <b>FECFrame</b> header of this <b>Data Slice</b> is normal mode
DSlice_CONST_CONF	1	0	The configuration of this <b>Data Slice</b> can be changed every <b>C2 frame</b>
DSlice_LEFT_NOTCH	1	0	There is no left neighboured <b>notch</b> band of this <b>Data Slice</b>
DSlice_NUM_PLP	8	00000001	There is 1 <b>PLP</b> in this <b>Data Slice</b>
PLP_ID	8	00000000	The <b>PLP</b> ID is 0x00
PLP_BUNDLED	1	0	This <b>PLP</b> is not bundled with other <b>PLP</b>
PLP_TYPE	2	10	This <b>PLP</b> is normal Data <b>PLP</b>
PLP_PAYLOAD_TYPE	5	00011	This <b>PLP</b> carries a TS
PSI/SI_REPROCESSING	1	1	PSI/SI of this <b>PLPs</b> TS is reprocessed
RESERVED_1	8	00000000	Reserved for future use
RESERVED_2	8	00000000	Reserved for future use
NOTCH_START	13	0000000000010	The <b>notch</b> band starts from 49 <sup>th</sup> carrier of this <b>C2 system</b>
NOTCH_WIDTH	8	00000001	The <b>notch</b> band ends at 71 <sup>th</sup> carrier of this <b>C2 system</b>
RESERVED_3	8	00000000	Reserved for future use
RESERVED_TONE	1	0	There is no reserved tone in this <b>C2 frame</b>
EARLY WARNING SYSTEM (EWS)	1	0	There is no emergency event
C2 VERSION	4	0000	The <b>C2 system</b> ID is 0x0
RESERVED_4	11	000000000000	Reserved for future use

Since all of the bits of the **L1-part2** signalling data are 234-bits, there is a 1-bit L1 block padding following the **L1-part2** signalling data.

## 8.4.5 FEC for the L1 signalling part 2

### 8.4.5.0 Introduction

This clause gives an overview on protection of L1 signalling part 2 based on BCH and LDPC codes. In particular, this clause briefly introduces the shortening of the BCH information part and puncturing of the LDPC parity part according to the length of signalling information: these are L1-specific operations which are not performed on the **PLP** data.

Basically, the protection mechanism for **L1-part2** in DVB-C2 is the same as that of DVB-T2, except some parameters and steps are slightly changed to be more suitable for DVB-C2.

### 8.4.5.1 Shortening of BCH Information part

The information bits of L1 signalling part 2 are protected by a concatenation of a BCH outer code and an LDPC inner code. The L1 signalling part 2 is first BCH-coded and the BCH-coded word is then further protected by a shortened and punctured 16 K LDPC code. For the protection of the L1 signalling part 2, the BCH code with  $K_{\text{bch}} = 7\,032$  and a 16 K LDPC code with  $K_{\text{ldpc}} = 7\,200$  defined in clause 6.1 in ETSI EN 302 769 [i.1] are used. The BCH-coded word corresponds to LDPC information bits. That is, the output of the BCH encoder is the input of the LDPC encoder. Note that  $K_{\text{ldpc}}$  is the same as  $(K_{\text{bch}} + 168) = 7\,200$  since the number of BCH parity bits in BCH encoding,  $N_{\text{bch\_parity}}$  is fixed as 168 bits.

The length of L1 signalling part 2 is variable, that is, it has different values case by case. Furthermore, the L1 signalling part 2 can be segmented into multiple blocks as described in clause 8.4.1 in ETSI EN 302 769 [i.1]. A segmented L1 signalling part 2 has a length less than or equal to  $N_{\text{L1part2\_max\_per\_Symbol}} = 4\,759$ .  $N_{\text{L1part2\_max\_per\_Symbol}}$  means the maximum number of **L1-part2** information bits for transmitting the coded **L1-part2** through one OFDM symbol.

For the shortening operation, an input parameter  $K_{\text{sig}}$  is first obtained from the length of the segmented L1 signalling part 2 as described in clause 8.4.2 in ETSI EN 302 769 [i.1]. Since  $K_{\text{sig}}$  is always less than  $K_{\text{bch}}$ , the BCH information part is shortened for encoding. More precisely,  $K_{\text{bch}}$  BCH information bits are filled with  $K_{\text{sig}}$  signalling bits and  $(K_{\text{bch}} - K_{\text{sig}})$  zero-padded bits for BCH encoding. Note that the shortening of BCH information may be regarded as the shortening of LDPC information since the LDPC information consists of BCH information bits plus parity bits.

All the BCH information bits are divided into  $K_{\text{ldpc}}/360 (= N_{\text{group}})$  bit-groups which are  $(N_{\text{group}} - 1)$  groups of 360 bits and 1 group of 192 bits as illustrated in figure 30. That is,  $K_{\text{bch}} = 7\,032$  BCH information bits are divided into  $7\,200/360 = 20$  bit-groups which are 19 groups of 360 bits and 1 group of 192 bits.

The shortening of the BCH information is performed on bit-group basis, in other words, the positions of the  $(K_{\text{bch}} - K_{\text{sig}})$  zero-padded bits for BCH encoding are allocated according to the bit-groups. For easy understanding, a concrete example of shortening and BCH encoding is given in the next clauses.

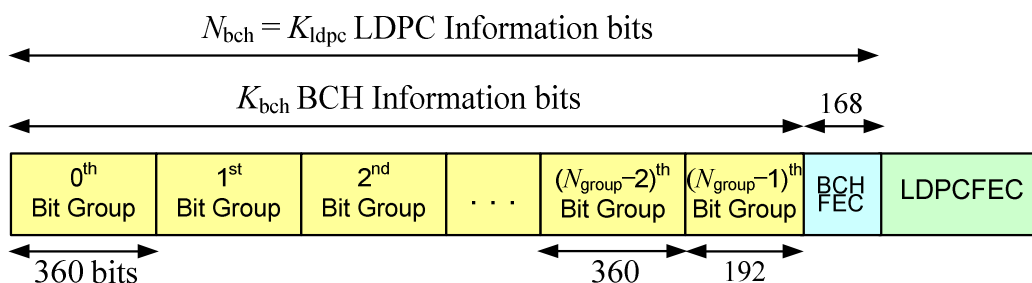


Figure 30: Format of data after LDPC encoding of L1 signalling part 2

### 8.4.5.2 Example for shortening of BCH information

Assume that the pure L1 signalling part 2 consists of 11 956 bits, i.e.  $K_{\text{L1part2\_ex\_pad}} = 11\,956$ .  $K_{\text{L1part2\_ex\_pad}}$  denotes the number of information bits of the L1 signalling part 2 excluding the padding field, L1\_PADDING, as described in clause 8.3.3 in ETSI EN 302 769 [i.1]. Then, the number of LDPC blocks needed for the protection of L1 signalling part 2,  $N_{\text{L1part2\_FEC\_Block}}$  is determined by:

$$N_{\text{L1part2\_FEC\_Block}} = \left\lceil \frac{K_{\text{L1part2\_ex\_pad}}}{N_{\text{L1part2\_max\_per\_Symbol}}} \right\rceil = \left\lceil \frac{11956}{4759} \right\rceil = 3$$

where  $\lceil x \rceil$  means the smallest integer larger than or equal to  $x$ .



Next, the length of L1\_PADDING field,  $K_{L1\_PADDING}$  is calculated as:

$$K_{L1\_PADDING} = \left\lceil \frac{K_{L1part2\_ex\_pad}}{N_{L1part2\_FEC\_Block}} \right\rceil \times N_{L1part2\_FEC\_Block} - K_{L1part2\_ex\_pad}$$

$$= \left\lceil \frac{11956}{3} \right\rceil \times 3 - 11956 = 2.$$

Then the final length of the whole L1 signalling part 2 including the padding field,  $K_{L1part2}$  is set as follows:

$$K_{L1part2} = K_{L1part2\_ex\_pad} + K_{L1\_PADDING} = 11956 + 2 = 11958$$

These  $K_{L1part2}$  signalling bits are divided into 3 ( $= N_{L1part2\_FEC\_Block}$ ) blocks consisting of  $K_{sig}$  bits which is calculated as

$$K_{sig} = \frac{K_{L1part2}}{N_{L1part2\_FEC\_Block}} = \frac{11958}{3} = 3986$$

Each block, with information size of  $K_{sig}$ , is protected by a concatenation of BCH outer code and LDPC inner code.

Here, 7 032 BCH information bits consist of a segmented block of 3 986 ( $= K_{sig}$ ) bits and

3 046 ( $= K_{bch} - K_{sig} = 7\,032 - 3\,986$ ) zero-padded bits. The positions for zero-padding are determined by the following procedure described in clause 8.4.3.1 in ETSI EN 302 769 [i.1].

First, all 7 032 BCH information bits denoted by  $\{m_0, m_1, \dots, m_{7031}\}$  are divided into 20 ( $= N_{group} = K_{ldpc}/360$ ) groups as follows:

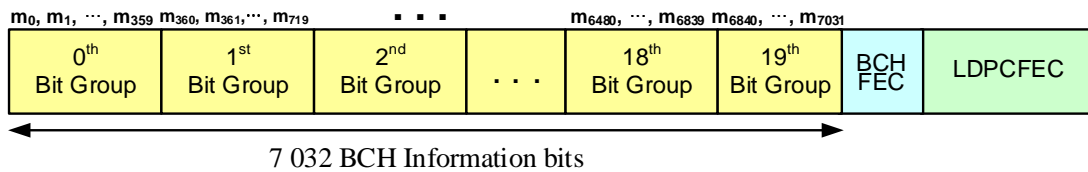
$$X_0 = \{m_0, m_1, m_2, \dots, m_{359}\}, (360 \text{ bits})$$

$$X_1 = \{m_{360}, m_{361}, m_{362}, \dots, m_{719}\}, (360 \text{ bits})$$

$$X_{18} = \{m_{6480}, m_{6481}, m_{6482}, \dots, m_{6839}\}, (360 \text{ bits})$$

$$X_{19} = \{m_{6840}, m_{6841}, m_{6842}, \dots, m_{7031}\}, (192 \text{ bits})$$

The bits in each bit-group correspond to the data field in ascending order as illustrated in figure 31.



**Figure 31: Allocation of BCH information bits in Data field**

Then, the shortening procedure is as follows:

Step 1) Compute the number of groups in which all the bits will be padded,  $N_{pad}$  as follows:

$$N_{pad} = \left\lfloor \frac{7032 - 3986}{360} \right\rfloor = \left\lfloor \frac{3046}{360} \right\rfloor = 8$$

Step 2) All information bits of 8 groups,  $X_{18}, X_{17}, X_{16}, X_{15}, X_{14}, X_{13}, X_{12}$ , and  $X_{11}$  are padded with zeros.

- Step 3) For the group  $X_4$ , 166 ( $= 7\,032 - 3\,986 - 360 \times 8$ ) information bits in the last part of  $X_4$  are additionally padded. In Steps 2) and 3), the choice of bit-groups is determined by table 28 in clause 8.4.3.1 in ETSI EN 302 769 [i.1].
- Step 4) Finally, 3 986 ( $= K_{sig}$ ) information bits are sequentially mapped to bit positions which are not zero-padded in 7 032 BCH information bits,  $\{m_0, m_1, \dots, m_{7031}\}$ , by the above procedure.

If 3 986 information bits which do not correspond to zero-padded position are denoted by  $\{s_0, s_1, \dots, s_{3985}\}$ , the result of above shortening procedure can be presented as follows:

$$m_i = s_i, \quad 0 \leq i < 1634,$$

$$m_i = 0, \quad 1634 \leq i < 1800,$$

$$m_i = s_{i-166}, \quad 1800 \leq i < 3960,$$

$$m_i = 0, \quad 3960 \leq i < 6840,$$

$$m_i = s_{i-3046}, \quad 6840 \leq i < 7032.$$

Figure 32 illustrates the shortening of the BCH information part in this case, i.e. filling the BCH information-bit positions that are not zero-padded with 3 986 information bits.

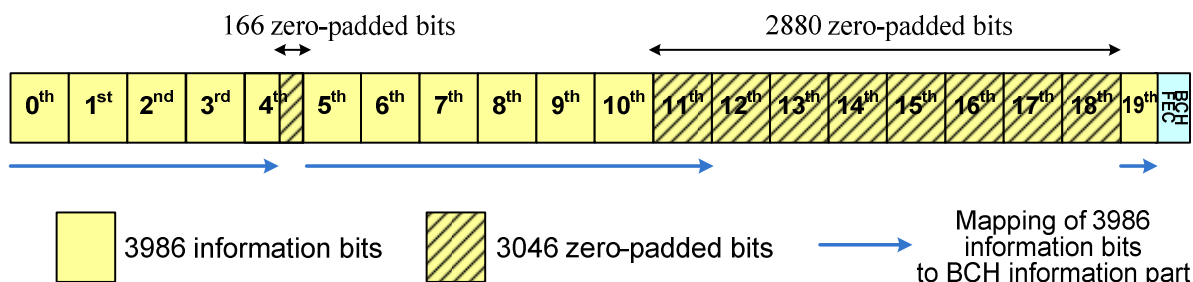


Figure 32: Example of Shortening of BCH information part

### 8.4.5.3 BCH encoding

The  $K_{bch}$  information bits (including the  $K_{bch} - K_{sig}$  zero padding bits) are first BCH encoded according to clause 6.1.1 in ETSI EN 302 769 [i.1] to generate  $N_{bch} = K_{ldpc}$  output bits ( $i_0 \dots i_{N_{bch}-1}$ ).

### 8.4.5.4 LDPC encoding

The  $N_{bch} = K_{ldpc}$  output bits ( $i_0 \dots i_{N_{bch}-1}$ ) from the BCH encoder, including the  $(K_{bch} - K_{sig})$  zero padding bits and the  $(K_{ldpc} - K_{bch})$  BCH parity bits, form the  $K_{ldpc}$  information bits  $I = (i_0, i_1, \dots, i_{K_{ldpc}-1})$  for the LDPC encoder. Note that for the protection of L1 signalling part 2, the number of BCH parity bits,  $N_{bch\_parity} (= K_{ldpc} - K_{bch})$  is fixed as 168. The LDPC encoder systematically encodes the  $K_{ldpc}$  information bits onto a codeword  $A$  of size  $N_{ldpc}$ :

$$A = (i_0, i_1, \dots, i_{K_{ldpc}-1}, p_0, p_1, \dots, p_{M_{ldpc}-K_{ldpc}-1}) = (i_0, \dots, i_{7199}, p_0, p_1, \dots, p_{8999})$$

according to clause 6.1.2 in ETSI EN 302 769 [i.1].

### 8.4.5.5 Puncturing of LDPC parity bits

The 16 K LDPC code with  $K_{\text{ldpc}} = 7\,200$  for L1 signalling part 2 has  $(N_{\text{ldpc}} - K_{\text{ldpc}}) = 9\,000$  parity bits. Since the code length of 16 K LDPC code,  $N_{\text{ldpc}}$  is 16 200, its effective LDPC code rates ( $R_{\text{eff}}$ ) is 4/9. All LDPC parity bits, denoted by  $\{p_0, p_1, \dots, p_{8999}\}$ , are divided into 25 ( $= Q_{\text{ldpc}}$ ) parity bit groups where each parity group is formed from a sub-set of the LDPC parity bits as follows:

$$P_0 = \{ p_0, p_{Q_{\text{ldpc}}}, p_{2Q_{\text{ldpc}}}, \dots, p_{359Q_{\text{ldpc}}} \} = \{ p_0, p_{25}, p_{50}, \dots, p_{8975} \},$$

$$P_1 = \{ p_1, p_{Q_{\text{ldpc}}+1}, p_{2Q_{\text{ldpc}}+1}, \dots, p_{359Q_{\text{ldpc}}+1} \} = \{ p_1, p_{26}, p_{51}, \dots, p_{8976} \},$$

$$P_2 = \{ p_2, p_{Q_{\text{ldpc}}+2}, p_{2Q_{\text{ldpc}}+2}, \dots, p_{359Q_{\text{ldpc}}+2} \} = \{ p_2, p_{27}, p_{52}, \dots, p_{8977} \},$$

$$P_{24} = \{ p_{Q_{\text{ldpc}}-1}, p_{2Q_{\text{ldpc}}-1}, p_{3Q_{\text{ldpc}}-1}, \dots, p_{360Q_{\text{ldpc}}-1} \} = \{ p_{24}, p_{49}, p_{74}, \dots, p_{8999} \}$$

where  $P_j$  represents the  $j$ -th parity group and  $Q_{\text{ldpc}} = 25$  is given in table 5b in ETSI EN 302 769 [i.1]. Each group has  $9\,000/Q_{\text{ldpc}} = 360$  bits, as illustrated in figure 33. Note that the number of parity bits of a DVB-C2 LDPC code is always a multiple of 360 and  $Q_{\text{ldpc}}$ .

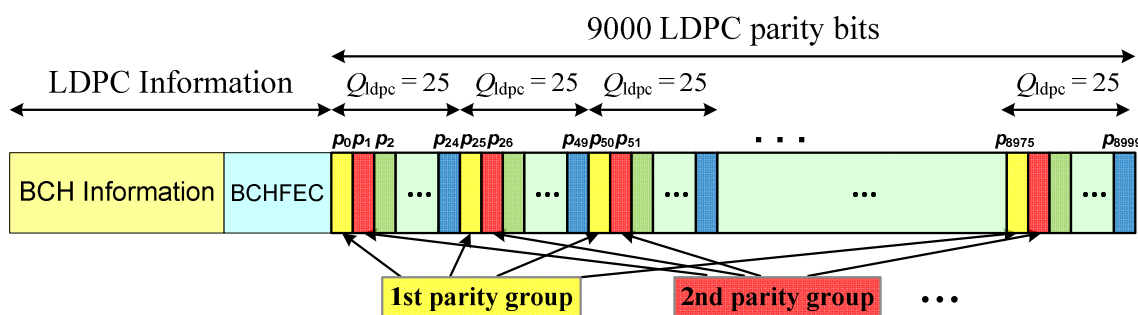


Figure 33: Parity-bit groups in an FEC block

When the shortening is applied to encoding of the signalling bits, some LDPC parity bits are punctured after the LDPC encoding. The punctured LDPC parity bits are not transmitted and they should be regarded as erasures in the receiver.

Similarly to the shortening of BCH information, the puncturing of LDPC parity bits is performed on parity-bit-group basis. Specifically, according to the number of LDPC parity bits to be punctured,  $N_{\text{punc}}$ , the positions of the parity bits are allocated by unit of parity-bit groups. The procedure for calculating  $N_{\text{punc}}$  is a little complicated as described in clause 8.4.2 in ETSI EN 302 769 [i.1] because there is a choice of time interleaving, the length of the L1 signalling part 2 can vary, and the number of modulated cells for each coded **L1-part2** shall be a multiple of the number of OFDM symbols for transmitting L1 part 2.

For easy understanding, a concrete example of calculating  $N_{\text{punc}}$  and puncturing of LDPC parity bits is given in clause 8.4.5.6.

### 8.4.5.6 Example for Puncturing of LDPC parity bits

Assume that the pure L1 signalling part 2 consists of 11 956 bits. Then,  $K_{\text{sig}}$  is 3 986 as calculated in clause 8.4.6.2. Note that the coded **L1-part2** is always transmitted using 16-QAM. For a given  $K_{\text{sig}} = 3\,986$  and modulation order 16-QAM,  $N_{\text{punc}}$  is determined by the following steps:

$$\text{Step 1) } N_{\text{punc\_temp}} = \left\lfloor \frac{6}{5} \times (K_{\text{bch}} - K_{\text{sig}}) \right\rfloor = \left\lfloor \frac{6}{5} \times (7032 - 3986) \right\rfloor = 3655$$

$$\text{Step 2) } N_{\text{L1part2\_temp}} = 3986 + 168 + 16200 \times (1 - 4/9) - 3655 = 9499$$

Step 3) For the sake of convenience, assume that the coded **L1-part2** is transmitted without an extended time interleaving, that is, the value of L1\_TI\_MODE is 00 or 01.

$$N_{L1part2} = \left\lceil \frac{N_{L1part2\_temp}}{2\eta_{MOD} \times N_{L1part2\_FEC\_Block}} \right\rceil \times 2\eta_{MOD} \times N_{L1part2\_FEC\_Block} = \left\lceil \frac{9499}{8 \times 3} \right\rceil \times 8 \times 3 = 9504. (\eta_{MOD} = 4)$$

$$\text{Step 4) } N_{punc} = N_{punc\_temp} - (N_{L1part2} - N_{L1part2\_temp}) = 3655 - (9504 - 9499) = 3650$$

Step 1 ensures that the effective LDPC code rate of the L1 signalling part 2,  $R_{eff\_L1part2}$  is always less than  $R_{eff\_16K\_LDPC\_1\_2}$  (= 4/9). Furthermore,  $R_{eff\_L1part2}$  tends to decrease as the information length  $K_{sig}$  decreases.

In general, for a fixed code-rate, the code performance is degraded as the information length decreases since the code length also decreases. On the other hand, for a fixed information length, it is clear that the code performance is improved as the code rate decreases. From these facts, the performance variation induced by the variable length of L1 signalling information can be reduced by code-rate control. In other words, the receiving coverage for the signalling information will be approximately invariant and stable, whatever the number of bits to be signalled. Step 1 for systematically calculating the number of puncturing bits is a simple code-rate control technique accomplished by adjusting the multiplicative coefficient.

Step 3 guarantees that  $N_{L1part2}$  is a multiple of the number of columns of the bit interleaver (described in clause 8.4.3.6 in ETSI EN 302 769 [i.1]) and that  $N_{L1part2}/\eta_{MOD}$  is a multiple of the number of OFDM symbols for transmitting L1 signalling part 2. Note that the latter condition should be satisfied when time interleaving is applied to L1 signalling part 2.

$N_{L1part2}$  means the number of the coded bits for each information block. After the shortening and puncturing, the coded bits of each block will be mapped to:

$$N_{MOD\_per\_Block} = \frac{N_{L1part2}}{\eta_{MOD}} = \frac{9504}{4} = 2376$$

modulated cells. The total number of cells in all  $N_{L1part2\_FEC\_Block}$  blocks,  $N_{MOD\_Total}$  is  $N_{MOD\_per\_Block} \times N_{L1part2\_FEC\_Block} = 2376 \times 3 = 7128$ .

Finally, for  $N_{punc} = 3650$  given in Step 4), the positions of bits to be punctured are determined as follows:

$$\text{Step 1) } \text{Compute } N_{punc\_groups} \text{ such that } N_{punc\_groups} = \left\lfloor \frac{3650}{360} \right\rfloor = 10.$$

Step 2) All parity bits of 10 parity bit-groups  $P_6, P_4, P_{13}, P_9, P_{18}, P_8, P_{15}, P_{20}, P_5, P_{17}$  are punctured.

Step 3) For the group  $P_2$ , 50 (= 3650 - 3600) parity bits in the first part of the group will be additionally punctured. That is, the parity bits  $p_2, p_{27}, p_{52}, \dots, p_{1227}$  are punctured. In Steps 2) and 3), the choice of parity bit-groups is determined by table 29 in clause 8.4.3.4 in ETSI EN 302 769 [i.1].

### 8.4.5.7 Removal of zero-padding bits

The  $(K_{bch} - K_{sig})$  zero-padding bits are removed and are not transmitted. This leaves a word consisting of the  $K_{sig}$  information bits, followed by the 168 BCH parity bits and  $9000 - N_{punc}$  (=  $N_{ldpc} - K_{ldpc} - N_{punc}$ ) LDPC parity bits without  $N_{punc}$  punctured bits, as illustrated in figure 34.

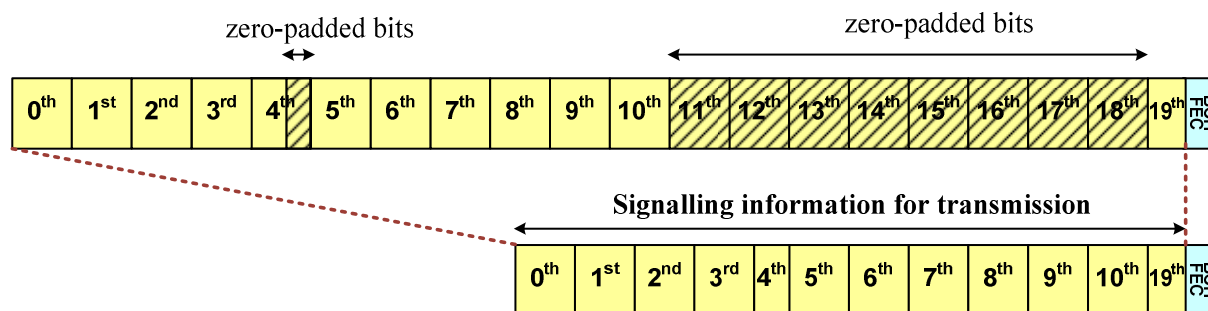


Figure 34: Example of removal of zero-padding bits

#### 8.4.5.8 Bit Interleaving and constellation mapping after shortening and puncturing

The LDPC codeword of length  $N_{L1part\ 2}$ , consisting of  $K_{sig}$  information bits, 168 BCH parity bits, and  $(9\ 000 - N_{punc})$  LDPC parity bits, are bit-interleaved using a block interleaver as described in clause 8.4.3.6 in ETSI EN 302 769 [i.1].

Each bit-interleaved LDPC codeword is mapped onto constellations as described in clause 8.4.4 in ETSI EN 302 769 [i.1].

## 8.5 Interleaving

### 8.5.1 Bit interleaving

Bit Interleaving is described in clause 6.1.3 of ETSI EN 302 769 [i.1].

### 8.5.2 Time interleaving

Time interleaving (TI) for **Data Slice** operates at **Data Slice** level as described in clause 9.4.4 of ETSI EN 302 769 [i.1]. A TI block is the set of **OFDM cells** within which time interleaving is performed; there is no time interleaving between **Data Slices** or TI blocks. No time interleaving is allowed between **Data Slice** and preamble. The **Data Slice** contains integer number of TI blocks. The TI block and **Data Slice** boundaries are synchronized in the time direction; the first TI block starts at the beginning of the **Data Slice** immediately after the preamble and the last TI block ends at the end of a **C2 frame**.

Each TI block of a **Data Slice** within a **C2 frame** contains exactly same number of **OFDM cells** (see clause of 9.4.4 of ETSI EN 302 769 [i.1]). The total number of cells within the TI block is kept constant once relevant parameters - scattered or continual pilot pattern, notch band and reserved dummy carriers for PAPR reduction - are fixed. Each TI block may not contain integer number of FEC blocks. It is not required that each TI block should contain the cells from the same **PLP**. The cells from different **PLPs** can be interleaved into one TI block. However, each TI block should contain all **data cells** carried by TI-depth OFDM symbols within the **Data Slice**.

The address generation equation in clause 9.4.4 of ETSI EN 302 769 [i.1] assumes ( $C$  columns  $\times$   $R$  rows) rectangular memory size for time interleaving. Each cell in the memory one-to-one maps to the **OFDM cell** in the **Data Slice** containing the TI block. All pilots and reserved dummy carriers for PAPR reduction should be taken into account in the address generation process but only the **data cells** should be read out for the final output. The addresses for these non-**data cells** may be temporarily generated but not used for interleaving.

The time interleaving is actually done over only **data cells** so the TI memory may be further reduced to exclude those non-**data cells** in practical implementation. However, the interleaver output should change neither the interleaving sequence nor the number of **data cells** allocated to each **OFDM symbol** within the **Data Slice** when such kind of optimal size of memory is used.

In the presence of **notch Bands** within the bandwidth of a **Data Slice**, i.e. Narrow Band **notches**, the carriers occupied by these **notches** shall be considered when calculating the addresses as in the case of pilots or reserved tones.

### 8.5.3 Frequency Interleaving

The frequency interleaver in DVB-C2 is applied to the payload cells of a given **Data Slice** from one OFDM **data symbol** to the next. The number of payload cells per **Data Slice** per OFDM **data symbol** ( $N_{DS}$ ) can vary from **data symbol** to **data symbol**.  $N_{DS}$  comprises the number of cells between ( $K_{DS,max} - K_{DS,min}$ ) minus the number of continual pilots, scattered pilots, reserved tones and cells that are located in **notches** within the **Data Slice**. Most of this information is carried in the Layer 1 signalling. For this purpose, assume that a function  $DataCells(slice\ number, data\ symbol\ number, L1\ info)$  exists in the modulator to provide  $N_{DS}$  for a given **Data Slice** number, data symbol number and from the Layer 1 information. The frequency interleaver shall be capable of interleaving payload cells of the largest possible **Data Slice** with  $\max(N_{DS})$  cells given that  $(K_{DS,max} - K_{DS,min}) \leq 3\ 408$ . This means that the interleaver shall be capable of dealing with  $N_{max}$  payload cells where:

$$N_{max} = (K_{DS,max} - K_{DS,min}) - N_{SP,Dx=24} - N_{CP}$$

and  $N_{SP,Dx=24}$  is the number of scattered pilots in 3 408 sub-carriers for the  $D_x = 24$  scattered pilot pattern and  $N_{CP}$  is the number of continual pilots in 3 408 sub-carriers.

The frequency interleaver memory is split into two banks: Bank A for even OFDM **data symbols** and Bank B for odd OFDM **data symbols**. Each memory bank comprises of  $N_{max}$  locations. DVB-C2 uses odd-only pseudo-random interleaving. In this the payload cells from even **data symbols** (data symbol number of form  $2n$ ) of the **Data Slice** are written into the interleaver memory Bank A in a sequential order and read out in a permuted order defined by  $H_0(q)$ . Similarly, payload cells from odd OFDM **data symbols** (data symbol number of form  $2n + 1$ ) of the **Data Slice** are written into interleaver memory Bank B in a sequential order and read out in a permuted order defined by  $H_1(q)$ . In each case, the permuted order addresses  $H_{\{0,1\}}(q)$  are provided by the pseudo-random address generator from clause 9.4.5 of ETSI EN 302 769 [i.1]. In order to produce a continuous stream of cells at its output, when Bank A is being written (incoming even **data symbol**), Bank B is also being read (outgoing previous odd **data symbol**) at the same time. Indeed, the sequential counter  $q$  doubles as both the sequential write address and the look up table index to each of the permutation functions  $H_{\{0,1\}}(q)$ . If all **data symbols** in the **Data Slice** contained  $N_{max} = C_{data}$  payload cells, then the number of write addresses for symbol number  $2n + 1$  would match the number of read addresses for data symbol number  $2n$  otherwise some **data cells** of **data symbol**  $2n$  will be skipped. Unfortunately  $N_{DS}$  can be different from **data symbol** to **data symbol**. Suppose  $N_{DS}(2n)$  is less than  $N_{DS}(2n + 1)$  then the pseudo-random address generator  $H_0(q)$  would have to produce more addresses than there are cells to be read from memory Bank A because the sequential write address counter  $q$  for Bank B would range from 0 to  $N_{DS}(2n + 1) - 1$  ( $> N_{DS}(2n)$ ). The case in which  $N_{DS}(2n) > N_{DS}(2n + 1)$  can also occur. In this case the sequential write address counter for Bank B would need to exceed  $N_{DS}(2n + 1) - 1$  as more  $H_0(q)$  read addresses are needed for Bank A. Recalling that the function  $DataCells(slice\ number, data\ symbol\ number, L1\ info)$  returns the number of payload cells in the current slice for the given **data symbol** and noting that HoldBuffer is a small amount of storage with write address  $wptr$  and read address  $rptr$ , the interleaving proceeds as follows at the start of even **data symbol** of number  $2n$ :

- 1)  $q = 0$ ;
- 2)  $C_{max} = \max(DataCells(slice\ number, 2n-1, L1\ info), DataCells(slice\ number, 2n, L1\ info))$ ;
- 3) Generate address  $H_1(q)$ ;
- 4)  $rdEnable = (H_1(q) < DataCells(slice\ number, 2n-1, L1\ info))$ ;
- 5)  $wrEnable = (q < DataCells(slice\ number, 2n, L1\ info))$ ;
- 6) if (rdEnable) Read cell  $q$  of output interleaved **data symbol**  $2n - 1$  from location  $H_1(q)$  of memory Bank B;
- 7) Store cell  $q$  of incoming un-interleaved **data symbol**  $2n$  into location  $wptr$  of HoldBuffer and increment  $wptr$ ;
- 8) if (wrEnable):
  - a) Write cell  $rptr$  of HoldBuffer into location  $q$  of memory Bank A and increment  $rptr$ .
  - b) If( $wptr == rptr$ ) reset both  $rptr = wptr = 0$ .
- 9) Increment  $q$ ;

- 10) if ( $q < C_{\max}$ ) go to 3.

Then with **data symbol**  $2n+1$  at the input of the interleaver:

- 1)  $q = 0$ ;
- 2)  $C_{\max} = \max(\text{DataCells}(\text{slice number}, 2n, \text{L1 info}), \text{DataCells}(\text{slice number}, 2n + 1, \text{L1 info}))$ ;
- 3) Generate address  $H_0(q)$ ;
- 4)  $\text{rdEnable} = (H_0(q) < \text{DataCells}(\text{slice number}, 2n, \text{L1 info}))$ ;
- 5)  $\text{wrEnable} = (q < \text{DataCells}(\text{slice number}, 2n + 1, \text{L1 info}))$ ;
- 6) if (rdEnable) Read cell  $q$  of output interleaved **data symbol**  $2n$  from location  $H_0(q)$  of memory Bank A;
- 7) Store cell  $q$  of incoming un-interleaved **data symbol**  $2n + 1$  into location  $wptr$  of HoldBuffer and increment  $wptr$ ;
- 8) if (wrEnable):
  - a) Write cell  $wptr$  of HoldBuffer into location  $q$  of memory Bank B and increment  $rptr$ .
  - b) If( $wptr == rptr$ ) reset both  $rptr = wptr = 0$ .
- 9) Increment  $q$ ;
- 10) if ( $q < C_{\max}$ ) go to 3.

The required width for each memory location depends on the resolution with which each cell is represented after QAM mapping.

Care should be taken to implement the interleaving function in the correct sense. As shown in the steps detailed above, the interleaver should work as follows:

- For each **data symbol**, the interleaver should write to the memory in normal order and read in permuted order.

## 8.6 Framing

The OFDM based **C2 frame** structure is shown in time and frequency direction in figure 35. The **C2 frame** structure comprises  $L_P$  Preamble Symbols ( $L_P \geq 1$ ) followed by  $L_{data}$  **Data symbols** in time direction. The beginning of the first Preamble Symbol marks the beginning of the **C2 frame**. The number of Preamble Symbols  $L_P$  depends on both the information length at the beginning of each L1 signalling part 2 block and the chosen L1 time interleaving depth.

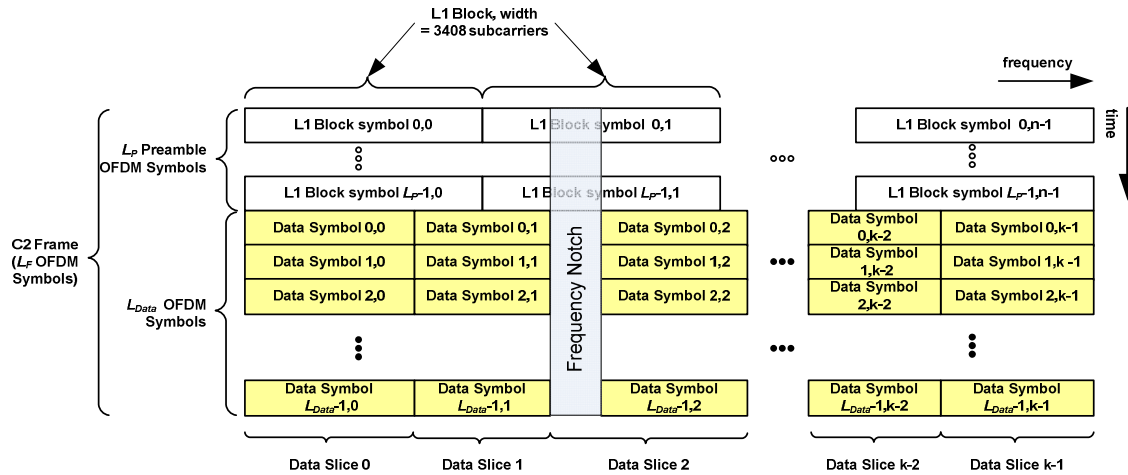
The data part of the **C2 frame** consists of  $L_{data}=448$  symbols (approximately 203,8 msec for  $GI = 1/64$  or 202,2 msec for  $GI = 1/128$ ,  $T_U = 448 \mu s$ ). The **C2 frame** duration is therefore given by:

$$T_F = (L_P + L_{data}) \times T_S,$$

where  $T_S$  is the total **OFDM symbol** duration.

The Preamble Symbols are divided in frequency direction into L1 block symbols of same bandwidth (3 408 subcarriers or approximately 7,61 MHz). The frequency specific preamble pilot pattern allows for reliable time and frequency synchronization (e.g. by correlation). The equidistant spacing of the L1 blocks allows to extract the L1 signalling in any receiver tuning position, even if the tuning position comprises parts of two neighbored L1 blocks. Since all L1 blocks of a C2 signal comprise the same information the complete L1 signalling information can be retrieved by reordering the OFDM carriers of the two L1 signalling blocks (see clause 10.1.1.5.1). The L1 signalling in the preamble contains all OFDM and **Data Slice** specific information that are needed to decode the desired **PLP**.

While the L1 signalling blocks have a fixed and constant bandwidth, **Data Slices** have an arbitrary bandwidth as a multiple of the pilot pattern specific granularity. This granularity depends on the chosen guard interval and has a value of 12 subcarriers for  $GI=1/64$  and 24 subcarriers for  $GI = 1/128$ . As an upper limit **Data Slices** shall not exceed the L1 block symbol bandwidth (i.e. 3 408 subcarriers).



NOTE: The **C2** frame starts with at least one Preamble Symbol ( $L_p$ ) followed by  $L$  Data symbols.

**Figure 35: The C2 frame structure**

Within each **Data Slice** **Data Slice** Packets of one or multiple **PLPs** are embedded. **Data Slice** Packets are not aligned to the C2 framing structure itself, i.e. one **Data Slice packet** may overlap over **C2** frames (being 'interrupted' by the preamble). As described in clause 7 of ETSI EN 302 769 [i.1] there are two formats of **Data Slice** Packets. Either they have a FECFRAME header that allow to synchronize to the packet and to extract all information that is needed to decode this **Data Slice** Packet (Type 2), or the 1st occurrence of a **Data Slice packet** is signalled in the preamble with a pointer mechanism (Type 1, requires constant modulation and coding).

Frequency **notches** can be inserted into the C2 signal across a **C2** frame. Frequency **notches** are used for two reasons. On the one hand notched frequencies reduce radiation of C2 signals from the cable networks on the air. In addition they are used to improve the C2 signal quality by excluding frequency ranges that are affected from interfering signals (e.g. CW carriers, etc.). The C2 standard [i.1] differs between narrowband and broadband **notches** and is described in more detail in clause 9.3.5 of ETSI EN 302 769 [i.1].

## 8.7 OFDM Signal Generation

### 8.7.0 Introduction of OFDM Signal Generation

The DVB-C2 specification introduces the new concept of Absolute OFDM, in which all OFDM subcarriers are seen relative to the absolute frequency 0 MHz instead of a centre frequency as used in other systems, e.g. DVB-T2. This leads to the formulas given in clause 10.1 of ETSI EN 302 769 [i.1] that mathematically define the transmitted signal, but are impractical for real implementations. Real implementations for OFDM signal generation are normally based on the Fast Fourier Transform and the equivalent lowpass representation of signals. However, the generation of a standard compliant DVB-C2 signal using the equivalent lowpass representation requires additional considerations. Unwanted phase jumps may be generated between adjacent OFDM symbols that could disturb the synchronization procedure within the receiver. The reasons for these phase jumps are described in detail in clause 8.7.1. Clauses 8.7.2 and 8.7.3 present practical solutions for the implementation of a standard compliant DVB-C2 transmitter.

#### 8.7.1 OFDM Modulation Using the Equivalent Lowpass Representation

Generally, the generation of the OFDM signal in the passband is impractical, as this requires extremely high sampling rates. Hence, the generation in the equivalent lowpass representation is normally used [i.22]. Afterwards, the signal is shifted from the equivalent lowpass representation into the desired passband.

The DVB-C2 standard ETSI EN 302 769 [i.1] defines the emitted passband signal by the following expressions:

$$s(t) = \text{Re} \left\{ \sum_{n=0}^{\infty} \left[ \frac{1}{\sqrt{K_{total}}} \sum_{l=0}^{L_p-1} \sum_{k=K_{min}}^{K_{max}} c_{m,l,k} \cdot \psi_{m,l,k}(t) \right] \right\}$$



where:

$$\psi_{m,l,k}(t) = \begin{cases} e^{j2\pi\frac{k}{T_U}(t-\Delta-lT_s-mT_F)} & mT_F + lT_S \leq t < mT_F + (l+1)T_S \\ 0 & \text{otherwise} \end{cases}$$

and:

$k$  denotes the carrier number;

$l$  denotes the **OFDM symbol** number starting from 0 for the first Preamble Symbol of the frame;

$m$  denotes the **C2 frame** number;

$K_{total}$  is the number of transmitted carriers, i.e.  $K_{total} = K_{max} - K_{min} + 1$ , assumed to be a multiple of 2;

$L_F$  total number of OFDM Symbols per frame (including the preamble);

$T_S$  is the total symbol duration for all symbols, and  $T_S = T_U + \Delta$ ;

$T_U$  is the active symbol duration;

$\Delta$  is the duration of the guard interval;

$c_{m,l,k}$  is the complex modulation value for carrier  $k$  of the **OFDM symbol** number  $l$  in **C2 frame** number  $m$ ;

$T_F$  is the duration of a frame,  $T_F = L_F T_S$ ;

$K_{min}$  is the carrier index of first (lowest frequency) active carrier;

$K_{max}$  is the carrier index of last (highest frequency) active carrier.

In order to generate this signal using the equivalent lowpass representation, the multiplication with a carrier signal, i.e.  $e^{j2\pi f_c t}$ , has to be included. The term within the sums then describes the equivalent lowpass representation of the signal. However, the carrier signal has to be compensated within the equation  $\Psi$  to obtain the same output signal:

$$s(t) = \frac{1}{\sqrt{K_{total}}} \cdot \text{Re} \left\{ e^{j2\pi f_c t} \cdot \sum_{m=0}^{\infty} \sum_{l=0}^{L_F-1} \sum_{k=K_{min}}^{K_{max}} c_{m,l,k} \cdot \Psi'_{m,l,k}(t) \right\} \quad (1)$$

with

$$\Psi'_{m,l,k}(t) = \begin{cases} e^{j2\pi\frac{k}{T_U}(t-\Delta-lT_s-mT_F)} \cdot e^{-j2\pi f_c t} & mT_F + lT_S \leq t < mT_F + (l+1)T_S \\ 0 & \text{otherwise} \end{cases} \quad (2)$$

Equation (2) cannot be directly transformed into the equation known from clause 9.5 of the DVB-T2 specification, ETSI TS 102 831 [i.4]. The reason is the second exponential term, i.e. the carrier signal. While the DVB-T2 equations are independent from the actual carrier frequency  $f_c$ , this initially will lead to phase jumps between adjacent OFDM symbols of the DVB-C2 signal. However, this effect can be avoided as explained in the following clauses.

The carrier frequency, which is not necessarily the centre frequency of the DVB-C2 signal, shall be defined as:

$$f_c = \frac{k_c}{T_U},$$

where  $1/T_U$  is the OFDM subcarrier spacing, and  $k_c$  is the OFDM subcarrier index at the carrier frequency  $f_c$ .

Furthermore,  $k$  shall be substituted by  $k = k' + k_c$ . This leads to:

$$\psi'_{m,l,k}(t) = \begin{cases} e^{j2\pi \frac{k'+k_c}{T_U}(t-\Delta-lT_s-mT_F)} \cdot e^{-j2\pi \frac{k_c}{T_U}t} & mT_F + lT_S \leq t < mT_F + (l+1)T_S \\ 0 & \text{otherwise} \end{cases} \quad (3)$$

which can be reformulated as:

$$\psi'_{m,l,k}(t) = \begin{cases} e^{j2\pi \frac{k'}{T_U}(t-\Delta-lT_s-mT_F)} \cdot e^{-j2\pi \frac{k_c}{T_U}\Delta(1+l+mL_F)} & mT_F + lT_S \leq t < mT_F + (l+1)T_S \\ 0 & \text{otherwise} \end{cases} \quad (4)$$

Equation (4) looks similar to the signal definition of the DVB-T2 signal as described in clause 9.5 of ETSI TS 102 831 [i.4]. However, both equations still differ in the last exponential term. This term is independent of the time  $t$  and causes a constant phase rotation for all OFDM subcarriers of a given OFDM symbol. Naturally, it is possible to choose  $k_c$  freely (and thus  $f_c$ ) and to compensate this phase rotation. However, this term can be avoided by choosing  $k_c$  properly. For this purpose, equation (4) can be written as:

$$\psi'_{m,l,k}(t) = \begin{cases} e^{j2\pi \frac{k'}{T_U}(t-\Delta-lT_s-mT_F)} \cdot e^{-j2\pi \frac{k_c}{T_U}T_U \left(\frac{\Delta}{T_U}\right)(1+l+mL_F)} & mT_F + lT_S \leq t < mT_F + (l+1)T_S \\ 0 & \text{otherwise} \end{cases} \quad (5)$$

where  $\left(\frac{\Delta}{T_U}\right)$  is the relative Guard Interval duration, i.e. 1/128 of 1/64. Additional simplification of (5) leads to:

$$\psi'_{m,l,k}(t) = \begin{cases} e^{j2\pi \frac{k'}{T_U}(t-\Delta-lT_s-mT_F)} \cdot e^{-j2\pi k_c \left(\frac{\Delta}{T_U}\right)(1+l+mL_F)} & mT_F + lT_S \leq t < mT_F + (l+1)T_S \\ 0 & \text{otherwise} \end{cases} \quad (6)$$

Hence, this leads to a common phase rotation of:

$$\varphi_{k_c} = -2\pi \cdot k_c \left(\frac{\Delta}{T_U}\right) \quad (7)$$

for all OFDM subcarriers between two consecutive OFDM symbols, which depends on the choice of the relative Guard Interval duration  $\left(\frac{\Delta}{T_U}\right)$  (i.e. 1/64 or 1/128) and the OFDM subcarrier  $k_c$  at the carrier frequency.

If  $k_c \cdot \left(\frac{\Delta}{T_U}\right)$  is integer, the phase shift can be removed from the equation as it becomes multiples of  $2\pi$ . Hence, if  $k_c$

is multiple of 128 for Guard Interval 1/128, or multiple of 64 for Guard Interval 1/64, equation (6) can be written as:

$$\psi'_{m,l,k}(t) = \begin{cases} e^{j2\pi \frac{k'}{T_U}(t-\Delta-lT_s-mT_F)} & mT_F + lT_S \leq t < mT_F + (l+1)T_S \\ 0 & \text{otherwise} \end{cases} \quad (8)$$

which is similar to the equation for the generation of a DVB-T2 signal. However, it has to be noted that the carrier frequency  $f_c$  is not the centre frequency of the DVB-C2 signal in most cases.

## 8.7.2 Calculation using the Fast Fourier Transform

### 8.7.2.0 Introduction

This clause describes two possible signal generations using the equivalent lowpass representation of the signal in the transmitter. The first approach uses the centre frequency with pre-distortion, while the second approach uses an optimized carrier frequency.

It has to be noted that similar considerations are also valid for the receiver. However, the receiver may use the continual pilots to remove the common phase rotations, which is similar to the compensation of the Common phase error introduced by phase noise.

#### 8.7.2.1 Generation Using the Centre Frequency of the Signal with Predistortion

The centre frequency of the DVB-C2 signal can be described as:

$$f_c = \frac{k_c}{T_U} \quad (9)$$

with

$$k_c = \frac{K_{\max} + K_{\min}}{2} \quad (10)$$

However, this would lead to unwanted common phase rotations between consecutive OFDM symbols. Hence, these rotations have to be compensated for generating a standard compliant signal. Therefore, the signal can be generated using the equations:

$$s(t) = \frac{1}{\sqrt{K_{\text{total}}}} \cdot \text{Re} \left\{ e^{j2\pi f_c t} \cdot \sum_{m=0}^{\infty} \sum_{l=0}^{L_F-1} \sum_{k=K_{\min}}^{K_{\max}} \left( c_{m,l,k} \cdot e^{j\varphi_{m,l}} \right) \cdot \Psi''_{m,l,k}(t) \right\} \quad (11)$$

with

$$\Psi''_{m,l,k}(t) = \begin{cases} e^{j2\pi \frac{k}{T_U} (t - \Delta - lT_s - mT_F)} & mT_F + lT_s \leq t < mT_F + (l+1)T_s \\ 0 & \text{otherwise} \end{cases} \quad (12)$$

and

$$\varphi_{m,l} = -\varphi_{k_c} \cdot (1 + l + m \cdot L_F) \quad (13)$$

where:

- $k$  denotes the carrier number;
- $k_c$  denotes the OFDM subcarrier at the carrier frequency  $f_c$ ;
- $k'$  denotes the carrier number relative to the OFDM subcarrier at the carrier frequency  $f_c$ , i.e.  $k' = k - k_c$ ;
- $l$  denotes the **OFDM symbol** number starting from 0 for the first Preamble Symbol of the frame;
- $m$  denotes the **C2 frame** number;
- $K_{\text{total}}$  is the number of transmitted carriers, i.e.  $K_{\text{total}} = K_{\max} - K_{\min} + 1$ ;
- $L_{F(m)}$  total number of OFDM Symbols per frame with index  $m$  (including the preamble);
- $T_S$  is the total symbol duration for all symbols, and  $T_S = T_U + \Delta$ ;

$T_U$  is the active symbol duration;

$\Delta$  is the duration of the guard interval;

$c_{m,l,k}$  is the complex modulation value for carrier  $k$  of the **OFDM symbol** number  $l$  in **C2 frame** number  $m$ ;

$T_F$  is the duration of a frame,  $T_F = L_F T_s$  ;

$K_{\min}$  is the carrier index of first (lowest frequency) active carrier;

$K_{\max}$  is the carrier index of last (highest frequency) active carrier;

$\varphi_{k_c}$  Phase jump between two consecutive OFDM symbols as calculated equation (7) of clause 8.7.1.

Practically, this generation is equivalent to the generation of a DVB-T2 signal, ETSI TS 102 831 [i.4]. The only difference is the additional phase correction term  $\varphi_{m,l}$  that linearly increases every **OFDM symbol** and compensates the unwanted phase rotations in the generated output signal. The data  $c'_{m,k}$  that is used for calculating the inverse FFT is the inner bracket of equation (11), i.e.  $(c_{m,l,k} \cdot e^{j\varphi_{m,l}})$ .

### 8.7.2.2 Generation Using the Optimum Carrier Frequency

As described in the previous clauses, a common phase rotation may be introduced to the system, depending on the carrier frequency. This common phase rotation can be compensated in order to obtain an output signal as defined in ETSI EN 302 769 [i.1]. Alternatively, this common phase rotation can be avoided by carefully choosing the carrier frequency. Therefore, the OFDM subcarrier at the carrier frequency can be chosen as:

$$k_c = \left\lfloor \frac{K_{\max} + K_{\min}}{2} \cdot \frac{\Delta}{T_U} + \frac{1}{2} \right\rfloor \cdot \frac{1}{\left( \frac{\Delta}{T_U} \right)} \quad (14)$$

where  $\Delta/T_U$  is the relative Guard Interval duration (i.e. 1/64 or 1/128). Practically, equation (14) obtains the carrier  $k_c$  that it is closest to the centre OFDM subcarrier  $(K_{\max} + K_{\min})/2$ , and additionally, generates multiples of  $2\pi$  in equation (7) of the previous clause.

Consequently, the optimum carrier frequency  $f_c$  is:

$$f_c = \frac{k_c}{T_U} \quad (15)$$

where  $1/T_U$  is the OFDM subcarrier spacing. Here, the resulting carrier frequency  $f_c$  is not the centre frequency of the OFDM signal in most cases. The maximum difference between carrier frequency and centre frequency can reach approximately 140 kHz (8 MHz channel raster).

If the carrier frequency  $f_c$  is chosen as described in clause 8.7.2, the transmitted signal according to clause 8.7.1 can be described as:

$$s(t) = \frac{1}{\sqrt{K_{\text{total}}}} \cdot \text{Re} \left\{ e^{j2\pi f_c t} \cdot \sum_{m=0}^{\infty} \sum_{l=0}^{L_F-1} \sum_{k=K_{\min}}^{K_{\max}} c_{m,l,k} \cdot \Psi'_{m,l,k}(t) \right\} \quad (16)$$

with

$$\Psi'_{m,l,k}(t) = \begin{cases} e^{j2\pi \frac{k}{T_U} (t - \Delta - lT_s - mT_F)} & mT_F + lT_S \leq t < mT_F + (l+1)T_S \\ 0 & \text{otherwise} \end{cases} \quad (17)$$

where:

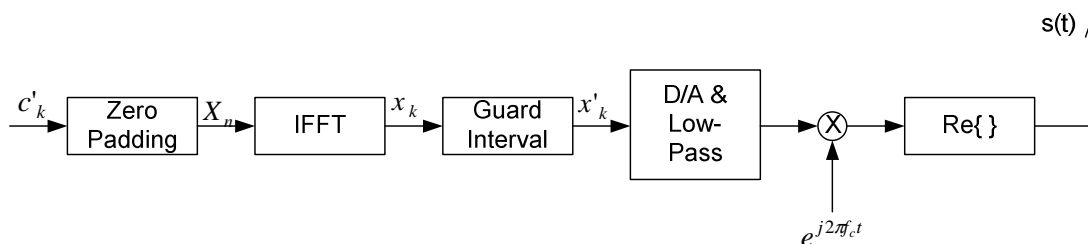
- $k$  denotes the carrier number;
- $k_c$  denotes the OFDM subcarrier at the carrier frequency  $f_c$ ;
- $k'$  denotes the carrier number relative to the OFDM subcarrier at the carrier frequency  $f_c$ , i.e.  $k'=k-k_c$ ;
- $l$  denotes the **OFDM symbol** number starting from 0 for the first Preamble Symbol of the frame;
- $m$  denotes the **C2 frame** number;
- $K_{total}$  is the number of transmitted carriers, i.e.  $K_{total} = K_{max} - K_{min} + 1$ ;
- $L_F$  total number of OFDM Symbols per frame (including the preamble);
- $T_S$  is the total symbol duration for all symbols, and  $T_S = T_U + \Delta$ ;
- $T_U$  is the active symbol duration;
- $\Delta$  is the duration of the guard interval;
- $c_{m,l,k}$  is the complex modulation value for carrier  $k$  of the **OFDM symbol** number  $l$  in **C2 frame** number  $m$ ;
- $T_F$  is the duration of a frame,  $T_F = L_F T_S$ ;
- $K_{min}$  is the carrier index of first (lowest frequency) active carrier;
- $K_{max}$  is the carrier index of last (highest frequency) active carrier.

The data  $c'_k$  that is used for calculating the inverse FFT are the coefficients  $c_{m,l,k}$  of equation (16).

## 8.7.3 OFDM Generation Block Diagram

### 8.7.3.0 Introduction

Figure 36 depicts the transmitter block diagram for the generation of the OFDM signal, which will be described in detail in the following clauses. Firstly, the signal is zero padded for preparation of the Inverse Fast Fourier Transform (IFFT). Then, the Guard Interval is added, the signal is converted from digital to analogue and finally, shifted to the desired passband frequency.



**Figure 36: Possible implementation of OFDM generation**

### 8.7.3.1 Zero Padding

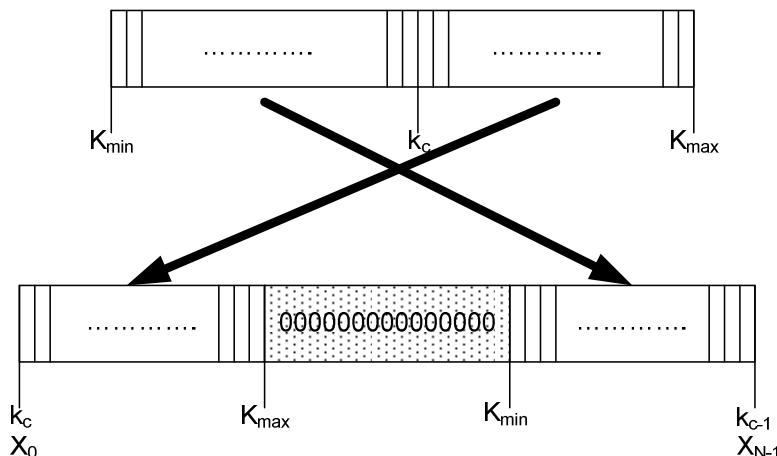
The Zero Padding is required to pre-condition the signal for the transformation of the frequency domain signal into the time domain using the Fast Fourier Transform. Firstly, the signal has to be stuffed in order to fit the **FFT size**  $N$ . Secondly, a realignment of the subcarrier positions is required to be able to use the FFT.

In order to be able to use the Fast Fourier Transform, e.g. based on the Radix 2 algorithm, it has to hold

$N = 2^p$ ,  $p = 1, 2, 3, 4, \dots$ . Furthermore, the value  $N$  shall be significantly higher than the actual number of used OFDM subcarriers in order to avoid alias effects, i.e.:

$$K_{total} = K_{max} - K_{min} + 1 \leq N = K_{total} + x \quad (18)$$

where  $x$  shall be at least 512 for practical implementations.



**Figure 37: Principle of the Zero Padding**

Figure 37 depicts the principle of the Zero Padding. In principle, it realizes a cyclic shift operation on the actually used OFDM subcarriers and inserts zeros to the  $x$  (see equation (18)) remaining positions. Mathematically this operation can be described as:

$$X(n)_{m,l} = \begin{cases} c'_{m,l,k_c+n} & 0 \leq n \leq K_{max} - k_c \\ 0 & \text{otherwise} \\ c'_{m,l,k_c+(n-N)} & N - (k_c - K_{min}) \leq n < N \end{cases} \quad \text{for } 0 \leq n < N \quad (19)$$

where  $X(n)_{m,l}$  (or  $X_n$  in short) is the  $N$  element input signal of the IFFT block.

### 8.7.3.2 IFFT Calculation

The signal has been generated within the frequency domain. The task of the inverse Fast Fourier Transform is the calculation of the corresponding time signal. This is achieved by means of:

$$x(k)_{m,l} = \frac{1}{\sqrt{K_{total}}} \sum_{k=0}^{N-1} X(n)_{m,l} \cdot e^{j2\pi \frac{k \cdot n}{N}} \quad (20)$$

for  $0 \leq k < N$ , where  $m$  is the OFDM symbol,  $l$  the **C2 frame** number,  $K_{total}$  the total number of active OFDM subcarriers, and  $x(k)_{m,l}$  has the short hand notation  $x_k$ .

### 8.7.3.3 Guard Interval Insertion

Figure 38 depicts the insertion of the Guard Interval. This is a cyclic copy of the last part of the useful **OFDM symbol** part, which is copied to the beginning. Mathematically, the OFDM symbol  $x'(k)$  including the Guard Interval is obtained as described in equation (21).

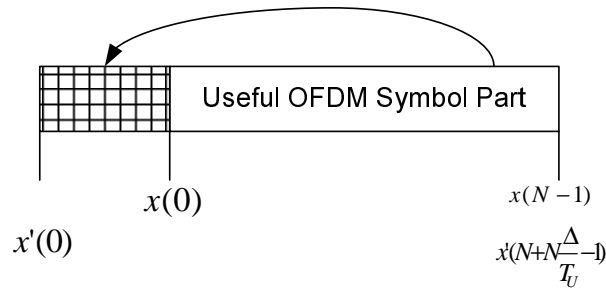


Figure 38: Generation of the Guard Interval

$$x'(k)_{m,l} = \begin{cases} x\left(k + N - N \cdot \frac{\Delta}{T_U}\right) & 0 \leq k < N \cdot \frac{\Delta}{T_U} \\ x\left(k - N \cdot \frac{\Delta}{T_U}\right) & N \cdot \frac{\Delta}{T_U} \leq k < N + N \cdot \frac{\Delta}{T_U} \end{cases} \quad (21)$$

#### 8.7.3.4 Digital to Analogue Conversion and Low-pass Filtering

The previous calculations have been made in the digital domain. The task of this block is the conversion into an analogue signal. Therefore, the signal  $x'(k)_{m,l}$  sampled with the sampling rate  $N/T_U$  has to be converted from digital to analogue **OFDM symbol** by OFDM symbol. This causes alias at multiples of the sampling rate as depicted in figure 39 that have to be removed by means of a low-pass filter. This filtering is simpler for higher distances between the wanted and the alias signals, which is the reason why small values of  $x$  for the zero padding (see equation (16) of clause 8.7.3.1) are impractical.

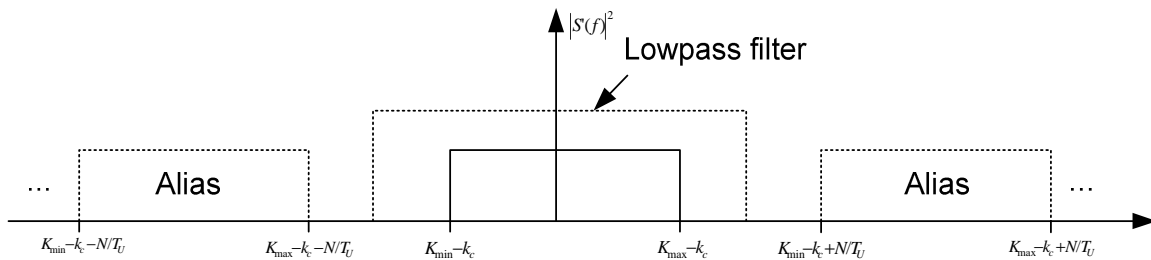


Figure 39: Spectrum of the digital signal and its aliases

#### 8.7.3.5 Frequency Shifting

Final step is the shifting of the equivalent lowpass signal into the wanted passband. This goal is reached by mixing the signal with the carrier frequency  $f_c$ , which is similar to a complex multiplication of the signal by  $e^{j2\pi f_c t}$  and the transmission of its real part.

## 8.8 Spectral Shaping

### 8.8.0 Introduction to the principles of the modulator output signal filtering

The effective bandwidth of a DVB-C2 signal is calculated by the difference between the highest and the lowest sub-carrier frequency. Its minimal and maximal limits are defined by the signalling mechanism chosen. The minimal effective bandwidth amounts to some 7,61 MHz in a system based on 8 MHz channels and to some 5,71 MHz in a system using 6 MHz channel spacing. It is defined by the requirement for a receiver that needs to be capable of receiving a complete **Data Slice** which may not exceed the maximal number of 3 408 sub-carriers equivalent to 7,61 MHz (see clause 9.4.1.2 in ETSI EN 302 769 [i.1]). The maximal effective bandwidth cannot exceed some 450 MHz in 8 MHz cable systems and some 338 MHz in 6 MHz cable systems. The limiting effect is caused by the range of values defined for the parameter assigning the position of the **Data Slice** to the sub-carrier number of the OFDM signal (see clause 9.4.3 in ETSI EN 302 769 [i.1]). The Power Spectral Density (PSD) of the DVB-C2 RF signal has a shape which is typical for signals based on OFDM. The example of figure 40 shows a DVB-C2 signal transmitted in an 8 MHz channel. The corresponding mathematical expression for calculating the PDF is given in the DVB-C2 specification [i.1]. Main characteristics of the PSD are listed as follows:

- 1) The in-band portion of the PSD occupying the frequencies between the lowest and the highest sub-carrier frequency  $K_{\min}/T_U = f_{\text{sc,min}} \leq f \leq f_{\text{sc,max}} = K_{\max}/T_U$  has a rather flat distribution. Due to the introduction of the Guard Interval in the time domain, small periodic ripples occur in the PSD in this frequency range. The ripples have a periodicity equal to the sub-carrier spacing of  $1/T_U$  and an elongation of some fractions of 1 dB. Because of their small elongation, the ripples do not have any implication on the RF transmission characteristic of a DVB-C2 signal. They are however mentioned at this stage for sake of completeness. For a good approximation, the PSD is assumed to have a white distribution within the range of the sub-carrier frequencies.
- 2) PSD has steep edges at frequencies below the smallest sub-carrier frequency  $f < f_{\text{sc,min}} = K_{\min}/T_U$  and above the highest sub-carrier frequency  $f > f_{\text{sc,max}} = K_{\max}/T_U$ . The steepness of the edges decreases exponentially with increasing distance to the sub-carrier frequencies.
- 3) The out-of-band portion of the PSD is attenuated by some 33 dB at frequencies of some 200 kHz distance to  $f_{\text{sc,min}}$  and  $f_{\text{sc,max}}$ , respectively. These frequencies correspond with the channel boundary of an 8 MHz channel. Consequently, all DVB-C2 signal power radiated at frequencies beyond these boundaries is considered to cause interference to signals transmitted in channels adjacent to the DVB-C2 channel.

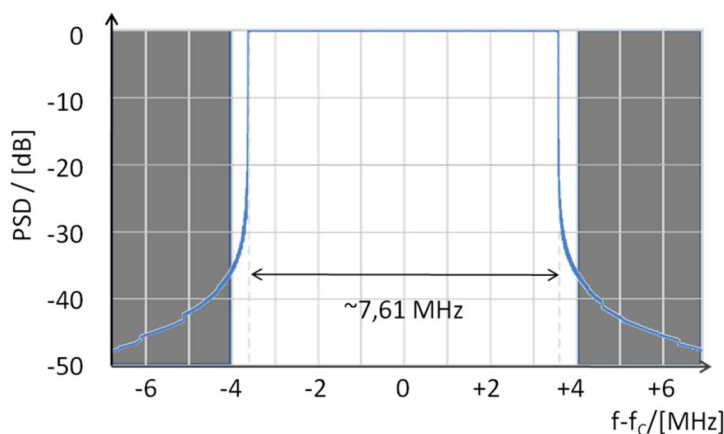
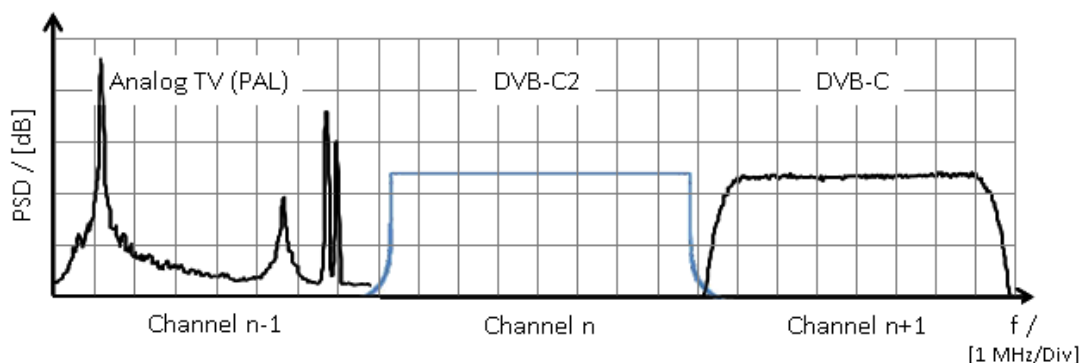


Figure 40: Power Spectral Density of a DVB-C2 RF signal transmitted in an 8 MHz channel



Although the effective bandwidth of a DVB-C2 signal can be flexibly assigned within the boundaries mentioned above, the introduction of DVB-C2 is considered to take place within the grid provided by the traditional channel raster either based on 8 MHz or on 6 MHz channels. In such a scenario, a DVB-C2 signal is injected in a traditional cable channel as illustrated in figure 41 by way of an example. The DVB-C2 signal is placed centric in an 8 MHz channel adjacent to both an analogue TV signal and a DVB-C signal. The power level, which is not standardized by DVB, was chosen to be equivalent to the level of the DVB-C signal. In fact, the adjustment of the DVB-C2 signal level provides a further degree of flexibility for the optimization of the DVB-C2 bandwidth efficiency in relation to the transmission conditions provided by the network. Nevertheless, the selection of the final DVB-C2 transmission level requires consideration of the out-of-band signal power radiated by a DVB-C2 signal exceeding the channel boundaries as described by item 3) above. This out-of-band signal power generates interference with the signals transmitted in the adjacent channels. The minimal transmission quality parameters relevant for cable networks such as the adjacent channel protection conditions are standardized by IEC and CENELEC in their standards series IEC 60728, and are defined in Part 1 [i.18].



**Figure 41: Qualitative depiction of an adjacent channel scenario comprising an analogue TV signal (measured PSD), a DVB-C2 signal (simulated PSD), and a DVB-C signal (measured PSD) in a cable system supporting 8 MHz channel spacing**

In annex D further items concerning an optimized frequency utilization are discussed.

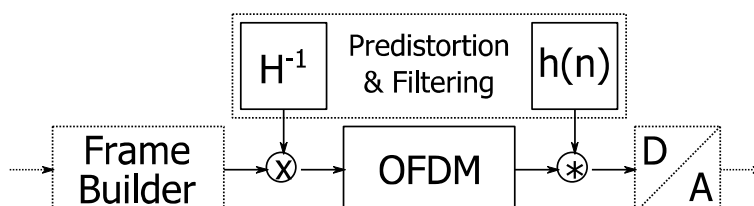
### 8.8.1 Optimized modulator output signal filtering with pre-distortion

Digital filtering improves the out-of-band emissions of the OFDM spectrum and therefore the isolation of the DVB-C2 channel towards adjacent channels. Filtering of the output signal is implementation specific; therefore it is not described in detail in ETSI EN 302 769 [i.1].

However, filtering of the output signal can have an impact on the signal quality and decoding performance, which is especially relevant for high constellation sizes such as 1 024-QAM or 4 096-QAM. Additionally, the channel impulse response length of the digital filter reduces the maximum tolerable echo length.

It is recommended to keep the filter length as short as possible while still providing the required out-of-band filtering effect. Shorter digital filter with high stop band attenuation however suffer from a significant amplitude ripple in the pass band, i.e. in the frequency range of the active OFDM subcarriers. This ripple can be compensated by an artificial pre-distortion of the active OFDM carriers, which is essentially inverse to the pass band channel transfer response of the applied filter.

In clause 4.2 of ETSI EN 302 769 [i.1] the building blocks of the transmitter side of the DVB-C2 signal processing are depicted before the signal is processed by the digital-to-analogue converter.



**Figure 42: Filtering and pre-distortion of a DVB-C2 signal**

Typically, the filtering of the output signal is applied in the digital signal processing domain by convoluting the signal with the filter kernel  $h(n)$  (or by the impulse response  $h(t)$  in the continuous time domain, respectively). In this approach, the distortions of the active carriers introduced by the filtering are compensated by multiplication of the active carriers with the inverse channel transfer function of the filter  $H^{-1}$  at the frequency locations of the active carriers. Depending on the concrete filter used the calculation of the inverse transfer function can be found analytically or by transformation of the impulse response into the frequency domain.

NOTE: It is known that this OFDM carrier specific pre-distortion does not degrade additionally the available guard interval length, as the cyclic convolution operation of the OFDM still holds true.

The filtering effect can be improved by an up sampled OFDM signal.

The OFDM generation, i.e. the IDFT and the insertion of the Guard Interval, is a linear time invariant process. Thus, the pre-distortion of the active carriers by the inverse channel transfer function of the filter results in a flat spectrum when synchronized to the **OFDM symbol** structure in the time domain, i.e. in case of a synchronized receiver.

When including the pre-distortion with filtering approach into the time domain signal representation of a DVB-C2 signal as provided in clause 10.1 of ETSI EN 302 769 [i.1], the (adapted) representation changes as follows:

$$s(t) = h(t) * \sum_l \left[ \delta(t - \Delta - l \cdot T_s) * \left( \text{rect} \left( \frac{t}{T_s} + \Delta \right) \cdot \sum_k H^{-1} \left( \frac{k}{T_U} \right) \cdot c_{l,k} \cdot e^{i2\pi \frac{k}{T_U} t} \right) \right]$$

where the symbols match the expression given in clause 10.1 of ETSI EN 302 769 [i.1].

## 8.9 Broadband DVB-C2 Systems

When scaling the bandwidth of a **C2 system** the following system specific requirement needs to be fulfilled: The useful symbol duration  $T_U$  and thus the carrier spacing  $1/T_U$  is a constant of the **C2 system** which shall not be changed.

Thus, the description of the emitted signal  $s(t)$  as provided in clause 10.1 of ETSI EN 302 769 [i.1] identifies the time continuous representation of a DVB-C2 signal for any combinations of the parameters  $K_{\min}$  and  $K_{\max}$  derived from any valid Layer 1 Signalling information (see example).

Considering the discrete signal generation using the a Discrete Fourier Transformation as described in clause 8.7.3 of the present document the following equation needs to be fulfilled:

$$T_U = T_C \cdot N \text{ with } N \in \{N \geq 4096 \wedge 128 | N\}$$

where  $T_U$  denotes the useful symbol duration,  $T_C$  is the sample rate of the output converter and  $N$  represents the DFT size for the OFDM signal generation. For the trivial case of deploying the minimum active bandwidth of 7,61 MHz for a DVB-C2 signal, the sample rate and the DFT size are well defined as  $T_C = T$  and  $N = 4096$  (see clause 10.1 of ETSI EN 302 769 [i.1]). Since the **elementary period** is a constant of the **C2 system** as well, the **elementary period** needs to be decoupled from the sample rate when scaling the bandwidth of the DVB-C2 signal, i.e. changing the DFT size. The DFT size  $N$  needs to be integer divisible by the least common multiple of the available guard interval fractions, which is 128, to maintain an integer number of samples when inserting the guard interval within the discrete time domain (see example).

EXAMPLE: A DVB-C2 system employing exactly one **Data Slice** of maximum bandwidth results in  $K_{\text{total}} = K_{\text{max}} - K_{\text{min}} + 1 = 3409$  active carriers, the minimum of which resulting in 7,61 MHz bandwidth of a DVB-C2 system. However, adding further **Data Slices** of arbitrary bandwidth scales the DVB-C2 system to higher bandwidth, approx. 450 MHz of which being the highest value that can be signalled.

## 8.10 Peak-to-Average Power Ratio of DVB-C2: theoretical considerations on basis of the Central Limit Theorem

The Peak-to-Average Power Ratio (PAPR) of a digital signal which is transmitted via cable networks is a critical parameter. High peak values may either cause severe intermodulation in cable amplifiers or, even worse, may cause clipping of signal components by lasers in optical transmission systems.

A DVB-C2 signal with 8 MHz bandwidth (3 408 subcarriers) has a PAPR of about 12,2 dB (see figure 43) due to the used OFDM modulation scheme. A DVB-C signal with 8 MHz bandwidth has a PAPR of 10,1 dB, about 2 dB lower than a DVB-C2 signal with 8 MHz bandwidth. Concern was raised that multiple DVB-C2 signals would cause even higher figures of PAPR as the peak values of the individual signals have to be added.

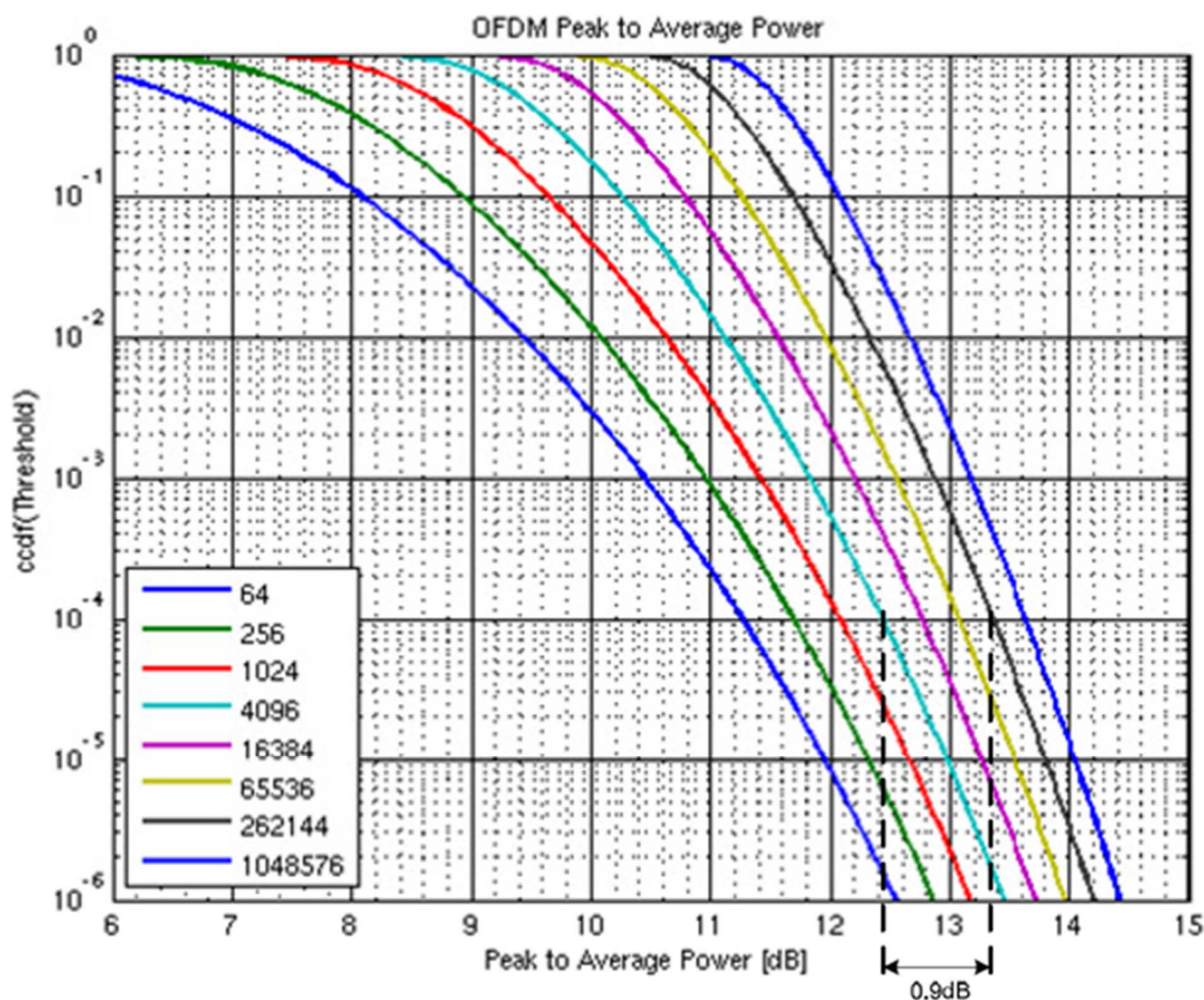


Figure 43: PAPR of different OFDM modulation schemes

According to theoretical considerations and simulations made by TNO it is not correct to add the peak values in order to calculate the Peak-to-Average Power Ratio (PAPR) for a set of digital signals. The Central Limit Theorem has to be applied, where the PAPR of a set of uncorrelated signals approaches asymptotically the Gaussian Probability Density Function (PDF) for an increasing number of contributing signals. Simulations have been made with DVB-C signals. Figure 44 shows the results. The dashed line is the Gaussian PDF. The blue line represents the PDF of a single DVB-C signal, which has a significantly lower PAPR than a signal which would have Gaussian PDF. In a scenario where 10 DVB-C signals are transmitted in a cable system, the PDF follows the magenta graph, which is already close to the Gaussian distribution. Finally the red curve represents 50 DVB-C signals and this curve almost completely follows the Gaussian distribution.

Based on those results, the logical extension to the DVB-C2 case is, that if a significant number of unsynchronized DVB-C2 channels are transmitted in a cable network, or if a DVB-C2 signal is transmitted with a large base of DVB-C channels, the Central Limit Theorem applies as well and the peaks of single DVB-C2 channels are irrelevant for the combined PDF of the total HFC signal-load, which follows the Gaussian distribution. PAPR reduction techniques are therefore not required specifically to minimize PAPR where C2 OFDM-based signals are present in a cable network.

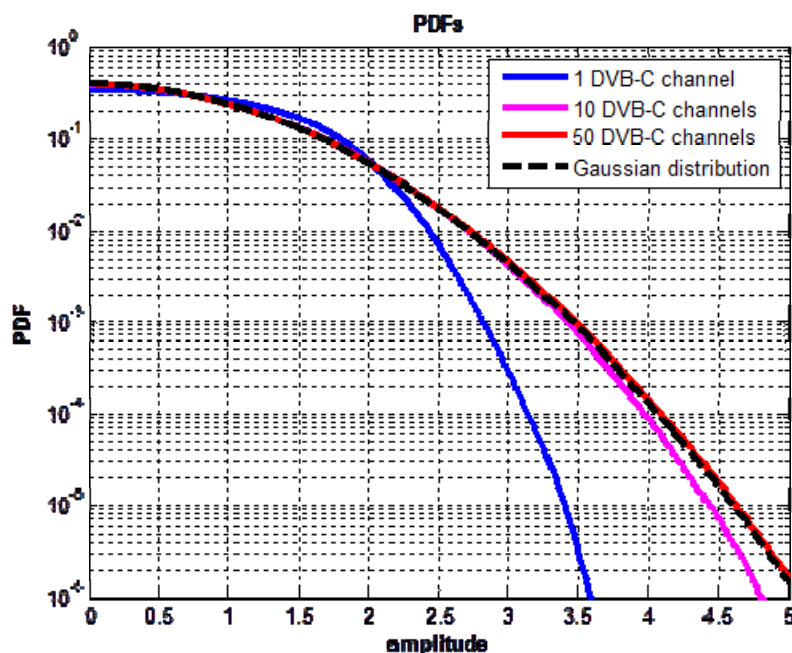


Figure 44: PDF function of 1,10, 50 DVB-C signals in a CATV network

In conclusion PAPR is not a relevant issue in case DVB-C2 signals are introduced in cable networks which are already carrying 10 and more DVB-C or DOCSIS® signals. This is the realistic scenario today in most of the CATV networks. In those networks the PAPR is today already about 13 dB and would not change with the added DVB-C2 signals. In the long term it can be assumed that DVB-C will be completely replaced by DVB-C2. In such a scenario it can be assumed that many (more than 20) uncorrelated DVB-C2 signal will be transmitted in cable networks. Here again the PDF will follow a Gaussian characteristic and the PAPR will be about 13 dB. In those scenarios the usage of the PAPR reduction mechanism in DVB-C2 (reserved tones) would not make sense and would only reduce the spectrum efficiency, as the reserved tones would not be available for carrying payload data.

However the mechanism for reduction of PAPR in DVB-C2 is not useless, as it may be utilized e.g. in case a single DVB-C2 signal is transmitted via a distribution system (e.g. via glass fibre). Simulations have shown that with appropriate algorithms for the generation of payload signals for the reserved tones a reduction of PAPR of about 3 dB is possible.

## 8.11 Concept and usage of the DVB-C2 modulator interface

### 8.11.0 Introduction

Cable network today usually provide bi-directional connectivity and therefore are capable to offer bi-direction services. However, traditional broadcasting services are playing an important role in the service offering of cable.

In case a traditional broadcasting service is targeted to be delivered to all customers connected to a network, it is self-evident that a cable operator will try to shift complex signal processing to the single centralized master headend in order to minimize the complexity of the signal processing in the local headends (see figure 45). Such effort may provide significant cost savings, especially in applications where there is a big number of local headends (hub / fibre node) in a cable network. DVB has therefore developed a modulator interface for DVB-C2, meeting those requirements and allowing significant cost savings and operational advantages for cable operators, applicable to any broadcasting type service. The technical specification has been standardized by ETSI TS 103 287 [i.25].

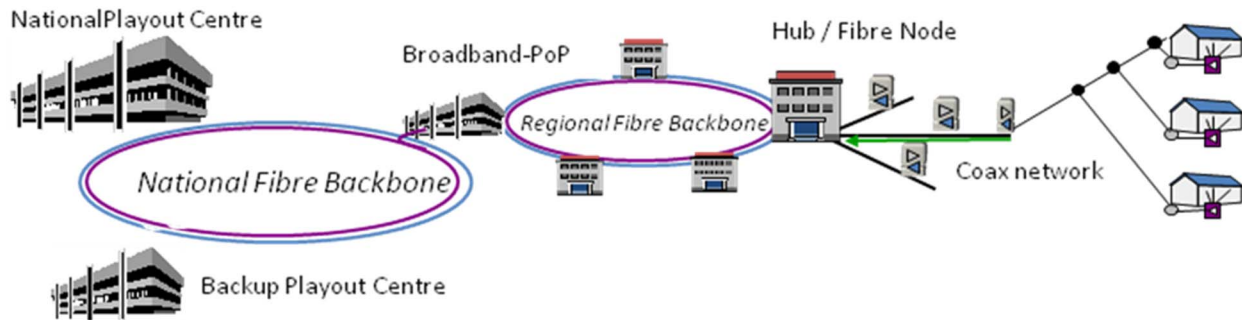
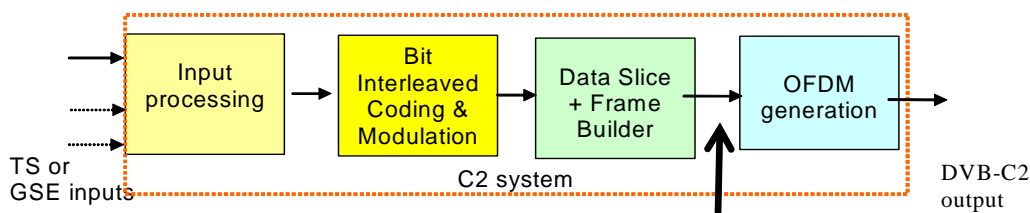


Figure 45: Example for the topology of a Hybrid Fibre Coax (HFC) cable network

### 8.11.1 Concept of the C2 Modulator Interface (MI)

The concept of the **C2 Modulator Interface** splits the modulator into a centralized part and an edge part (see figure 46). The block diagram does already give a clear indication, that complex processing blocks are located in the centralized part of the C2-modulator. As the DVB-C2 specification does provide significant levels of flexibility, it was a challenge for the design of the C2-MI to cover all technical DVB-C2 solutions the standards allows.



### DVB-C2 Modulator Interface

Figure 46: DVB-C2 modulator block diagram and location of the C2 Modulator Interface

The **C2-MI** supports any DVB-C2 compliant modulator configuration which includes the following options:

- Modulator output bandwidth up to 450 MHz
- Change of modulation parameters on a frame by frame basis
- Service specific modulation and robustness parameters
- Full flexibility of the 2 layer multiplexing structure
- Narrowband and broadband **notches**

The **C2-MI** specification defines a Time Division Multiplexing (TDM) frame, which allows delivering the payload of all subcarriers of a **C2 frame** to a **C2-MI** edge device. This TDM framing structure is optimized in terms of overhead.

### 8.11.2 Usage of the C2 Modulator Interface

In the starting phase of the DVB-C2 deployment the major application will be the transmission of broadcasting services encoded and multiplexed in a **Transport Stream** (single **PLP** and a single **Data Slice**) in a traditional frequency raster (8 or 6 MHz, depending on the country). In this case the **Transport Stream** can be encoded and multiplexed in a central master headend and delivered (e.g. via IP-multicast) to the different local headends via a fibre-based backbone network. Usually such backbone systems are able to deliver **Transport Streams** up to several hundred Mbit/s. This would allow extending the usage of the MI to **C2 systems** occupying several traditional cable channels (e.g. 64 MHz with up to 664,9 Mbit/s payload).

In case of single **PLP** configurations the **C2-MI** allows to use a special TDM schemes which reduces the overhead to about 15 %, whereas for multi **PLP** configuration application the overhead is in the range of 26,3 % to 45,1 %, depending on the chosen modulation schemes. For details see table 19 and table 20 in ETSI TS 103 287 [i.25].

The **C2-MI** specification [i.25] gives 4 examples of typical applications of the **C2-MI** interface in Annex B.

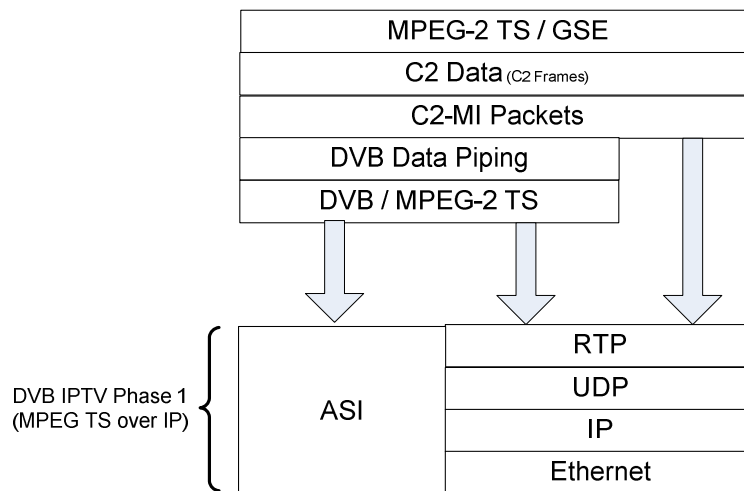
### 8.11.3 Usage of the fragmentation mechanism

In case the payload of a **C2 system** exceeds the maximum capacity of a channel of the backbone system, the payload frames shall be fragmented into  $n$  fragments, characterized by the carrier number of the lowest subcarrier and the carrier number of the highest subcarrier associated to that fragment of a **C2 system**. A **C2-MI** system, capable to handle fragmentation will be able to recognize the different fragments and to reassemble to complete payload of that **C2-System** in the **C2-MI** edge device.

**NOTE:** It is recommended that **C2-MI** systems supporting fragmentation are labelled as class IIBN (see Annex C of ETSI TS 103 287 [i.25]).

### 8.11.4 Transport of **C2-MI** packets

The **C2-MI** specification offers three options for the transport of **MI** packets over cable backbone systems to the **C2-MI** edge devices. Figure 47 shows the related protocol stack. Two solutions are based on the **MPEG-2 Transport Stream**. The first can be used in cable systems where the Asynchronous Serial Interface (**ASI**) is used as the standard interface to the backbone system for signal distribution. The second option fits for backbone configurations where **IP** multicast is used. For backbone systems which do not require a **MPEG-2 TS** layer a simplified **IP** transport mechanism is available. For more details see clauses 4.2 and 8 of ETSI TS 103 287 [i.25].



**Figure 47: The C2-Modulator Interface protocol stack**

### 8.11.5 Reserved tones for PAPR reduction

The **DVB C2-System** provides the possibility to reduce the Peak-to-Average Power Ratio (**PAPR**) of the output signal of a **DVB-C2** modulator by means of reserved tones (for details see clause 9.7 in ETSI EN 302 769 [i.1]). Usually the reduction of **PAPR** is not necessary in fully digital cable networks (mix of **DVB-C** and **DVB-C2** signals), as discussed in clause 10.8 in the present document. However, in case **PAPR** is needed in a configuration where the **C2-MI** is used, the modulation parameters of the reserved tones shall be defined in the **C2-MI** edge device, as it is only possible to detect the relevant output level characteristics at the edge side.

## 9 Network

### 9.0 Introduction

The network provides the transport and distribution capability to deliver the **DVB-C2** and other signals to a number of customer receivers. Apart from the conveyance of the signals, the network will add distortion signals like thermal noise from the amplifiers, echo and intermodulation products associated with the non-linear behaviour of the amplifiers and optical transmitters. To warrant good services, the network and the composite signal load should be designed for a sufficiently strong signal level and an appropriate low distortion signal level for all signals and at all homes.

The network design and the signal load and signal quality concerns a business trade off targeted at a most economical delivery of the broadcast and narrowcast services. This trade off results in a maximization of the network load in terms of the number of carriers and of the signal level. In practise the network load is limited by the non-linear character of the active components; the network will be operated close to the level of overloading of the active components.

DVB-C2 will be added as a new digital transmission technology next to DVB-C and analogue TV and FM radio. DVB-C2 supports different modulation and error protection schemes, each requiring an appropriate signal level and signal quality. From the viewpoint of costs, operators will be inclined to apply the modulation scheme with the highest capacity per channel, thus providing a further stimulus for a high composite signal level. The network challenge of DVB-C2 concerns the appropriate maximum DVB-C2 signal level that yields a satisfactory trade off between the network capacity in terms of the number of analogue and of digital channels at a specific modulation scheme and the quality of these signals in terms of signal level and signal-to-noise ratio and carrier to interference-plus-noise ratio (SNR and CINR).

## 9.1 Components of a cable network

### 9.1.0 Introduction

The network consists of an ensemble of active components, splitters, multitaps and fibre and coaxial cables connecting the DVB-C2 transmitter in the head end and the DVB-C2 receiver in the customer home. The network can be split into the operator's network part and the customer in-home network part with the network termination outlet as a demarcation point of both domains. From the perspective of the operator the wall outlet is considered as the system outlet. The operator will warrant a minimum signal quality delivered at the system outlet.

As a rule, the operator will deliver signals at the system outlet with a sufficient but limited margin in terms of signal level, SNR and echo to convey the signals in the customer's home. This margin is a business choice of the operator. Operators may deliver a high quality signal that allows passive in-home distribution to many receivers like analogue TV sets, STBs, IDTVs and cable modems.

#### 9.1.1 The operator part of the network

The operator part of the network should be constructed and maintained so that all signals at all homes have an appropriate signal level. The required signal characteristics for FM radio, analogue TV and DVB-C signals, including the minimum and maximum values, are specified in IEC 60728-1 [i.18]. Requirements for DVB-C2 have not been defined so far.

#### 9.1.2 The customer part of the network

The in-home coaxial network of a customer may range from a single coaxial lead directly connecting a customer receiver like an analogue TV set, STB, IDTV or home gateway up to an extended coaxial network with branching point and coaxial cables to different rooms with or without a home amplifier. Installations and components may range from high quality down to poor. Cabling and connectors with inferior shielding and low quality ohmic splitters are commonly used. Coaxial cables are not always connected to a receiver port or properly terminated with an impedance of 75  $\Omega$ . Often, the cumulative attenuation of splitters and coaxial cables is rather high. In particular the poor in-home networks may deteriorate the signal level and signal quality.

## 9.2 Distortion signals

### 9.2.1 Echo

#### 9.2.1.0 Introduction of standardized echoes profiles

Echo is caused by reflection of the signals at impedance transitions occurring at connectors, components, damaged cables, incorrectly (or not) terminated cables. For the development of the DVB-C2 standard [i.1], the worst-case echo as specified by the IEEE 802.14 [i.17] model has been assumed as a reference. This IEEE 802.14 [i.17] worst-case echo specification is shown figure 48. To validate this echo requirement, the echo in two live networks and the impact of the customer in-home network have been studied. The results of this study are briefly summarized below; the full study can be found in [i.20].

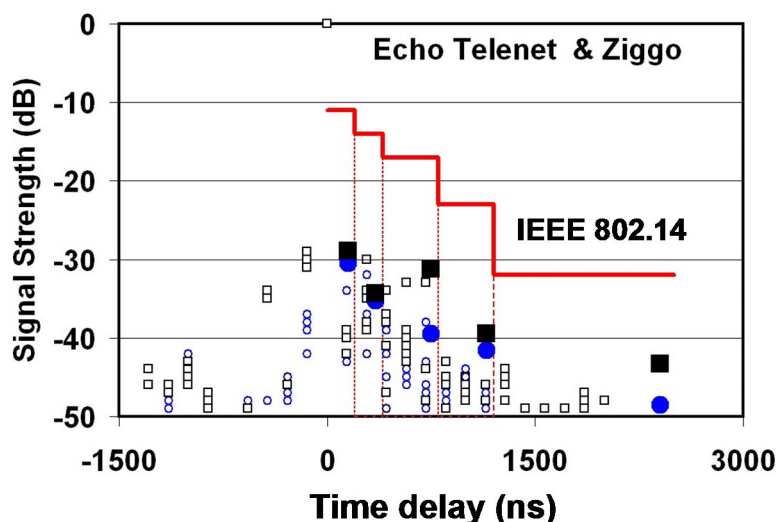


Figure 48: Worst case echo

NOTE 1: Figure 48 shows the worst case echo as specified in the IEEE 802.14 [i.17] channel model (solid red line) and the echo as measured at several end amplifiers in the networks of Ziggo (NL, cascades of 2 amplifiers) and Telenet (B, cascades of 11 amplifiers).

NOTE 2: The open symbols represent the delay and magnitude of the bins of the measurements. The solid symbols indicate the composite signal power of all the bins that fall within a delay interval of the IEEE 802.14 [i.17] mask.

### 9.2.1.1 Echoes caused by the operator network

In the networks of Ziggo (NL, cascades of 2 amplifiers) and Telenet (B, cascades of 11 amplifiers) the echo was measured at 4 multitaps using a DVB-C analyser [i.17]. The signal delay and magnitude is indicated in the open symbols of table 19. To compare the echo with the IEEE 802.14 mask [i.17], the signal power of all bins with positive and negative delay within the delay ranges of the IEEE 802.14 [i.17] mask were calculated (solid circles and squares). The result shows that in these networks the echo does not exceed the IEEE 802.14 mask [i.17]. In fact there is a margin of about 10 dB.

To create a reflexion in the forward direction, a backward and a next forward reflection is needed. In addition, the signal has to travel backward and forward between the two reflection points. In total the reflection suffers a signal loss equal to twice a reflection loss plus the attenuation of passing the coaxial cable twice. IEC 60728-1 [i.18] provides minimum requirements for reflection loss of passive components of 18 dB for low frequencies up to 10 dB for high frequencies. Most modern components have significantly better loss figure for the high frequencies. Table 19 shows some estimates of the signal loss and delay for cable segments of different length and for high and low frequencies. The estimates demonstrate that the echo contributed in the HFC network is roughly 10 dB better than required by the IEEE 802.14 [i.17] echo mask.

Table 19: Loss and delay of echoes

Length segment [m]	250		50		10		2	
Frequency MHz}	140	800	140	800	140	800	140	800
Reflection loss (2x) [db]	36	20	36	20	36	20	36	20
Attenuation @ 3 dB/100 m (200 MHz) [db]	15	30	3	6	0,6	1,2	-	-
Total Loss [dB]	51	50	39	26	37	21	36	20
Delay [ns]	2 500		500		100		20	

### 9.2.1.2 Echoes caused by the in-house network of the customer

Inferior in-home coaxial networks may severely degrade the echo, in particular when resistive splitters are used and coaxial cable is not properly terminated. However, echo measurements showed that even in such inferior in-home networks the echo still does not exceed the IEEE 802.14 [i.17] mask.



## 9.2.2 Ingress

### 9.2.2.1 Terrestrial broadcast services

Ingress of DVB-T signals is known to occur in the vicinity of the DVB-T transmitters. The use of high quality leads and passive components is an effective remedy.

### 9.2.2.2 Human activity in the home environment

In home activity such as switching on and off and plugging in- and out of electric equipment may cause severe burst events. The impact of such burst events on the DVB-C2 performance is not yet known.

### 9.2.2.3 Mobile services (Digital Dividend)

The EC intends to allocate the 790 MHz to 862 MHz frequency band for future mobile communication services (e.g. using the fourth generation mobile technology: LTE). If this allocation is adopted, mobile transmission in the in-home environment will severely distort the DVB-C2 signals. The impact of this use currently is topic of study.

## 9.2.3 Nonlinear behaviour of components

### 9.2.3.0 Introduction

In cable networks the amplifiers operate at a high output level. Because of this high output signal level, the non-linear behaviour of the amplifiers becomes apparent. Commonly in the field of HFC engineering, the component is assumed to behave according to the non-linear response function:

In the current practice only 2<sup>nd</sup> and 3<sup>rd</sup> order intermodulation is taken into consideration. Commonly these are referred to as the CSO and CTB beats. IEC 60728-1 [i.18] provides a detailed and complete description of the measurement of the second and third order intermodulation clusters (the CSO and CTB beats) at a series of frequencies when applying a load of un-modulated carriers. Next the carrier-to-intermodulation ratio (CIR) is calculated for all measurement frequencies, and for both the CSO beats ( $CIR_{CSO}$ ) and the CTB beats ( $CIR_{CTB}$ ). This measurement is used to specify the maximum output level of active components. The  $CIR_{CSO}$  and  $CIR_{CTB}$  values at all carrier frequencies are measured at different carrier levels. The  $CIR_{CSO}$  and  $CIR_{CTB}$  levels are frequency dependent, and a minimum or worst-case value for both the  $CIR_{CSO}$  and  $CIR_{CTB}$  can be appointed for a specific carrier level. Next the component maximum output level is defined as the carrier level in dB $\mu$ V for which respectively a 60 dB worst-case  $CIR_{CSO}$  and  $CIR_{CTB}$  value is found. If it is assumed that only 2<sup>nd</sup> and 3<sup>rd</sup> order intermodulation contributes to the worst-case  $CIR_{CSO}$  and  $CIR_{CTB}$  values.

In the DVB-C2 deployment scenarios the HFC network will carry a mixed load of analogue and digital services. The non-linear nature of the amplifiers results in the generation of intermodulation products of the analogue and digital carriers. The intermodulation products can be distinguished in three types:

- Narrow band intermodulation products, the CSO and CTB cluster beats.
- Broadband, random noise-like intermodulation products.
- Impulse noise.

### 9.2.3.1 Narrowband cluster beats

From the viewpoint of intermodulation, the analogue signal is dominated by the narrow band carrier; the broadband video signal and the FM audio signals have a lower spectral density and can be neglected. Because of the convolution, intermodulation of an analogue signal with another analogue signal produces a narrowband distortion signals (beats) at a number of well-defined frequencies. Intermodulation of all analogue carriers thus generates clusters of the beats at specific frequencies, the well-known cluster beats. Figure 49 shows the spectrum of the cable signals and distortion products for a cascade of amplifiers with a mixed load of PAL and digital carriers [i.19]. The figure is obtained from simulation using a 2<sup>nd</sup> and 3<sup>rd</sup> order component model. The narrowband cluster beats are shown in blue.

### 9.2.3.2 Broadband random noise

In contrast to the intermodulation of a (narrowband) analogue carrier with an analogue carrier, intermodulation of an analogue signal with a (broadband) digital carrier, either DVB-C or DVB-C2, and intermodulation of a digital carrier with a digital carrier yields a broadband, random noise like distortion signal. The red curve in figure 49 shows the broadband (random noise) distortion signal. This curve shows a periodic fine structure which reveals the period spectral density of the raster of 8 MHz DVB-C and DVB-C2 carriers. The simulation was performed for relatively low DVB-C and DVB-C2 signal levels, and therefore the broadband distortion signal is dominated by the intermodulation of the digital carriers with the analogue carriers; intermodulation of digital carriers with digital carriers is still negligible at these signal levels. The fine structure thus can be assigned to the convolution of period raster of analogue carriers with the periodic raster of digital carriers. For higher digital carrier levels, this fine structure becomes faint.

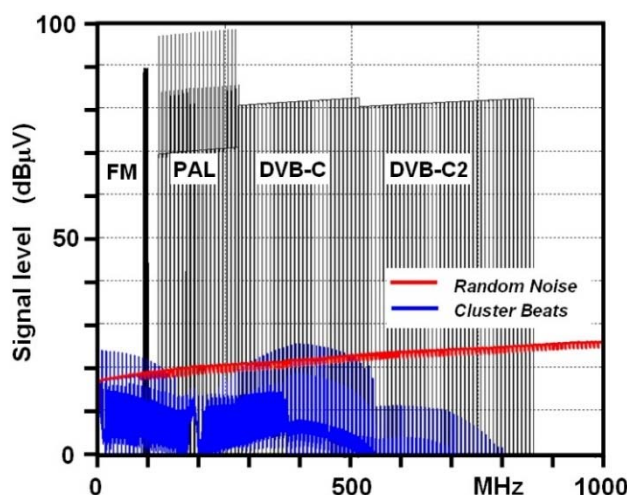


Figure 49: Spectrum of cable signals and intermodulation products

NOTE: Figure 49 shows the spectrum of cable signals and intermodulation products for a cascade of amplifiers with a mixed load of Pal and digital carriers. The figure obtained from simulation using a 2<sup>nd</sup> and 3<sup>rd</sup> order component model.

### 9.2.3.3 Impulse noise

At high composite loads, impulse events start occurring [i.20]. The generation of impulse events is demonstrated in figure 50. Figure 50 shows the probability density function (PDF) of a baseband real time distortion signal sample as recorded with a fast capturing system. The RF distortion signal is from a single amplifier (left window) and a cascade of 8 amplifiers (right window) with a composite load of 96 digital carriers and had a centre frequency of 423 MHz and 5 MHz bandwidth. At a low carrier level P1, the PDF matches a Gaussian distribution. At higher carrier levels P2 and P3, the PDF reveals a tail reflecting the occurrence of impulse events. Statistical analysis showed that these impulse events have a random time distribution and a peak duration of about 100 ns associated with the 5 MHz bandwidth resolution of the RF capturing system.

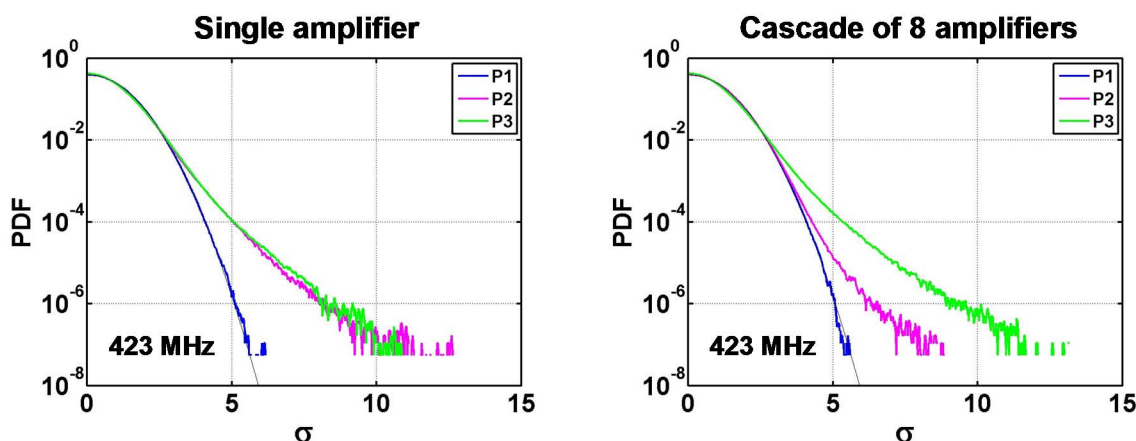


Figure 50: Probability density function (PDF) of the distortion signal caused by amplifiers

NOTE: Figure 50 shows the probability density function (PDF) of the distortion signal caused of a single amplifier (left window) and a cascade of 8 amplifiers (right window) with a composite load of 95 digital carriers. The PDF are recorded for a low carrier level (P1) where the non linear behaviour is not notable, at a high carrier level (P2) and at a very high carrier level (P3). All curves are normalized to the average distortion signal power. The grey curve shows the PDF for random noise (Gaussian distribution). The deviations at P2 and P3 are associated with impulse events.

## 9.3 Signal Requirements

### 9.3.0 Introducion

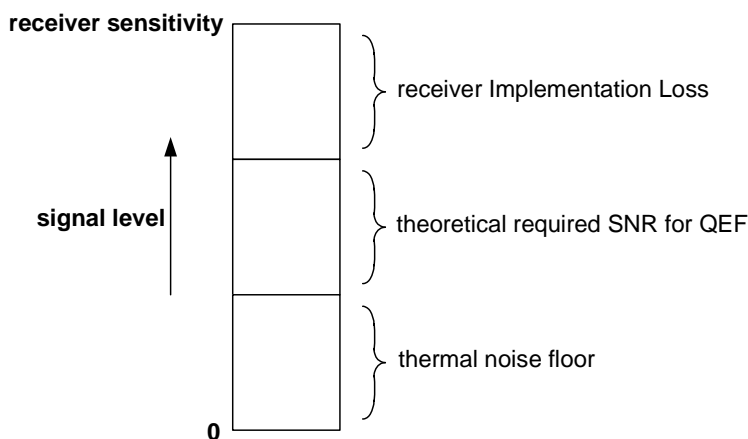
The networks will carry a mixed network load of FM radio, analogue TV (PAL or SECAM), DVB-C and DVB-C2 signals. For all services the appropriate signal levels and the signal quality levels shall be warranted.

### 9.3.1 Signal levels

The minimum signal levels at the system outlet for FM radio, analogue TV and DVB-C are specified in IEC 60728-1 [i.18], clause 5.4.

Currently, IEC 60728-1 [i.18] provides no requirements for the DVB-C2 signal level at the cable system outlet. A guideline for the DVB-C2 signal level can be derived from the sensitivity figure of a DVB-C2 receiver and by allocating a maximum signal loss associated with the in-home coaxial network between the system outlet and the DVB-C2 receiver. Since the DVB-C2-based digital TV service is intended as an addition to the existing analogue TV services, an operator may base his service concept on the scenario that digital TV will be watched on one TV set in the living room while the analogue services can be watched elsewhere in the home. In this scenario, the coaxial branch between the system outlet and the DVB-C2 receiver will consist of a splitter and a short coaxial lead which typically corresponds with some 7 dB loss.

The receiver sensitivity is composed of the thermal noise floor, the minimum signal-to-noise ratio for quasi error-free reception and the implementation loss and follows from a straightforward power budget calculation as illustrated in figure 51.



**Figure 51: Budget calculation for receiver sensitivity**

The thermal noise floor of broadband cable technologies with an 8 MHz channel width is about 4 dB $\mu$ V. The required signal-to-noise ratio for quasi error free reception depends on the specific DVB-C2 modulation and protection schemes as given in table 22 and figure 78 in clause 11.2 of the present document. Based on the implementation of the state-of-the-art DVB-C receivers, the implementation margin is estimated 11 dB for the DVB-C2 1 024 and lower QAM modulation modes and 12 dB for the 4 096-QAM modulation mode. Table 19 lists the breakdown and the minimum DVB-C2 signal level.

**Table 20: Estimated minimum DVB-C2 signal level at the system outlet**

Modulation	Code Rate	Noise Floor (dB $\mu$ V)	SNR (dB)	Implementation margin (dB)	In-home margin (dB)	Minimum signal level (dB $\mu$ V)
16-QAM	4/5	4	<b>10,7</b>	10	7	<b>31,7</b>
	9/10	4	<b>12,8</b>	10	7	<b>33,8</b>
64-QAM	2/3	4	<b>13,5</b>	10	7	<b>34,5</b>
	4/5	4	<b>16,1</b>	10	7	<b>37,1</b>
	9/10	4	<b>18,5</b>	10	7	<b>39,5</b>
256-QAM	3/4	4	<b>20,0</b>	11	7	<b>42,0</b>
	5/6	4	<b>22,0</b>	11	7	<b>44,0</b>
	9/10	4	<b>24,0</b>	11	7	<b>46,0</b>
1 024-QAM	3/4	4	<b>24,8</b>	11	7	<b>46,8</b>
	5/6	4	<b>27,2</b>	11	7	<b>49,2</b>
	9/10	4	<b>29,5</b>	11	7	<b>51,5</b>
4 096-QAM	4/5	4	<b>31,4</b>	12	7	<b>54,4</b>
	5/6	4	<b>32,4</b>	12	7	<b>55,4</b>
	9/10	4	<b>35,0</b>	12	7	<b>58</b>

## 9.3.2 Signal quality requirements

### 9.3.2.0 Introductory remarks

The required quality of the signals at the system outlet for FM radio, analogue TV and DVB-C are specified in IEC 60728-1 [i.18].

### 9.3.2.1 Analogue TV

The carrier-to-noise ratio ( $C/N$ ) requirements for analogue TV is specified in IEC 60728-1 [i.18], clause 5.8.

The carrier-to-intermodulation ratio ( $C/I$ ) is defined as the ratio between the carrier signal and the weighted sum of the CSO and CTB cluster beats measured as specified in IEC 60728-1 [i.18], clause 4.5.3. The requirement is specified in IEC 60728-1 [i.18], clause 5.9.3.

### 9.3.2.2 FM radio

The carrier-to-noise ratio ( $C/N$ ) is specified in IEC 60728-1 [i.18], clause 5.8.

### 9.3.2.3 DVB-C

The minimum composite intermodulation noise ration (CINR) is specified in IEC 60728-1 [i.18], clause 5.8.

### 9.3.2.4 DVB-C2

For DVB-C2 currently IEC 60728-1 [i.18] does not provide the signal quality requirements. However, DVB-C2 has been designed and optimized for the existing HFC networks. The simulation scenarios used for the system validations have been defined with the minimum network requirements as specified in IEC 60728-1 [i.18] in mind. This includes:

- **Echo**

The echo should not exceed the IEEE 802.14 [i.17] specification, see clause 9.2.1.

- **Narrowband interference**

An operator should expect 3 narrow-band cluster-beats per 8 MHz channel when analogue services are transmitted. Each cluster-beat has a bandwidth of 30 kHz with high signal levels, as shown in figure 51. However, only few OFDM subcarriers will be affected by this noise.

In a special case it may be assumed that the power spectral density of the narrow-band interferer is significantly higher than the useful signal. Thus, the affected OFDM subcarriers do not carry any useful information for the decoding process and can be treated as **notches**. Figure 52 depicts a simulation where 44 subcarriers ( $44 \cdot 2,232\text{kHz} \approx 100\text{kHz}$  for 8 MHz channel raster) have been notched (see figure 48). In case of the applied modulation parameters 256-QAM and LDPC code rate 3/4 approximately 0,15 dB increased SNR is required for error-free reception. Furthermore, the influence of omitting the use of the frequency interleaver is also shown in this figure 53. Here, approximately 0,1 dB would have to be added additionally to the minimum required SNR value for error free reception.

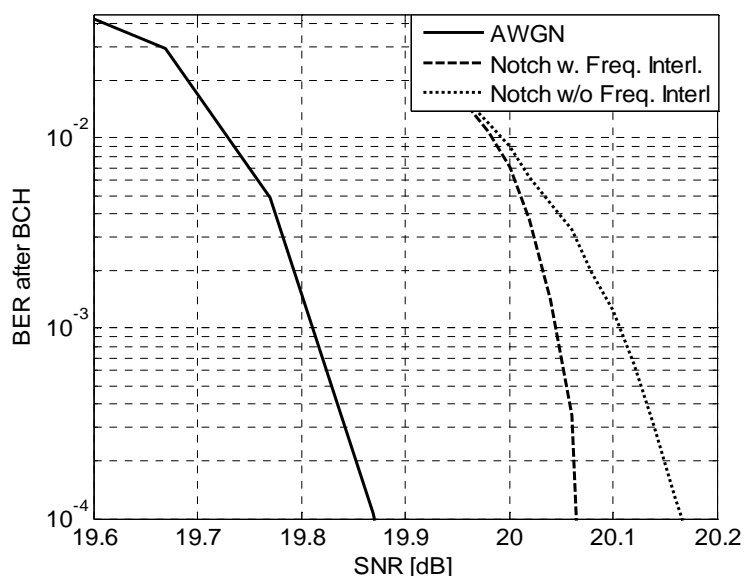


Figure 52: Impact of a narrowband interferer (100 kHz bandwidth) on the SNR requirements of a DVB-C2 signal (8 MHz bandwidth), with and without frequency interleaving

- SNR

First minimum requirements for the SNR for DVB-C2 are obtained from system calculations as shown in table 22 of clause 11.2. However, in case of high network loads this minimum SNR may turn out too optimistic due to the possible occurrence of impulsive noise.

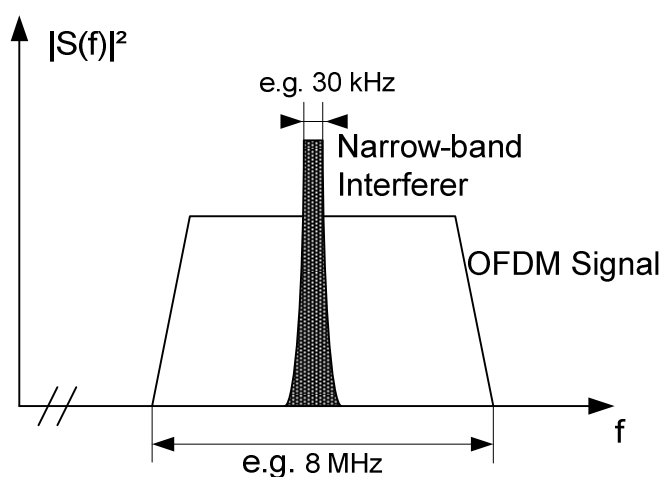


Figure 53: DVB-C2 signal with narrow-band interferer (cluster beat), only few OFDM subcarriers will be affected

## 9.4 Network optimization

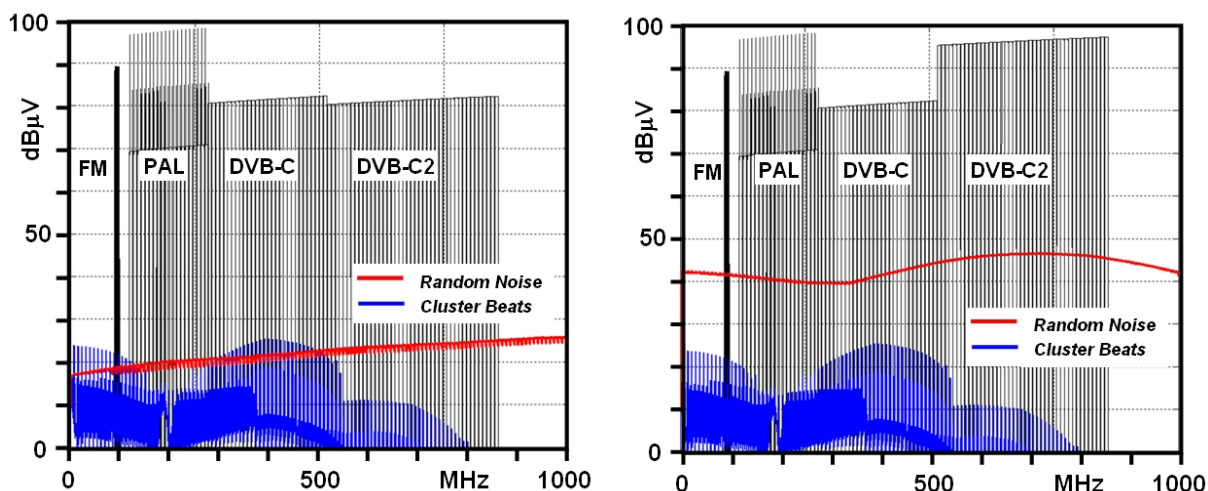
### 9.4.0 Introduction

Operators will have to establish the optimum DVB-C2 carrier level for their networks, which allows robust and error free DVB-C2 transmission at a sufficiently high modulation profile whereas at the same time the other services (FM radio, analogue TV and DVB-C) are sufficiently protected.

The following subclauses of clause 9.4 summarize and explain the issues relevant in network optimization for DVB-C2 deployment. In particular the optimization of the DVB-C2 signal level is addressed.

#### 9.4.1 The effect of the DVB-C2 carrier level

Figure 54 gives an illustration of the impact of the new DVB-C2 services on the nature and level of the distortion signals. When deploying DVB-C2, an operator will not raise or lower the signal level of the existing FM radio, analogue TV services and the DVB-C. Instead these signal levels will be conserved. For DVB-C2 instead, the operators faces the problem of determining the appropriate signal level. The higher the signal level, the higher the throughput. Figure 54 shows the level of the intermodulation products for a low DVB-C2 signal level (left panel) and a high DVB-C2 signal level (right panel).



**Figure 54: Full spectra at the output port of the distribution amplifier of a cascade**

**NOTE:** Figure 54 shows the full spectra at the output port of the distribution amplifier of a cascade. The random noise (thermal noise and broadband intermodulation products) and the narrowband composite cluster beats are respectively shown in red and blue. The left figure shows a simulation for of a low DVB-C2 signal level, the right figure for a 16 dB higher DVB-C2 level. The signal levels in dBμV refer to the level as measured with a spectrum analyzer with 300 kHz bandwidth resolution. Thus the real DVB-C, DVB-C2 and random noise intermodulation signal levels are about 14 dB higher than shown.

Comparison of both the windows of figure 54 shows that raising the DVB-C2 level does not change the cluster beat spectrum (blue) whereas the random noise spectrum (red) is roughly 20 dB higher. Both observations are consistent with the earlier mentioned origin of the intermodulation distortion products, see clause 9.2.3. The analogue TV carrier level is not changed and the number has not changed, and so the number and magnitude of the narrowband cluster beats is conserved. In contrast, since the DVB-C2 carrier level is increased, a much higher signal level of broadband random-noise is found, which is associated with digital-analogue and digital-digital intermodulation.

The above analysis shows that when introducing digital services this will have no impact on the multiple frequency intermodulation interference (the CSO and CTB cluster beats) to the analogue TV services. It may contribute to the reduction of the carrier-to-noise ratio of analogue services.

## 9.4.2 Impact of the DVB-C2 signal level on DVB-C2 performance

Figure 54 schematically shows the impact of the DVB-C carrier level on the signal quality and performance of DVB-C services for a single component or a cascade with a *mixed analogue and digital load*. For a load with DVB-C2 carriers there is no information available yet; however, a qualitatively similar behaviour can be expected. Both the CINR and the bit error rate before interleaving and forward error coding are shown.

Based on this understanding, three ranges with a different distortion signal environment have to be distinguished in case of a mixed load of analogue and digital carriers, as indicated in figure 55:

### a) a low carrier level:

In this range the distortion signal is composed of the thermal noise of the amplifier(s) and the narrowband cluster beats generated by the intermodulation of the analogue TV services. As explained in clause 9.4.1, the number and amplitude of the cluster beats does not depend on the signal of the digital carriers. Raising the DVB-C signal level yields a proportional improvement of the CINR. The CINR curve in this range has a slope +1.

### b) a moderate carrier level:

In this transition region, the generation of broadband noise by the intermodulation of a digital carriers with analogue carriers (second and third order non-linear behaviour) becomes notable, but with no or small effect on the CINR. This broadband random noise is additive to the thermal noise of the amplifiers and the narrowband cluster beats of range A.

### c) a high carrier level:

At this carrier level broadband intermodulation products overrule the thermal noise of the amplifiers as demonstrated by a steep decline of the CINR curve in this range. As a rule, the CINR curve approaches an asymptote with slope -4 which shows that 5<sup>th</sup> order intermodulation products do dominate the distortion signal. In clause 9.4.4 a further explanation is given of the order of the dominant intermodulation products. Although the CINR for this high carrier level range still is sufficiently high for quasi-error free transmission in case of noise with a random noise power density distribution (arbitrary white Gaussian noise, AWGN), the bit error rate increases dramatically in this range. This high bit error rate level agrees with the occurrence impulse noise events.

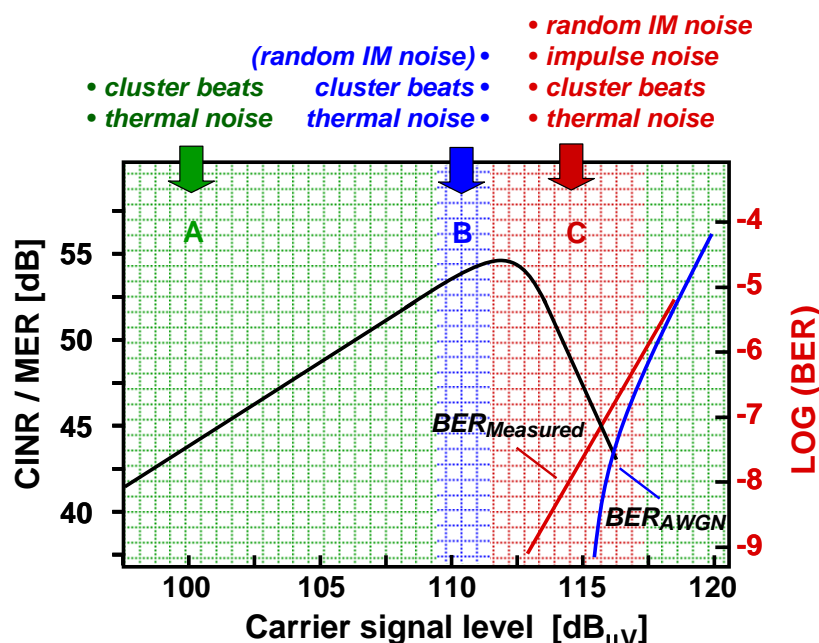


Figure 55: Schematic diagram explaining the nature of the quality of the signal of a DVB-C signal

NOTE: Figure 55 shows the schematic diagram explaining the nature of the quality of the signal of a DVB-C carrier for an amplifier or a cascade with a mixed analogue and digital load when increasing the carrier signal level. The top panel shows both the CINR and the bit error rate before interleaving and forward error coding. Additionally the figure indicates the nature of the distortion products present at the different carrier levels. The lower panel shows schematically shows the signal level of the intermodulation products and of the thermal noise for 8 MHz and 50 kHz measurement resolution. The carrier levels refer to the output of an amplifier.

Although the above understanding is based on data obtained for components with a load of DVB-C carriers, a comparable behaviour can be expected for DVB-C2 carriers. When considering the DVB-C2 signal level, operators shall be aware of the generation of the above intermodulation products.

Irrespective of the DVB-C2 signal level, the spectrum will contain narrow band clusters beats. These narrowband cluster beats will interfere with specific subcarriers of the DVB-C2 signal.

### 9.4.3 Impact of the DVB-C2 signal level on analogue TV services

Raising the DVB-C2 signal level above a specific value will increase the random noise distortion level and eventually result in the generation of impulse events. It does not affect the number and magnitude of the cluster beats. Therefore, higher DVB-C2 signal level will reduce the C/N of the analogue TV signals whereas the CINR is not affected.

The noise level in the C/N of the video carrier refers to the noise measured in the full TV channel, see IEC 60728-1 [i.18], clause 4.6. Therefore, this C/N will degrade only when the level of broadband intermodulation products approaches the thermal noise level; the C/N and CINR are related. Thus, the C/N of an analogue TV signal will start to degrade when the digital carrier level approaches the level of maximum CINR.

### 9.4.4 Non linear behaviour of active components in case of digital loads

This clause contains preliminary findings based on first simulations and measurement of higher order intermodulation interference effects in cable networks caused by digital signals. It will be subject to future revisions of the DVB-C2 implementation guidelines to further elaborate on the impact of non linear distortions in cable networks caused by digital signals.

In the existing understanding of signal degradation associated with non-linear behaviour of components, generally only 2<sup>nd</sup> and 3<sup>rd</sup> order effects are considered. However, analysis of degradation data strongly suggests that not the 2<sup>nd</sup> and 3<sup>rd</sup> order nonlinear behaviour degrades the digital signal, but 4<sup>th</sup> and 5<sup>th</sup> order effects. This hypothesis is based on the following three observations and arguments [i.21]:

- 1) Most of the CINR curves of a component with a digital load of 96 DVB-C carriers have an asymptote with slope -4 associated with 5<sup>th</sup> order intermodulation for high carrier levels. In addition, as a rule only a limited transition range from the low carrier level part with slope +1 is seen, and without indications for intermediate regions with 2<sup>nd</sup> or 3<sup>rd</sup> order dominance. Figure 56 gives two samples of such CINR curves obtained from two different amplifiers with a load of 96 digital carriers. The curves were measured with 8 MHz bandwidth resolution. Next to the measured curves, the figures show the CINR curves from simulations using a 2<sup>nd</sup> and 3<sup>rd</sup> order component model.
- 2) In case of a load of unmodulated carriers, the CINR curves for the CSO and CTB beats can be recorded separately and with a measurement resolution of 50 kHz. These curves respectively do show the ranges with dominant 2<sup>nd</sup> and 3<sup>rd</sup> order intermodulation. An example of  $CINR_{CSO}$  and  $CINR_{CTB}$  curves and the occurrence of dominant 2<sup>nd</sup> and 3<sup>rd</sup> order degradation can be found in figure 53.



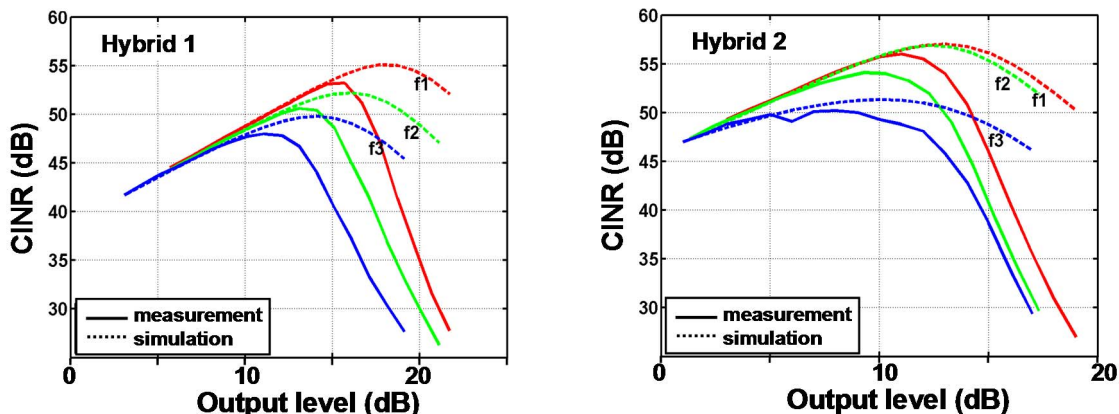


Figure 56: Measured and simulated CINR curves for 119 MHz (f1), 420 MHz (f2) and 855 MHz (f3)

NOTE 1: Figure 56 shows the measured and simulated CINR curves for 119 MHz (f1), 420 MHz (f2) and 855 MHz (f3) as obtained for a component with a composite load of 96 digital carriers. The bandwidth resolution was 8 MHz. The measured curves show a high carrier level asymptote with a slope -4. For the simulation a 2<sup>nd</sup> and 3<sup>rd</sup> order component model is used. The simulated and measured curves clearly are not congruent, showing that 2<sup>nd</sup> and 3<sup>rd</sup> order intermodulation does not dominate the CINR curves.

- 3) The absence of visible or measureable 2<sup>nd</sup> and 3<sup>rd</sup> order degradation in case of digital carriers can be explained logically and straightforwardly, namely, as long as the 2<sup>nd</sup> and 3<sup>rd</sup> order intermodulation dominates the intermodulation distortion signal, these 2<sup>nd</sup> and 3<sup>rd</sup> order intermodulation products have a smaller signal power than the thermal noise generated by the component. To understand this point, it is helpful to compare the cases of a composite load of digital (broadband) carriers and of unmodulated and uncorrelated (narrowband) carriers, with the same number of carriers and the same average signal level. Thus both cases have a system load with a same composite signal power level, albeit composed of broadband signals and narrowband signals in the respective cases. Evidently, the signal power of the intermodulation products will be the same in both cases as well. However, the intermodulation products will be different in nature: the broadband load of digital carriers will generate broadband random noise more or less evenly distributed in the frequency domain. In contrast, in case of the load of unmodulated carriers, narrowband cluster beats are generated. Stated differently; in case of the broadband signals the distortion signal power is completely smeared out over the full frequency range whereas in case of the unmodulated carriers the distortion signal is concentrated in a limited number of cluster beats with high spectral power density. Taking the thermal noise of the amplifier into consideration, the broadband intermodulation signal level is below or equal to this thermal noise level whereas the cluster beats peak well above the thermal noise level, as illustrated in figure 57.

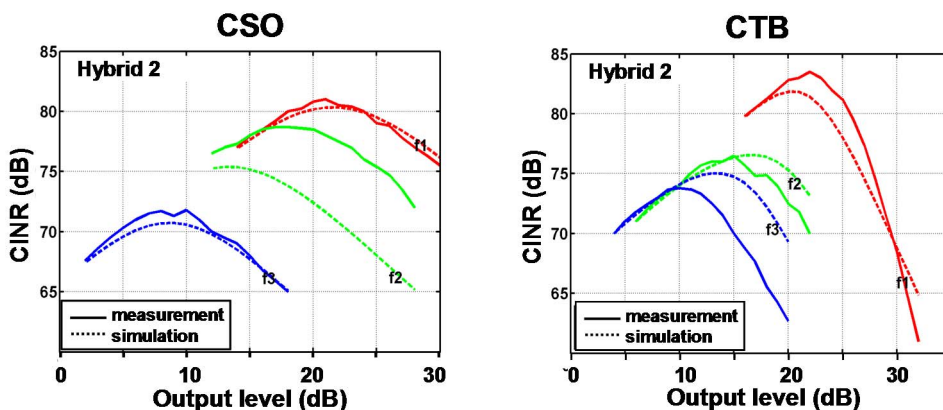


Figure 57: Measured and simulated CINR curves

NOTE 2: Figure 57 shows the measured and simulated CINR curves as obtained for a component with a (sloped) CENELEC load with 42 unmodulated carriers for 119 MHz (f1), 420 MHz (f2) and 855 MHz (f3). Measurement resolution was 50 KHz. The left panel shows the CINR for the CSO beats; the right panel shows the CINR for the CTB beats. For the simulation a 2<sup>nd</sup> and 3<sup>rd</sup> order component model is used. The measured curves show a high carrier level asymptote with a slope -1 and -2 for the CSO and CTB CINR curves respectively. The simulated and measured curves have congruent shapes, showing that indeed 2<sup>nd</sup> and 3<sup>rd</sup> order intermodulation dominates the CINR curves.

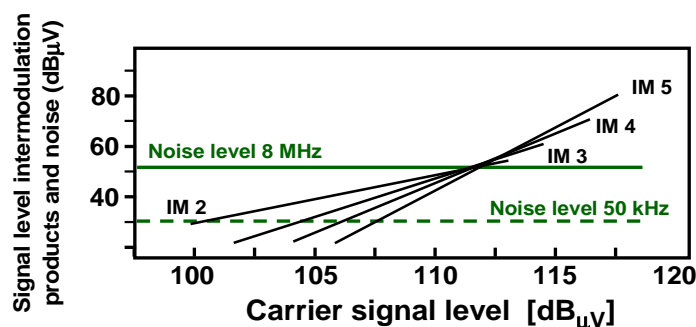


Figure 58: Schematic diagram of the signal level of the intermodulation products

NOTE 3: Figure 58 shows the schematic diagram of the signal level of the intermodulation products as a function of the carrier level. The noise levels are indicated for 50 kHz and with 8 MHz bandwidth resolution.

## 10 Receivers

### 10.1 Synchronization Procedure

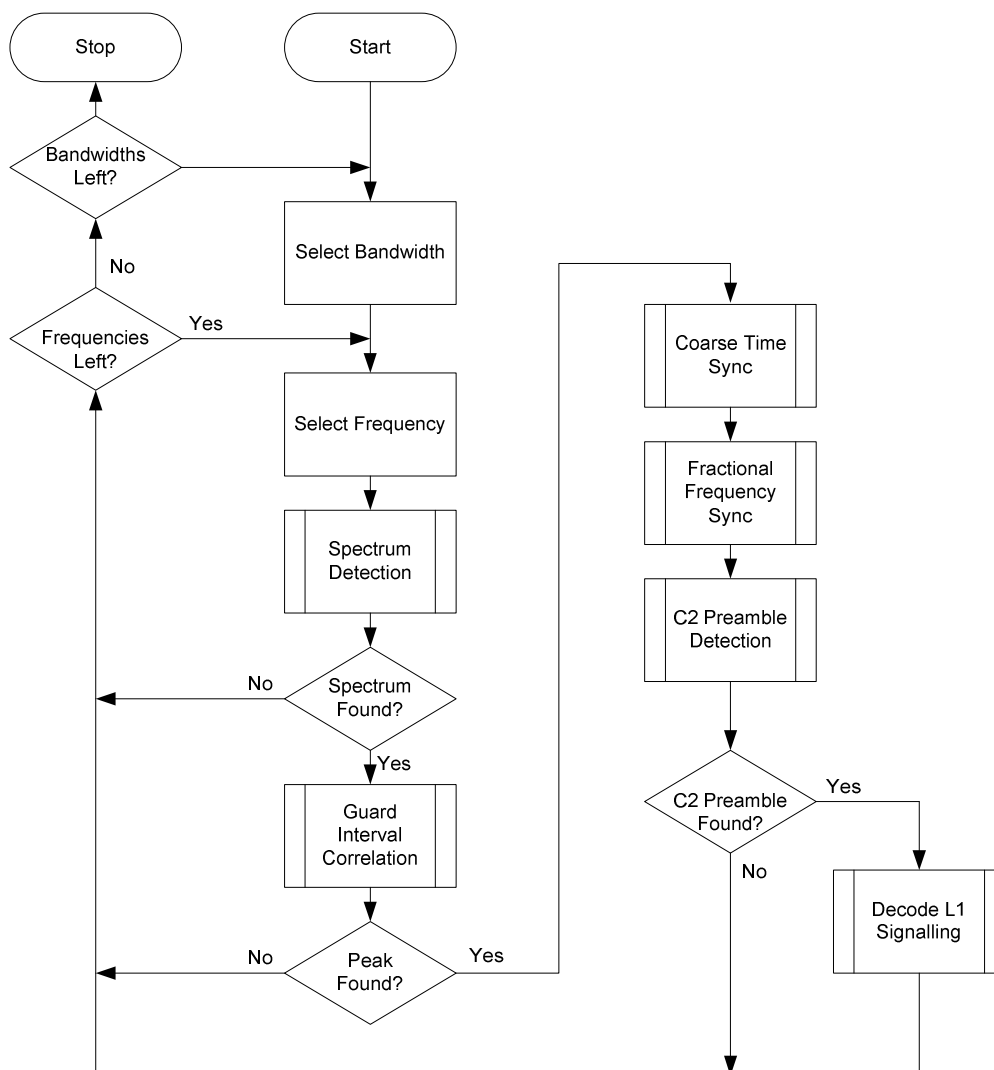
#### 10.1.0 Introduction

The DVB-C2 signalling scheme, consisting of Layer 1 and Layer 2 signalling, allows the receiver to acquire all relevant information required to tune to the targeted service. This clause describes procedures of the receiver to detect and process the relevant information.

#### 10.1.1 Initial Acquisition

##### 10.1.1.0 Initial Acquisition flow diagram

The initial acquisition is performed after the first switch-on of the receiver to detect the available DVB-C2 signals. The procedure works as depicted in figure 59. Details on the different blocks are given in the following clauses.

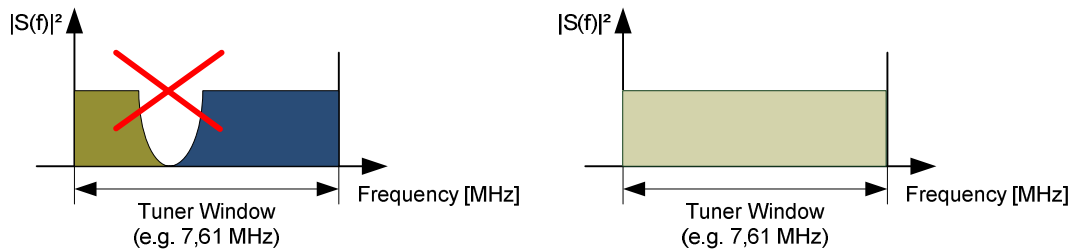


**Figure 59: Initial acquisition flow chart**

Firstly, the DVB-C2 receiver selects one of the possible signal bandwidth, i.e. 8 MHz or 6 MHz. Then, it chooses a possible DVB-C2 signal frequency and tries to detect whether a possible DVB-C2 spectrum is available within the tuner window. If a spectrum has been found, the receiver tries to evaluate if the signal is an OFDM signal. Next, the receiver tries to synchronize onto the OFDM signal and tries to find the DVB-C2 preamble. If the preamble has been found, it is decoded and the detection of the next DVB-C2 signal starts.

#### 10.1.1.1 Spectrum Detection

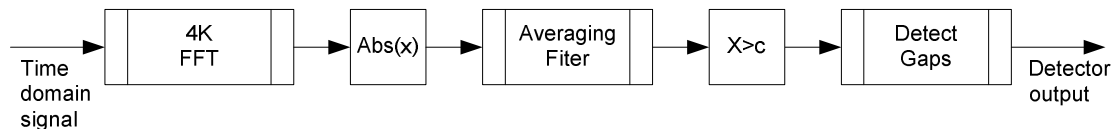
The spectrum detection is required to tune correctly to a DVB-C2 signal. In order to be able to decode the complete **L1-part2** signalling, the receiver shall be able to receive at least 7,61 MHz (or 5,71 MHz in the 6 MHz mode) of one DVB-C2 signal. It is especially required that the receiver does not try to decode the **L1-part2** signalling of two separate DVB-C2 signals. Additionally, the receiver should not try to decode a specific part of a DVB-C2 signal that included broadband **notches**.



**Figure 60: Principle of the spectrum detection, false spectrum (left hand side) and correct spectrum (right hand side)**

One means to overcome this problem is the application of spectrum detection. Figure 60 depicts this approach. The left spectrum has frequency areas in which no energy is transmitted. Hence, this signal contains a broad-band notch at this position or the receiver is tuned onto two separate signals. Consequently, the receiver has not found a correct spectrum and shall tune to another frequency.

A correct tuning position is the right figure. The spectrum does not contain any broad-band **notches**. Hence, the signal may be a valid DVB-C2 signal and the receiver shall continue the synchronization process. However, it has to be mentioned that this signal may naturally contain narrow band **notches**, which may be placed in each valid tuner window. Therefore, the receiver shall treat narrow-band **notches** like a DVB-C2 signal.



**Figure 61: Possible implementation of the spectrum detector**

A possible implementation of the spectrum detector is depicted in figure 61. The receiver uses the 4K-FFT block of its OFDM demodulator to transform the time domain signal into the frequency domain. Next, it calculates the absolute value of the different frequencies and uses a mean filter between adjacent frequencies. By means of a threshold the receiver tries to estimate if the frequencies are used or not. Lastly, a gap detector counts the frequency gaps and tries to estimate whether the tuning position contains a possible DVB-C2 signal without broadband **notches** or not.

#### 10.1.1.2 Guard Interval Correlation

Most signals within the cable environment are not OFDM signals. Therefore, the detection whether the signal is an OFDM signal is extremely useful. As the Guard Interval is a cyclic copy of the useful part of the OFDM symbol, the receiver can try to correlate the Guard Interval against the useful part. If a peak (or several peaks in consecutive OFDM symbols) is found, the receiver can assume that the signal is an OFDM signal with the assumed parameters. Details of the synchronization algorithm are given in [i.12].

#### 10.1.1.3 Coarse Time and Fractional Frequency Synchronization

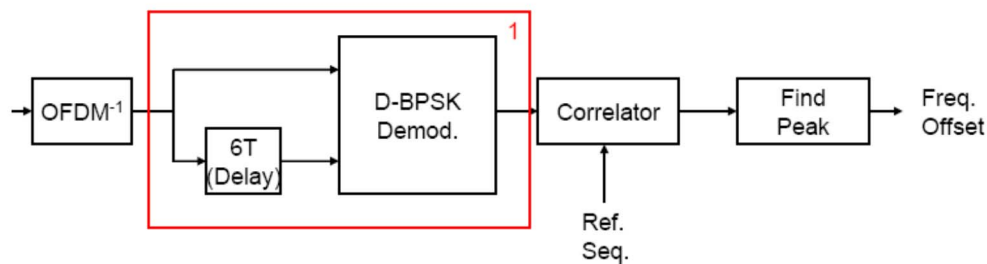
The coarse time and the fractional frequency synchronization can also be achieved by means of the Guard Interval. For details see [i.12].

#### 10.1.1.4 Preamble Detection and Synchronization

Within this clause it is assumed that the temporal synchronization to the OFDM symbols as well as the fractional frequency offset compensation has already been achieved. Thus, it is possible to demodulate the data modulated on each OFDM sub-carrier correctly, but a channel estimation cannot be performed and the actual carrier index  $k$  is unknown. If it is possible to detect the index  $k$  of the received OFDM sub-carriers, a complete synchronization with the data stream is achieved. Also the channel estimation is possible in this case, as the transmitted values of the pilots are known.

The synchronization to the preamble is shown in figure 62. The demodulation part (red box) already assumes a correct demodulation of the OFDM sub-carriers, but without channel estimation and knowledge of the absolute sub-carrier  $k$ . The block performs a DBPSK demodulation between two OFDM sub-carriers that have a distance of  $D_p=6$ , i.e. the distance of the preamble pilots. Assuming the channel conditions are nearly static between two pilots in the frequency direction, which is true due to the very short echoes in cable networks, the DBPSK demodulator outputs the Pilot Scrambling Sequence  $w_k^P$ , i.e. the modulation of the pilots before the differential encoding. Naturally, this sequence can only be calculated for the pilot positions, i.e.  $k$  is multiple of 6, while the output on the other positions depends on the signalling data will be random like.

In order to find the sequence, the output of the DBPSK demodulator is used within a correlator. The reference sequence is exactly the sequence  $w_k^P$  for the pilot positions in the desired frequency range. If  $k$  is not a pilot position the value of the sequence is assumed with 0. If the demodulated and the receiver-generated sequence are the same, a significant correlation peak occurs. By means of this peak, the receiver is able to estimate the offset in multiples of OFDM sub-carriers and correct it.

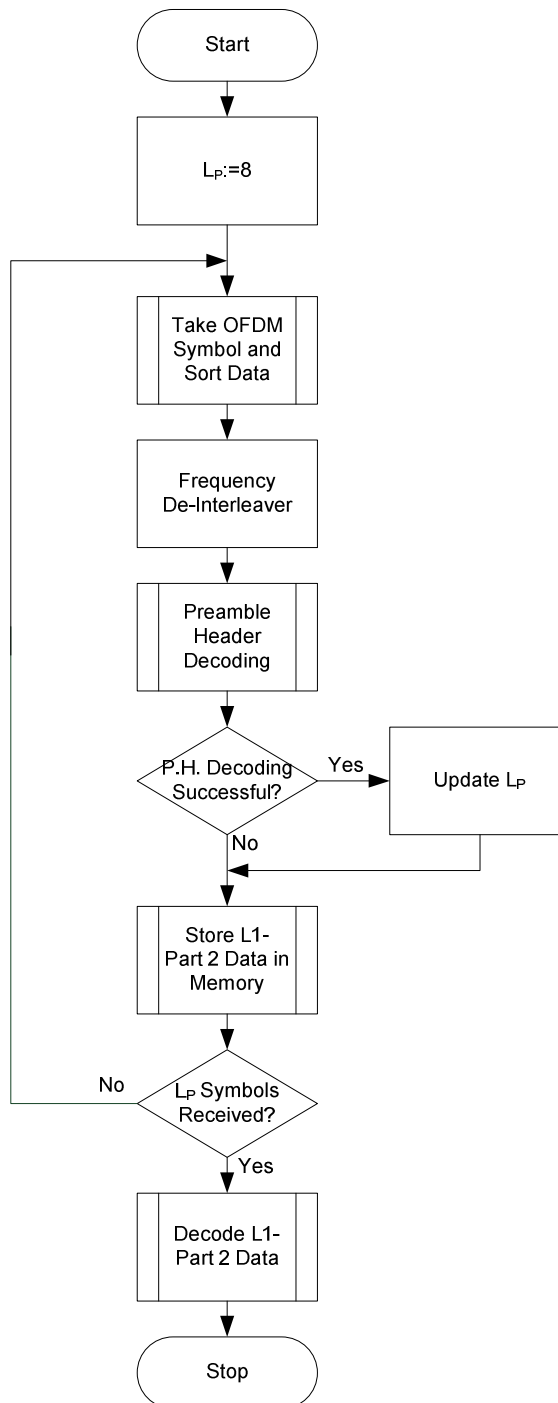


**Figure 62: Synchronization to the preamble by means of the preamble pilots**

### 10.1.1.5 Preamble Data Decoding Procedure

#### 10.1.1.5.0 Introduction

The information transmitted within the preamble is cyclically repeated within the L1 Signalling Blocks every 3 408 OFDM sub-carriers, or 7,61 MHz in 8 MHz operation. However, it is not ensured that the tuning window of the receiver front-end is aligned to an L1 Signalling Block. Additionally, the number of **preamble symbols**  $L_p$  is not known in advance. Therefore, the receiver may use the block diagram of figure 63.



**Figure 63: Preamble Decoding Procedure**

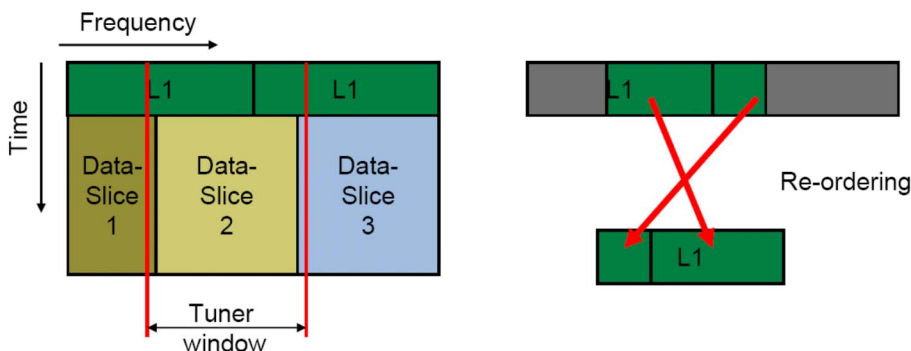
It is assumed that the receiver knows the position of the first **preamble symbol**. This is trivial if synchronization is already achieved, as the receiver knows the position of the previous preamble and the number of payload OFDM symbols  $L_F$ , which have been signalled in the previous preamble. If the receiver has tuned to the preamble recently, it will recognize the start of the preamble by means of the correlation as described in the clause above. In the very rare cases in which the receiver has tuned into the preamble and missed the first **preamble symbol**, the decoding procedure as depicted in figure 63 can be applied. Naturally, the decoding of the L1 - Part 2 signalling data will fail and the receiver has to wait for the next complete preamble.

At the beginning the receiver does not know the number of preamble OFDM symbols. Thus, he simply assumed the maximum number, which is  $L_p=8$ . Then, it takes the first **OFDM symbol** and sorts the sub-carriers and applies the frequency de-interleaving. Then it tries to decode the Preamble Header. Out of its parameters the receiver is able to calculate the correct number of preamble OFDM symbols. If the decoder was able to decode the data (see **Data Slice Packet Header** decoding), it can set the correct number of **preamble symbols**. If it was not able to decode it, it tries to decode it within the next OFDM symbol. When the number of preamble OFDM symbols is reached, the receiver decodes the L1 Signalling - Part 2 data.

The assumption of the maximum number of preamble OFDM symbols is required for increased robustness in case of preamble time interleaving. If e.g. the first **preamble symbol** is lost due to an impulsive noise event, the receiver does not know the length of the preamble. Hence, it has to try to obtain it during the following OFDM symbols. If it finds the parameters the decoding process may continue and will most probably be successful due to the time interleaving. On the other hand, if the preamble was built by a single OFDM symbol, the receiver will not be able to decode the **preamble header** in the next OFDM symbol, as it is already a **data symbol**. Hence, it will try to decode the maximum number of **OFDM symbol** for the preamble (which is 8). This will naturally fail, especially as the LDPC decoder will not converge. However, this should not be any problem, as the decoding process of the payload will fail anyway due to the missing signalling data, if no other means are used.

#### 10.1.1.5.1 Data Sorting

The tuning bandwidth of the receiver front-end is always wider than one L1 Signalling Block. Therefore, the receiver is able to receive the information within two blocks, and obtain the complete information by sorting of the data. Figure 64 shows an example. The receiver is optimally tuned to receive **Data Slice 2**. However, it is not aligned to the L1 Signalling Blocks. However, it is able to recover the complete information by sorting two blocks as shown in figure 64.



**Figure 64: Recovery of a L1 Signalling Block**

**NOTE:** Data-Slices do not have to be aligned with the L1 Signalling Blocks and therefore the tuning position does not need to be aligned with the L1 Signalling Blocks as well, information is obtained by sorting of the data of two partially received **L1-part2** Signalling Blocks.

One additional aspect is the presence of **notches** within the preamble, which may be required if no power shall be transmitted on specific frequencies. For this purpose the DVB-C2 specification distinguishes between narrow and broadband **notches**. The width of narrow-band **notches** is limited to few OFDM sub-carriers only. The loss of these few sub-carriers can be compensated by means of the FEC of the preamble easily.

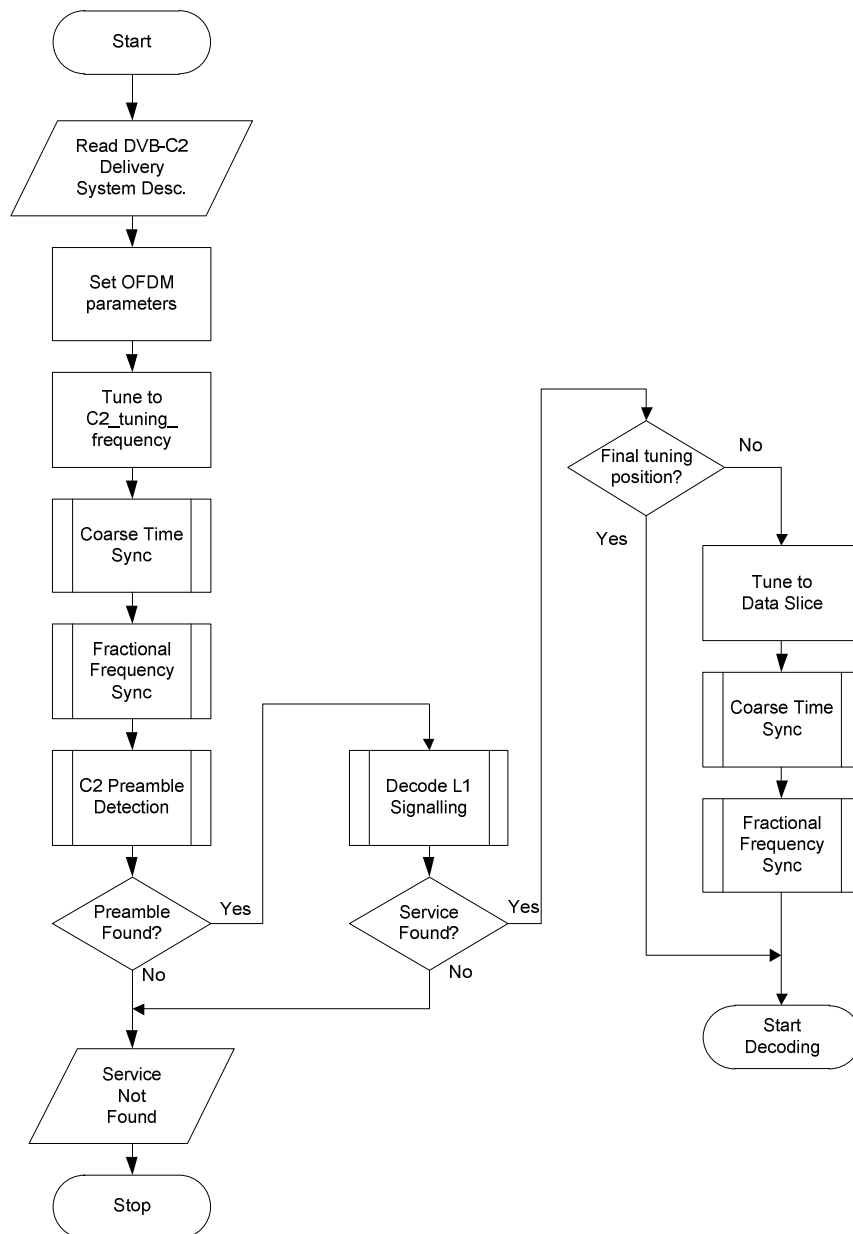
#### 10.1.1.5.2 Preamble Header Decoding

See **Data Slice Packet Header** decoding.

If the L1 block comprises several copies of **L1-part2** cells to fill the entire band of the L1 block, accumulation of those **L1-part2** cells improves the resultant SNR therefore decoding performance. In addition, if the preamble is composed of more than one OFDM symbol, the copy from each **preamble symbol** can also be used for the accumulation since all L1 headers carry the same information.

## 10.1.2 Channel Tuning Procedure

For the tuning to a specific service, the receiver has already the information of the *C2\_delivery\_system\_descriptor*, which is located in the NIT. This descriptor enables the mapping of a **Transport Stream** to the corresponding **Data Slice** (*data\_slice\_id*) and the **PLP** (*PLP\_id*) that carries this **Transport Stream** within the **C2 system**. For that purpose it includes the OFDM parameters, the tuning frequency to obtain the Layer 1 - part 2 signalling information.



**Figure 65: Channel tuning block diagram**

Figure 65 depicts the complete tuning procedure. Firstly, the system sets the OFDM parameters and the tuning frequency as described within the *C2\_delivery\_system\_descriptor* and then tries to synchronize onto the DVB-C2 stream. If the DVB-C2 signal (especially the DVB-C2 preamble) is not found, the service is no longer present and the tuning process is stopped. If the preamble is found the receiver decodes the corresponding Layer 1 - part 2 signalling (**L1-part2**). If the desired service is not present within the signalling, the service does not exist and the tuning process stops. If the service is present, the receiver recognizes from the *C2\_tuning\_frequency\_type* field if it has already tuned to its final **Data Slice** specific tuning position. If this is true it can immediately start to decode the service. Otherwise the receiver calculates the tuning position out of the Layer 1 - part 2 signalling (**L1-part2**) information, re-tunes to this position and starts to decode the desired service.



## 10.1.3 Preamble Time De-interleaver

### 10.1.3.1 Phase of time de-interleaving

As the L1 TI block is synchronized to preamble boundary in time direction, the receiver can begin preamble time de-interleaving after **C2 frame** detection. The receiver first detects the preamble and decodes L1 header to get information about preamble time interleaving parameters. Then, the receiver can immediately start de-interleaving with a de-interleaver buffer.

### 10.1.3.2 Pre-processing to time de-interleaving

The pre-processing from preamble synchronization to preamble time de-interleaving process is depicted in figure 62, which is a de-interleaving process as a counterpart to the time interleaving shown in figure 28 of ETSI EN 302 769 [i.1].

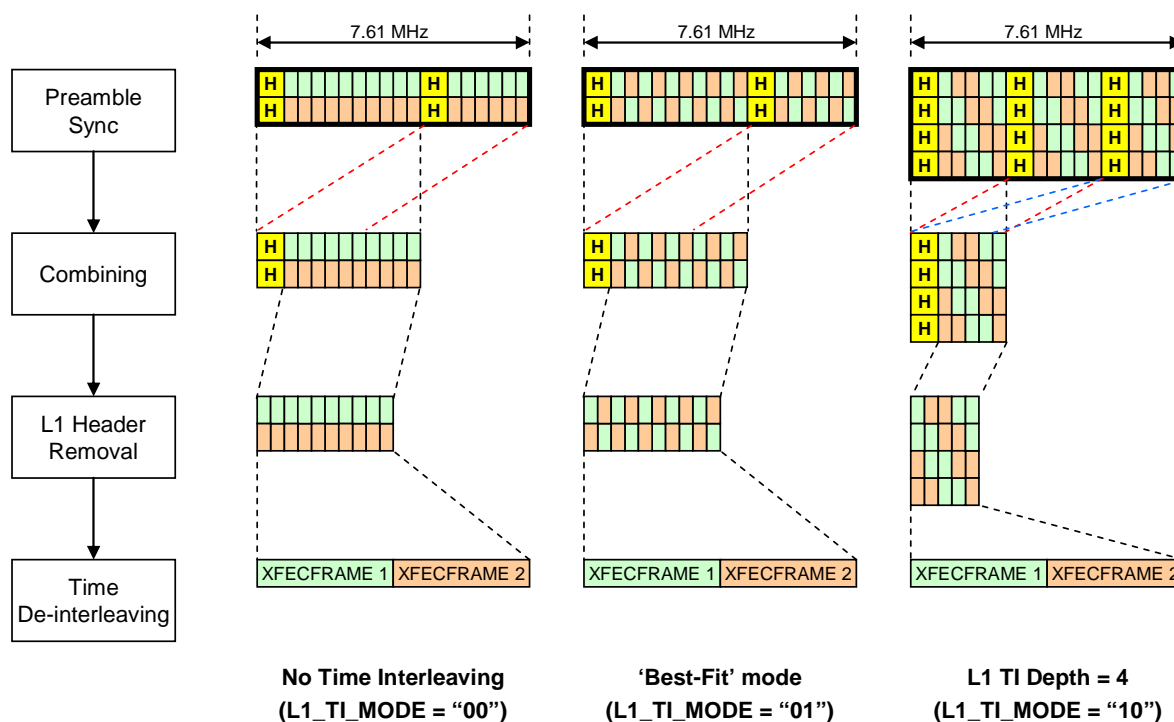


Figure 66: Time de-interleaving of L1 part 2 data

After the receiver detects the preamble and synchronizes to the L1 signalling block (occupying 3 408 carriers or 7,61 MHz bandwidth), it detects and decodes the L1 header. Regarding L1 header decoding process, there are two possible ways. One way is to decode only the first L1 header and get parameters, L1\_INFO\_SIZE and L1\_TI\_MODE. The other way is to combine all L1 headers inside L1 signalling block and decode the combined L1 header to get advantage of SNR gain. As the header signals include synchronization sequence, it is possible to detect and combine all L1 headers without decoding parameter L1\_SIZE\_INFO from the first L1 header signalling.

Next, the receiver may combine all L1 TI blocks inside the L1 signalling block to get better decoding performance like as in the case of L1 header decoding described above. The number of combined L1 TI blocks is determined by the number of copied L1 TI blocks to fill the entire L1 signalling block in the transmitter side. The combining of L1 TI blocks in different locations are indicated by dashed lines of different colours (red and blue) in combining stage of figure 66. As the L1 header is not included in preamble time interleaving and the interleaving sequence is same for every L1 TI block within the preamble, the same pattern is reserved for every L1 TI block so the TI blocks can be combined without any interference.

Finally, the preamble time de-interleaving process is applied according to the L1\_TI\_MODE after L1 header removal as shown in figure 66. Note that the interleaving depth may be different for the same amount of **L1-part2** data.

### 10.1.3.3 Memory-efficient implementation of time de-interleaver

Memory efficient way of de-interleaving for the preamble is very similar to that of de-interleaving for the **Data Slice** (see clause 10.2). The address generation and memory handling are exactly same. The only difference is that for the preamble de-interleaving, neither preamble pilot nor reserved dummy carrier for PAPR reduction is included even in address generation. Therefore, the selective reading process like equation (25) of clause 10.2 is not necessary.

### 10.1.3.4 Disabled time interleaving

The L1 time de-interleaving may be disabled in cases where the L1 time interleaving is not applied to the preamble in the transmitter side for the fast access to the **L1-part2** data or short latency application. In this case, it may still be useful to use the de-interleaver memory as a buffer, but the read and write sequences will both be identical.

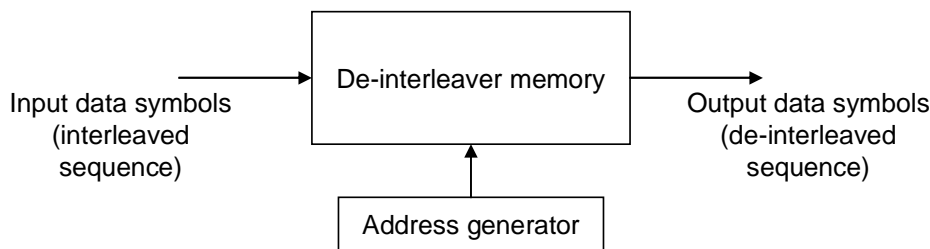
## 10.2 Time de-interleaving of payload data

### 10.2.1 Phase of time de-interleaving

As the TI block is synchronized to **Data Slice** boundary in the time direction, the receiver can begin time de-interleaving after **C2 frame** detection. The receiver first detects the preamble and decodes **L1-part2** data to get information about time interleaving depth. Then, the receiver can immediately start de-interleaving with a de-interleaver buffer.

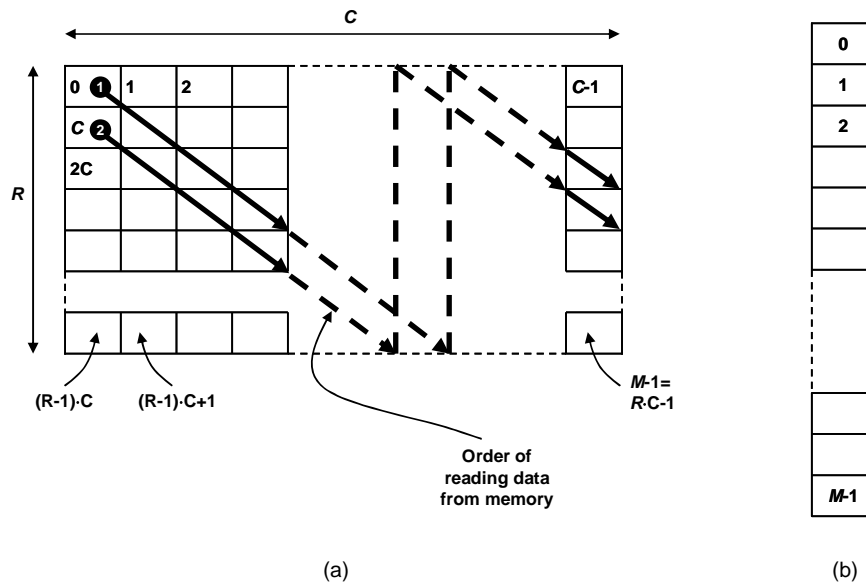
### 10.2.2 Memory-efficient implementation of time de-interleaver

The permutation function for the time interleaver is the same for each TI block within the **C2 frame** and so it might appear that two blocks of memory are required in the de-interleaver. However, a more efficient method exists using only one TI block's worth of memory. The memory-efficient de-interleaver is shown in the block diagram shown in figure 67.



**Figure 67: De-interleaver block diagram**

During each TI-block (except the first) **data cells** are read out one at a time from the de-interleaver memory according to the addressing sequence produced by the address generator. For each cell read out, a new cell from the input is written into the memory at the same address, as this memory location has just been cleared by reading the output cell.



**Figure 68: Implementation of de-interleaver memory**

NOTE: Figure 68 shows the implementation of the de-interleaver memory, showing (a) the conceptual implementation with rows and diagonals which defines the de-interleaving sequence and (b) the actual implementation as a sequential block of memory.

The time interleaver is defined to write data input into diagonal direction and read out row-wise from ( $R$  rows  $\times$   $C$  columns) buffer memory. The de-interleaver will therefore write data into  $R$  rows and read along the diagonal direction as shown in figure 68 (a). The line number 1 in diagonal direction is first read and the line number 2 is next read and so on. The total memory of the interleaver is therefore defined by:

$$M = R \times C$$

The actual block of memory to be used will be implemented as a sequential block of memory, as shown in figure 68 (b), and the main problem to be solved is to calculate the correct address sequence. The addresses of the elements of the memory will be calculated by index  $i$ , where:

$$0 \leq i \leq M - 1$$

The address generation can be better understood and explained by using 2-dimensional (2-D) memory shown in figure 68 (a). The addresses for sequential memory shown in figure 68 (b) can be easily calculated from the addresses of 2-D memory by simple conversion. A general equation for address generation can be obtained from following implementation perspective.

The address for the cell of 2-D memory can be defined by a coordination  $(c_{i,j}, r_{i,j})$  of which the element is a row and a column index for the  $i$ -th data of  $j$ -th TI block. To read the data in diagonal direction as shown in figure 68 (a), the "READ" address for  $i$ -th output data of a TI block can be calculated by adding an increment  $s_{i,j}$  to the row index  $r_{i,j}$  of the "WRITE" address for  $i$ -th input data of the same block. The amount of increment changes on a column by column basis.

For the first TI block ( $j=0$ ), after all input data are written row-wise into memory, the READ address of the first data output ( $i=0$ ) is same as the WRITE address of the first data input; there is no increment in  $r_{0,0}$  and the READ address is set to  $(0, 0)$ . For the next address generation hereafter, the increment  $s_{i,0}$  is added to the  $r_{i,0}$  of the WRITE address to get the  $r_{i,0}$  of the READ address. The increment  $s_{i,0}$  increases by 1 as the data index  $i$  increases until the  $r_{i,0}$  becomes  $(R-1)$ ; the last row is met. Whenever the last row is met, the increment for the  $r_{i,0}$  of the next READ address is reset to 0; the row index rolls up to the first row index. This process continues until the  $c_{i,0}$  becomes  $(C-1)$ ; the last column is met. The first line in diagonal direction will be found by addressing the  $(c_{i,0}, r_{i,0})$  coordinates in the order:

$$(0, 0), (1, 1), (2, 2), \dots, (R-1, R-1), (R, 0), (R+1, 1), (R+2, 2), \dots, (C-1, (C \bmod R)-1)$$

This completes one cycle of reading the first line in diagonal direction in figure 68(a). Then, the process over second line begins by setting the READ address to that of the first cell of second row and the same way of reading is repeated. The address sequence for the second line in diagonal direction will be:

$$(0, 1), (1, 2), (2, 3), \dots, (R-2, R-1), (R-1, 0), (R, 1), (R+1, 2), \dots, (C-1, C \bmod R)$$

When the last cycle of reading is finished, the whole process of reading over first TI block is completed.

The following equation generates  $(c_{i,0}, r_{i,0})$  of the READ address as described above:

$$\begin{aligned} c_{i,0} &= i \bmod C \\ s_{i,0} &= c_{i,0} \bmod R \\ r_{i,0} &= [s_{i,0} + (i \operatorname{div} C)] \bmod R \end{aligned} \quad (22)$$

The term  $(i \operatorname{div} C)$  can be interpreted as the WRITE address for data input into the  $(i \operatorname{div} C)$ -th row and the increment  $s_{i,0}$  is added to get the row index of READ address.

This describes the addresses to be generated to read out the first TI-block of de-interleaved data. However, the same address will also be used to write the new **data symbols**. Therefore, when this second input TI-block has all been stored, it too will be need to be read out in de-interleaved sequence.

Once again, the row index  $r_{i,1}$  of the READ address for the second TI block may be calculated by adding an increment  $s_{i,0}$  to  $r_{i,0}$  ( $r_{i,0}$  becomes the row index of the "WRITE" address for the second TI block). As another but simpler way, the increment itself can be accumulated as the TI block index  $j$  increases. Since the increment step is fixed to  $s_{i,0}$  according to the de-interleaving rule, accumulation is identical to multiplying  $s_{i,0}$  by block index  $j$ . By applying the accumulation of increment for  $j$ -th block, the general equation of (22) will be:

$$\begin{aligned} c_{i,j} &= i \bmod C \\ s_{i,j} &= (j \times c_{i,0}) \bmod R \\ r_{i,j} &= [s_{i,j} + (i \operatorname{div} C)] \bmod R \end{aligned} \quad (23)$$

Note that  $c_{i,0}$  as well as  $c_{i,j}$  are independent of block index  $j$  so  $c_{i,0}$  is replaced by  $c_{i,j}$  in generating the increment  $s_{i,j}$ .

Returning to the address generation for the sequential memory shown in figure 68(b), the coordinate address  $(c_{i,j}, r_{i,j})$  of 2-D memory can be converted into the linear address for the  $i$ -th data of  $j$ -th TI block by:

$$L_{i,j} = C \cdot r_{i,j} + c_{i,j} \quad (24)$$

The corresponding linear address for the 2-D memory is shown in figure 68(a).

Consider as an example the case when the de-interleaver has 8 rows and 12 columns. The address sequence generated for this TI block would be 0, 13, 26, 39, 52, 65, 78, 91, 8, 21, 34, 47, 12, 25, 38, 51, ..., 92, 9, 22, 35.

Note that the time de-interleaver for the **Data Slice** should output only **data cells** even if the addresses for the pilots and reserved dummy carriers are temporarily generated. For this operation, the time de-interleaver should know in advance the exact positions of those non-**data cells** and skip the memory reading process:

$$\begin{aligned} & \text{for } (i = 0; i < M; i = i + 1) \{ \\ & \quad \text{Generate Address } L_{i,j}; \\ & \quad \text{if } (c_{i,j}, r_{i,j}) = \text{Data Cell Address} \\ & \quad \quad \text{Read / Write Cell at } L_{i,j}; \\ & \} \end{aligned} \quad (25)$$

The time de-interleaver memory may be further reduced not to include non-**data cells** in practical implementation. However, the de-interleaver output should not change the de-interleaving sequence and the number of **data cells** allocated to each **OFDM symbol** within **Data Slice** when such kind of optimal size of memory is used.

A practical implementation to generate the above address sequence is straightforward. When the receiver starts to de-interleave the cells of a **Data Slice** after its synchronization, the address generator resets its increment  $s_{0,0}$  for the row index to 0. Once the address generation is initiated, the only value to be stored for the use in next de-interleaving frame is the TI block index  $j$  used for the previous de-interleaving frame. The increment itself can be generated by using the sample index  $i$  so needs not be stored.

### 10.2.3 Disabled time interleaving

The time de-interleaving may be disabled in cases where the interleaving is disabled in the receiver side for short latency application. In this case, it may still be useful to use the de-interleaver memory as a buffer, but the read and write sequences will both be identical.

## 10.3 Frequency de-interleaving of payload data

The frequency de-interleaver in DVB-C2 is applied to the equalized payload cells of a given **Data Slice** from one OFDM **data symbol** to the next. The number of payload cells per **Data Slice** per OFDM **data symbol** ( $N_{DS}$ ) can vary from **data symbol** to **data symbol**.  $N_{DS}$  comprises the number of cells ( $K_{DS,max} - K_{DS,min}$ ) minus the number of continual pilots, scattered pilots, reserved tones and cells that are located in **notches** within the **Data Slice**. Most of this information is either carried in or can be derived from the Layer 1 signalling. For this purpose, assume that a function  $DataCells(slice\ number, data\ symbol\ number, L1\ info)$  exists in the demodulator to provide  $N_{DS}$  for a given **Data Slice** number, data symbol number and from the Layer 1 information. The frequency de-interleaver shall be capable of de-interleaving received payload cells of the largest possible **Data Slice** with  $\max(N_{DS})$  cells given that  $(K_{DS,max} - K_{DS,min}) \leq 3\ 408$ . This means that the interleaver shall be capable of dealing with  $N_{max}$  payload **data cells** where:

$$N_{max} = (K_{DS,max} - K_{DS,min}) - N_{SP,D_X=24} - N_{CP}$$

and  $N_{SP,D_X=24}$  is the number of scattered pilots in 3 408 sub-carriers for the  $D_X = 24$  scattered pilot pattern and  $N_{CP}$  is the number of continual pilots in 3 408 sub-carriers.

The frequency de-interleaver memory is split into two banks: Bank A for even OFDM **data symbols** and Bank B for odd OFDM **data symbols**. Each memory bank comprises of  $N_{max}$  locations. Similar to the transmitter, a DVB-C2 receiver should use odd-only pseudo-random de-interleaving. In this the equalized payload cells from even OFDM **data symbols** (data symbol number of form  $2n$ ) of the **Data Slice** are written into the de-interleaver memory Bank A in a permuted order defined by the sequence  $H_0(q)$  and read out in a sequential order. Similarly, equalized payload cells from odd OFDM **data symbols** (data symbol number of form  $2n+1$ ) of the **Data Slice** are written into de-interleaver memory Bank B in a permuted order defined by the sequence  $H_1(q)$  and read out in a sequential order. In each case, the permuted order addresses  $H_{l,0,1}(q)$  are provided by the pseudo-random address generator from clause 9.4.5 of ETSI EN 302 769 [i.1]. In order to produce a continuous stream of cells at the de-interleaver output, when Bank A is being written (incoming even **data symbol**), Bank B is also being read (outgoing previous odd **data symbol**). Indeed, there is a sequential counter  $q$  used as the sequential read addresses and also as the lookup index to each of the permutation functions  $H_{l,0,1}(q)$  which provide the write addresses. If all **data symbols** in the **Data Slice** contained  $N_{max} = C_{data}$  payload cells, then the number of write addresses for data symbol number  $2n+1$  shall match the number of read addresses for **data symbol** number  $2n$  otherwise some **data cells** of **data symbol**  $2n$  will be skipped. Unfortunately  $N_{DS}$  can be different from **data symbol** to **data symbol**. Suppose  $N_{DS}(2n)$  is less than  $N_{DS}(2n+1)$  then the pseudo-random address generator  $H_1(q)$  would have to produce more addresses than there are cells to be read from memory Bank A because the sequential LookUp-Table (LUT) indices for generating the permuted write addresses for writing to Bank B would range from 0 to  $N_{DS}(2n+1)-1$ . The case in which  $N_{DS}(2n) > N_{DS}(2n+1)$  can also occur. In this case the sequential read address counter for Bank B would need to exceed  $N_{DS}(2n+1)-1$  as more  $H_0(q)$  write addresses are needed for Bank A. Recalling that the function  $DataCells(slice\ number, data\ symbol\ number, L1\ info)$  returns the number of payload cells in the current slice for the given **data symbol** and noting that HoldBuffer is a small amount of storage with write address  $wptr$  and read address  $rptr$ , the de-interleaving proceeds as follows at the reception of an even **data symbol** number  $2n$ :

- 1)  $q = 0$ ;
- 2)  $C_{max} = \max(DataCells(slice\ number, 2n-1, L1\ info), DataCells(slice\ number, 2n, L1\ info))$ ;

- 3) Generate address  $H_0(q)$ ;
- 4)  $rdEnable = (q < DataCells(slice\ number, 2n-1, LI\ info))$ ;
- 5)  $wrEnable = (H_0(q) < DataCells(slice\ number, 2n, LI\ info))$ ;
- 6) if (rdEnable) Read cell  $q$  of output de-interleaved **data symbol**  $2n - 1$  from location  $q$  of memory Bank B;
- 7) Store cell  $q$  of incoming interleaved **data symbol**  $2n$  into location  $wptr$  of HoldBuffer and increment  $wptr$ ;
- 8) if (wrEnable):
  - a) Write cell  $rptr$  of HoldBuffer into location  $H_0(q)$  of memory Bank A and increment  $rptr$ .
  - b) If( $wptr == rptr$ ) reset both  $rptr = wptr = 0$ .
- 9) Increment  $q$ ;
- 10) if ( $q < C_{max}$ ) go to 3.

Then with **data symbol**  $2n+1$  at the input of the de-interleaver:

- 1)  $q = 0$ ;
- 2)  $C_{max} = \max(DataCells(slice\ number, 2n, LI\ info), DataCells(slice\ number, 2n+1, LI\ info))$ ;
- 3) Generate address  $H_1(q)$ ;
- 4)  $rdEnable = (q < DataCells(slice\ number, 2n, LI\ info))$ ;
- 5)  $wrEnable = (H_1(q) < DataCells(slice\ number, 2n+1, LI\ info))$ ;
- 6) if (rdEnable) Read cell  $q$  of output de-interleaved **data symbol**  $2n$  from location  $q$  of memory Bank A;
- 7) Store cell  $q$  of incoming interleaved **data symbol**  $2n+1$  into location  $wptr$  of HoldBuffer and increment  $wptr$ ;
- 8) if (wrEnable):
  - a) Write cell  $rptr$  of HoldBuffer into location  $H_1(q)$  of memory Bank B and increment  $rptr$ .
  - b) If( $wptr == rptr$ ) reset both  $rptr = wptr = 0$ .
- 9) Increment  $q$ ;
- 10) if ( $q < C_{max}$ ) go to 3.

The required width for each memory location depends on the resolution with which each cell is represented after channel equalization. Each de-interleaver memory cell would hold at least: the complex cell information and the channel state information for the cell.

## 10.4 Use of Pilots

Pilots can be used for typically the following four purposes:

- C2-frame synchronization.
- Integer frequency offset estimation.
- Channel estimation.
- Common phase error estimation.

The C2-frame synchronization is introduced in clause 10.1, the frequency offset and channel estimation details are discussed in clause 10.5 and the common phase error estimation method is described in clause 10.5.1.

## 10.5 Phase noise requirements

### 10.5.0 Introduction

Phase noise is specified as single sideband phase noise power in a 1 Hz bandwidth at a frequency  $f$  from the carrier frequency. The unit of  $L(f)$  is dBc/Hz, representing the noise power relative to the carrier power contained in a 1 Hz bandwidth centred at a certain offsets from the carrier (e.g. at 10 kHz offset).

DVB-C2 uses OFDM and DVB-C single carrier QAM. Therefore the effect of phase noise on the signal will be different for DVB-C2 compared to DVB-C. Generally the phase-noise considerations are the same as for DVB-T/T2.

Phase noise added to an OFDM signal causes two distinct effects: common phase error and ICI.

Low-frequency noise gives rise to common phase error. Common phase error is a common rotation of all the constellations transmitted in one OFDM symbol. Because it is common for all carriers, it is possible to measure it and cancel it out. The DVB-C2 standard [i.1] includes continual pilots which, amongst other uses, can be used for the purpose of cancelling common phase error, according to the method described in clause 10.5.1. Taking this and other possible cancellation measures into account, the impact of common phase error is expected to be negligible.

High-frequency noise causes inter carrier interference (ICI). ICI is a form of crosstalk between carriers and manifests itself as an additional noise term, degrading the constellation SNR. The ICI part of the phase-noise may be approximated by integrating  $L(f)$  from half the carrier spacing to half of the signal bandwidth on both sides of the carrier. An accurate calculation requires the use of weighting functions, e.g. as described in [i.13].

In practice, tuners will vary in their phase-noise spectra, and manufacturers should therefore calculate the ICI value caused by the phase noise values of their tuners. Because ICI behaves like AWGN manufacturers then can estimate the implementation loss of their tuners.

#### 10.5.1 Common phase error correction

Low-frequency phase noise causes common phase error. Since the random rotation of common phase error are (by definition) the same for every carrier within the same **OFDM symbol** they can be measured (and thus corrected if desired) by using the Continual Pilots (CPs).

CPs are specified carriers which transmit reference information in every symbol. The reference information is a function of the carrier index,  $k$ .

The random change in common phase error from one symbol to the next can be measured as follows. The received CPs are differentially demodulated on each CP carrier by multiplying the current symbol by the complex conjugate of the same carrier in the previous symbol. The values are summed for all the CPs in one symbol. The argument of the resulting complex number is the common phase error of the current symbol with respect to the previous one. This can be used to correct the current symbol. In effect the very first symbol received is treated as a reference of zero common phase error.

The summation of the results for a large number of CPs reduces the effect of normal additive noise by averaging. This is important since this additive noise will itself contribute a phase-noise component to the measurement - it is important for this to be substantially smaller than the common phase error which is being corrected, otherwise the correction process will have a negative impact. It may also be necessary to exclude from the calculation any CPs on carriers that have been found to be suffering from CW or any interference.

The common phase error-corrected symbols are then used in all subsequent processing.

All CP locations within **data symbols** coincide with Continual Pilot locations within **preamble symbols**. Therefore it is possible to apply the same common phase error correction for preamble and **data symbols** if the common phase error calculation is based solely on the CP locations (of **data symbols**).

## 10.5.2 Channel Equalization

### 10.5.2.1 Overview

#### 10.5.2.1.1 The need for channel estimation

The received carrier amplitudes output by the receiver FFT are not in general the same as transmitted - they are affected by the channel through which the signal has passed on its way from the transmitter.

Consider the *channel extent*, which could be variously described as: the duration of the impulse response from the first significant component to the last; or the shortest duration which can be chosen without excluding any significant impulse-response components; or, more practically, the shortest duration which can be chosen so that at least  $X$  % of the total signal energy is included, where  $X$  % is some substantial proportion, e.g. 99,9 %.

Provided this channel extent of the channel's impulse response does not exceed the guard interval  $T_G$ , and that correct OFDM synchronization is maintained, the received (complex) carrier amplitudes can be given by:

$$Y_{k,l} = H_{k,l} X_{k,l} + N_{k,l}$$

where:

$X_{k,l}$  represents the complex modulation-'symbol' (constellation) applied on carrier  $k$ , symbol  $l$ ;

$Y_{k,l}$  represents the corresponding received carrier-amplitude;

$H_{k,l}$  represents the (complex) frequency response of the channel during symbol  $l$ , sampled at the carrier frequency, i.e.  $H_{k,l} = H_l(f_k)$ ; and

$N_{k,l}$  represents the additive receiver noise.

**NOTE:** A further explanation of what happens is as follows. The received signal is the transmitted signal convolved with the channel impulse-response. The addition of a COFDM guard-interval (also known as a cyclic prefix) has the effect of converting this linear convolution into a cyclic one (provided the channel extent does not exceed the guard interval). Cyclic convolution in time corresponds to multiplication in frequency, when the two are related by a DFT operation.

The relationship can also be written as a matrix equation; the effect of the channel is to multiply by a simple diagonal matrix - unless orthogonality is lost (e.g. because the channel extent is too great), whereupon more entries in the matrix become non-zero.

The receiver needs knowledge of the  $H_{k,l}$  if it is to interpret the  $Y_{k,l}$  in the best way. One simple way for re-scaling the received constellations is equalization: dividing by our *best estimate*  $H'_{k,l}$  of the complex channel response  $H_{k,l}$  pertaining to each **data cell**. It will be clear that provided the estimate  $H'_{k,l}$  is noise-free, this operation does not change the signal-to-noise ratio. The SNR will however already be degraded by the channel response as it is known that: weakly-received carriers have a poorer SNR than others.

Dividing by the estimated channel response is somewhat similar to a zero-forcing equalizer, and would normally be deprecated on the grounds of aggravating the effects of noise. However, under the assumption that coded OFDM is being used, it is possible to simply weight the soft-decision values fed to the error corrector appropriately to take account of the different SNRs with which the data on the various carriers are received. (These weighted soft decisions are also known as metrics).

An alternative method simply inputs both  $Y_{k,l}$  and  $H'_{k,l}$  into the metric calculation without re-scaling the  $Y_{k,l}$  to the standard size first - the result is equivalent but perhaps less intuitive.



### 10.5.2.1.2 Obtaining the estimates

The channel estimate can be derived from the known information inserted in certain **OFDM cells** - a term used for the entity conveyed by a particular combination of carrier (location in frequency) and symbol (location in time). These cells containing known information are known as *pilot cells*; they are affected by the channel in exactly the same way as the data and thus - barring the effect of noise - precisely measure the  $H_{k,l}$  for the cell they occupy. The measured channel estimate for this cell is calculated as:

$$H'_{k,l} = \frac{Y_{k,l}}{X_{k,l}} = H_{k,l} + \frac{N_{k,l}}{X_{k,l}}$$

In principle this calculation can be done for any cell where the transmitted information  $X_{k,l}$  is known. In most cases this will be the scattered-pilot cells (SPs). However, continual pilots (CPs) are also available for a smaller proportion of cells.

To obtain the estimates of the channel response for every **data cell**, the normal approach is to *interpolate* between the values  $H'_{k,l}$  (which are only available for those  $\{k, l\}$  corresponding to transmitted pilots) to provide values for every cell.

### 10.5.2.2 Fundamental limits

The channel response  $H_{k,l}$  in general varies with both *time* (symbol index  $l$ ) and *frequency* (carrier index  $k$ ). The temporal variation corresponds to external causes such as Doppler shift and spread, but also to instantaneous error in the receiver's frequency tracking. The variation with frequency is a symptom of channel selectivity, itself caused by the channel comprising paths having different delays.

In effect, the receiver samples the channel response  $H_{k,l}$  by measuring it for the cells  $\{k, l\}$  within which pilot information has been transmitted. For the most part, the SPs are used for this purpose, but where other types of pilot information are present, they can also be used if desired. The SPs constitute a form of 2-D sampling grid, and in consequence there are limits, according to the Nyquist criterion, on the rates of variation of the channel response with time and frequency that can be measured using SPs.

Clause 9.6.2 of ETSI EN 302 769 [i.1] defines the SP patterns of DVB-C2, and introduces terms  $D_X$  and  $D_Y$  to characterize them.  $D_X$  is the separation between pilot-bearing carriers, so if  $D_X = 3$ , say, then every third carrier contains SPs - but not in general within a single symbol. This is because there is a diagonal pattern, which repeats every  $D_Y$  symbols. So, on carriers that are pilot-bearing, an SP occurs, and a measurement can be made, in every  $D_Y$ th symbol. Symbols occur at the rate  $f_S = 1/T_S = 1/(T_U + T_G)$ . It follows that the Nyquist limit for temporal channel variation that can be measured is  $\frac{\pm 1}{2D_Y(T_U + T_G)}$  Hz.

Given suitable temporal interpolation, either a measurement or an interpolated estimate of the channel response for every cell on the pilot-bearing carriers will be available. Estimates for the remaining cells can then be found by frequency interpolation between the pilot-bearing carriers. Since these are spaced by  $D_X$  carriers, or  $D_X f_U = D_X / T_U$  Hz, it follows that the maximum Nyquist channel extent, or spread between the first and last paths in a channel that can be supported, is  $T_U / D_X$  sec.

Note that this approach is a variables-separable one leading to a rectangular Nyquist area on a diagram of Doppler versus delay. This rectangular area corresponds to a (dimensionless) timewidth-bandwidth product having the value:

$$\frac{1}{D_Y(T_U + T_G)} \cdot \frac{T_U}{D_X} = \frac{1}{D_X D_Y (1 + GIF)}$$

where  $GIF = T_G / T_U$  is the **guard-interval fraction**.

It can be shown that the same sampling grid of channel measurements can be interpreted to produce other shapes of supportable 'area', in general non-rectangular, but whose total area remains the same.

Note that the spacing between pilots in just one particular symbol is in general greater than  $D_X$  carrier spacings, being  $D_X D_Y$ , the inverse of the scattered-pilot density. It follows that if no temporal interpolation is performed, and channel estimates are obtained solely by frequency interpolation within one symbol, then the applicable Nyquist limit for channel extent is tighter than described above, being  $\frac{T_U}{D_X D_Y}$ . DVB-C2 pilot patterns have nevertheless been chosen so that frequency-only interpolation is both possible and sensible in certain scenarios.

### 10.5.2.3 Interpolation

#### 10.5.2.3.1 Limitations

In principle, channel variations within the Nyquist supported area described in the previous clause can be measured. However, the Nyquist limit can only be very closely approached when using an interpolator having a very large number of taps. This is unattractive on grounds of cost and complexity, but is also bounded by other practical constraints.

The frequency interpolator can only make use of the finite number of pilot-bearing carriers.

Similarly, the temporal interpolator clearly cannot access measurements from before the time the receiver was switched on or the current radio-frequency channel was selected. Much more importantly, the length of the temporal interpolator is tightly limited by the fact that the main signal stream awaiting equalization has to be delayed while the measurements to be input into the temporal interpolator are gathered. This delay has a large cost in terms of memory needed, but also in terms of the delay it introduces before the programme material can be delivered to the viewer. The temporal interpolator is thus usually much more constrained in its size than the frequency interpolator.

It follows that the full Nyquist area cannot in practice be supported.

An *ideal* interpolator, whose bandwidth (or as appropriate, 'time-width' as defined later) is chosen to be some fraction  $x < 1$  of the Nyquist limit can also reduce the noise on the channel estimate by the same factor  $x$ . Note however that practical interpolators are unlikely to realize this good a result. Perversely, simple linear interpolation does achieve noticeable noise reduction, but in this case at the expense of poor interpolation accuracy through most of the wanted bandwidth.

#### 10.5.2.3.2 Temporal interpolation

The presumption is that most receivers, on grounds of simplicity and minimizing delay, will use at most simple linear temporal interpolation. To begin the temporal interpolation the simplest case of interpolation between scattered pilots will be considered.

In DVB-C2 both of two scattered pilot patterns have  $D_y = 4$ . It follows that simply to perform simple linear interpolation requires three ( $D_y - 1$ ) symbols' worth of storage for the main data stream.

Although it is unlikely to have any Doppler distortions in cable channel environment, the channel estimation may vary over time due to a residual synchronization error. It is depending upon the residual synchronization error, but once good synchronization is achieved, the linear interpolation should be quite accurate. It is also possible to reduce the interpolation bandwidth to have better noise reduction, if better synchronization can be achieved.

Temporal interpolation accuracy can be increased somewhat without increasing main data stream memory by using "one-sided" interpolator designs having more taps than a simple linear interpolator.

NOTE: Due to the chosen pilot density in DVB-C2, temporal interpolation might not be necessary.

### 10.5.2.3.3 Frequency interpolation

Fortunately the frequency interpolator can use rather more taps. This is in fact necessary, since the 'time-width' (an analogous term, for frequency-domain sampling, to the common use of bandwidth in relation to time-domain sampling) of the interpolator now has to be an appreciable fraction of the Nyquist limit. Depending on an interpolation method, the necessary time-width (if the full guard-interval range of delay is to be supported) can be 19 % or 75 % Nyquist; the 19 % applies for the temporal and frequency interpolation method case whilst the 75 % is for the frequency only interpolation case. Considering DVB-C2 has 4 096-QAM and CR9/10, it is required accurate interpolation. I.e. generally requires high order interpolation filter. It is, however, possible to dramatically reduce the interpolation filter order by having large transition region, which is possible in the temporal and frequency interpolation case due to the small necessary time-width compare to the Nyquist limit. Furthermore, there is a case for making the time-width (if possible) a little greater than simply the guard-interval duration, in order to cope better with a degree of timing error, or with channels whose extent is not strictly contained within the guard-interval duration.

Suppose  $N$  taps are used. For most carriers, the interpolator would make use of estimates/measurements from  $N/2$  pilot-bearing carriers on one side plus  $N/2$  pilot-bearing carriers on the other side. However, as the carrier to be estimated approaches the upper or lower limit of the OFDM spectrum, this is no longer possible, as some of the interpolator taps 'fall off the edge'. In this case it is still possible to use an  $N$ -tap interpolator, but its taps shall be chosen so that it becomes progressively more one-sided. The performance is slightly compromised, but remains better than the alternative of using fewer and fewer interpolator taps as the edge is approached.

## 10.6 Tuning to a Data Slice

As soon as the receiver performed all initial synchronization tasks (i.e. being synchronized in time, frequency and C2 framing) it can be tuned to the **Data Slice** where the required **PLP** is embedded. This clause illustrates the needed steps and provides two simple examples.

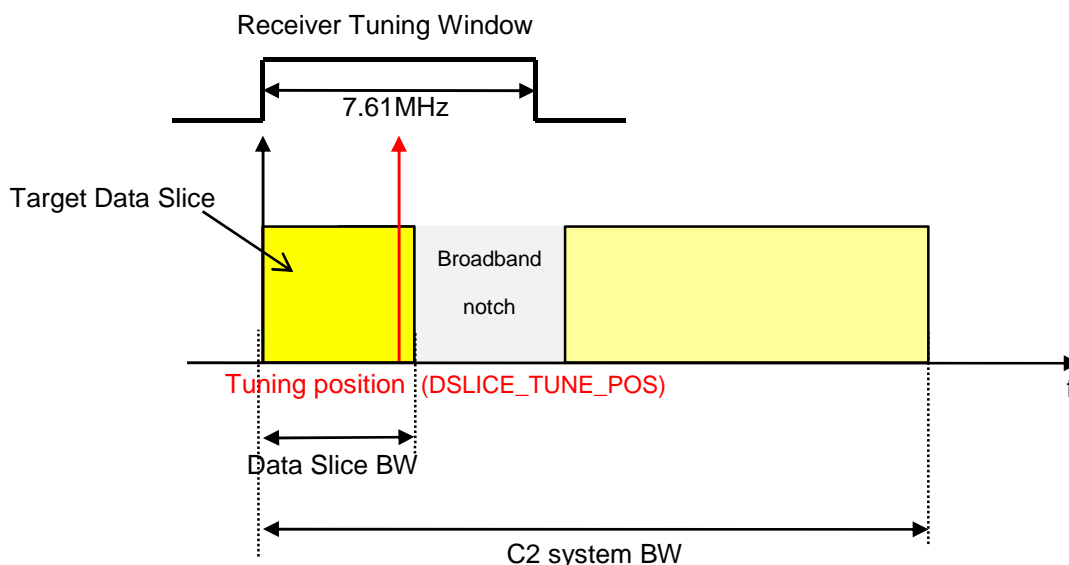
During initial acquisition the C2 receiver tunes to an arbitrary frequency within the C2 signal (i.e. the tuning window should not include frequencies outside the defined C2 bandwidth). As soon as the receiver is able to recognize the preamble it can start to compensate offsets and to extract the **L1 part2** signalling within the **preamble symbols**. The **L1 part2** signalling list contains all physical layer specific information of the C2 signal and especially the partitioning of the C2 signal into **Data Slices** and **notches**. Note that typically a tuning position is not aligned to the **L1-part2** information block repetition rate (3 408 subcarriers). Therefore the **L1-part2** signalling has to be retrieved after the Rx FFT by simple QAM symbol reordering (see clause 10.1.1.5).

The decoding of the L1 signalling itself is described in clause 10.1.1.5.

The tuning to a **Data Slice** is defined in an unambiguous way by several **L1 part2** parameters. First of all the parameter **DSLICE\_TUNE\_POS** defines the tuning position of the receiver. This field indicates the tuning position of the associated **Data Slice** relative to the **START\_FREQUENCY** and has a bit width of 13 bits or 14 bits according to the **GUARD\_INTERVAL** value. When **GUARD\_INTERVAL** is '00', the bit width of this field is 13 bits and indicates the tuning position in multiples of 24 carriers within the current **C2 frame**. Otherwise the bit width of this field is 14 bits and indicates the tuning position in multiples of 12 carriers within the current **C2 frame** relative to the **START\_FREQUENCY**.

According to the C2 standard [i.1], **DSLICE\_TUNE\_POS** shall be a value at least 1 704 carriers from the edge of a broadband notch or the start or end of the **C2 system**. The intention of this requirement is to ensure that the cells of a complete L1 block are available within the tuner receiving window.

The only possible exception for this requirement is for **Dependent Static Data Slices** (see clause 6.10.1.2), as these standalone **Data Slices** might have a bandwidth below 3 408 subcarriers, i.e. a successful L1 block decoding is not possible. In that case it might happen that `DSLICE_TUNE_POS` has a value below 1 704 carriers from the edge of a broadband notch. Figure 69 shows a related example, where the difference between tuning position and the start of the broadband notch is below 1 704 subcarriers:



**Figure 69: DSLICE\_TUNE\_POS example with broadband notch**

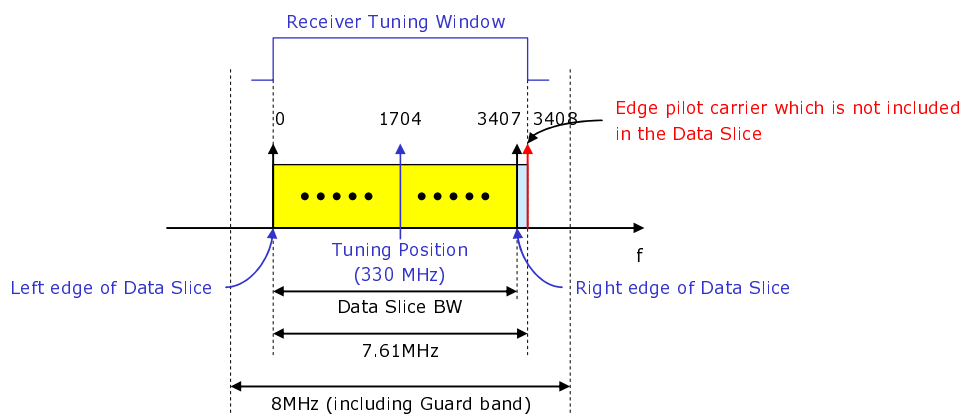
Generally, the tuning position should be configured by the transmitter so that the undesired interference (outside the **C2 system BW**) is outside the receiver tuning window. Any interference from the broadband notch should be considered when choosing parameters of the **Dependent Static Data Slice**, too.

Beside the `DSLICE_TUNE_POS` the start and stop carriers of a **Data Slice** are given in the **L1-part2** signalling list. These two parameters (`DSLICE_OFFSET_LEFT` and `DSLICE_OFFSET_RIGHT`) are given as offset values in relation to the `DSLICE_TUNE_POS` value. As for other **Data Slice** parameters the granularity (24 subcarriers or 12 subcarriers) and the bit field width (8 bit or 9 bit) depend on the chosen `GUARD_INTERVAL` value.

A typical receiver performs the receive FFT across its reception window (typically 4k FFT within an 8 MHz reception window) and selects these subcarriers that are assigned to the selected **Data Slice**. In the following two examples are given:

EXAMPLE 1: This simple scenario is taken from clause 8.4.4:

<code>GUARD_INTERVAL:</code>	00	Guard Interval is 1/128
<code>NUM_BUNDLED_CH:</code>	00001	The C2 signal width is 7,61 MHz
<code>START_FREQUENCY:</code>	000000000001100000010000	Start Frequency is 330 MHz = subcarrier $330E6 \times 448 \text{ usec} = 147\,840 \text{ usec}$ 24 carriers granularity -> $6\,160 = 0x1\,810 = 1100000010000$
<code>DSLICE_TUNE_POS:</code>	0000001000111	The tuning position of this <b>Data Slice</b> is 1 704 <sup>th</sup> carrier frequency of this <b>C2 system</b> 24 carriers granularity -> $71 = 0x47 = 0000001000111$
<code>DSLICE_OFFSET_LEFT:</code>	10111001	The left edge of this <b>Data Slice</b> is start frequency (apart from tuning position as much as 1 704 carrier spacing)



**Figure 70: Relation between Data Slice and tuning window positioning**

EXAMPLE 2: In this scenario the tuning position is outside a rather narrow **Data Slice**:

GUARD\_INTERVAL: 00 Guard Interval is 1/128

NUM\_BUNDLED\_CH: 00100 The C2 signal width is 31,61 MHz (4 x 8 MHz - 0,39 MHz)

START\_FREQUENCY: 000000000001100000010000  
Start Frequency is 330 MHz = subcarrier 330E6 x 448 usec = 147 840 usec  
24 carriers granularity -> 6 160 = 0 x1 810 = 1100000010000

DSLICE\_TUNE\_POS: 0000001000111 The tuning position of this **Data Slice** is 1 704<sup>th</sup> carrier frequency of this **C2 system**  
24 carriers granularity -> 71 = 0x47 = 0000001000111

DSLICE\_OFFSET\_LEFT: 00000010 The Data slice starts 48 carriers right from the tuning position

DSLICE\_OFFSET\_RIGHT: 01000111 The Data slice ends 1 703 carriers right from the tuning position

## 10.7 Buffer Management

NOTE: Clause 7.5 of the present document includes refinements of annexes C and G of the DVB-C2 specification [i.1], which will appear as a non-editorial update in the specification in the future. Thus, the explanations stated in this clause will refer to clause 7.5 instead of the current versions of the DVB-C2 specification [i.1].

In the course of the DVB-C2 Validation and Verification process the testing of the stream synchronization aspects introduced the necessity to define a more precise handling of the ISSY data calculation and insertion within the **DVB-C2 system** processing. The specification defines the details about Input Stream Synchronization in clause annex C, including the insertion order of ISSY data fields of the baseband header. Although annex G of the specification describes the **Transport Stream** regeneration, the exact meaning and use of the ISSY data fields remain unspecified.

The following explanations shall help to understand the meaning and use of the ISSY fields as referred to in annex C:

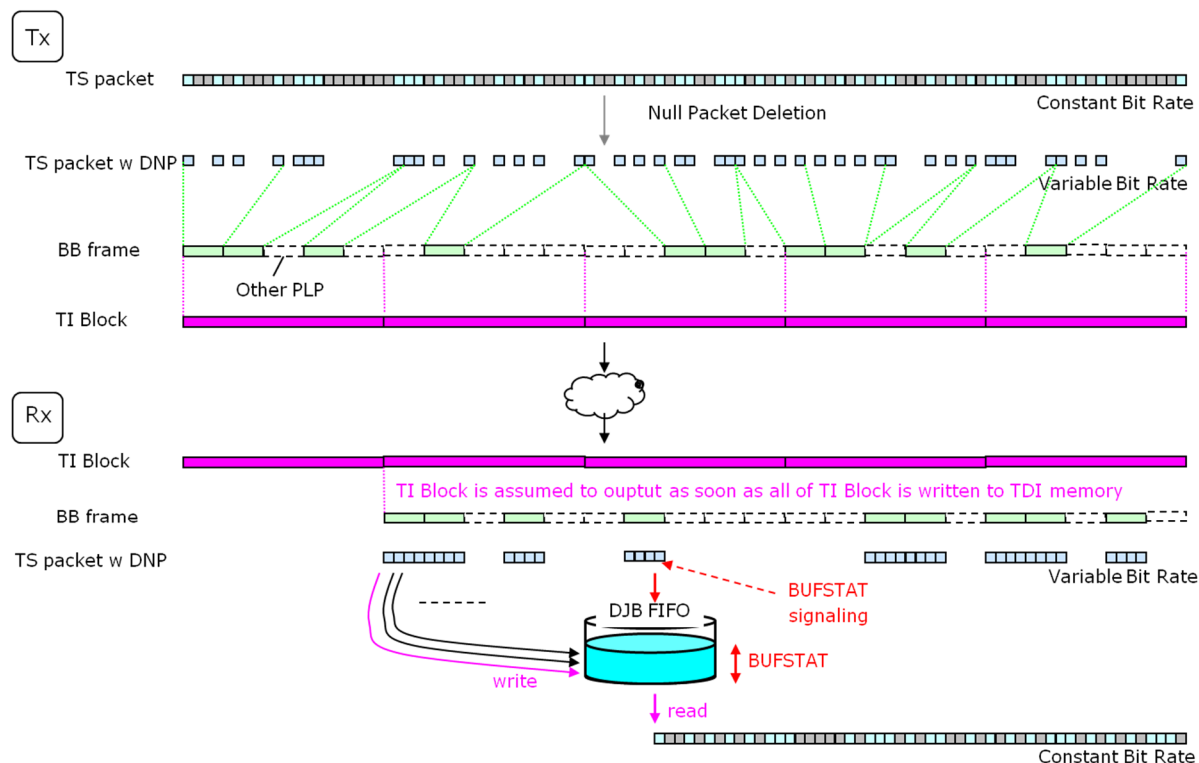
- BUFS

The BUFS field is the quantity of memory the receiver should allocate for the DeJitter Buffer (DJB) in order to prevent an overflow of the buffer and thus to ensure an uninterrupted processing of a single or group of transported streams based on the ISSY calculations of the transmitter.

- BUFSTAT

The BUFSTAT field signals the level of the reception FIFO buffer at the time the ISSY field is received (in the BaseBand frame (High Efficiency Mode) or **Transport Stream** packet compound (Normal Mode), respectively) The value of the BUFSTAT field semantically refers to the quantity of bits of the transported stream before the reinsertion of Null Packets (in case of the stream being a **Transport Stream**).

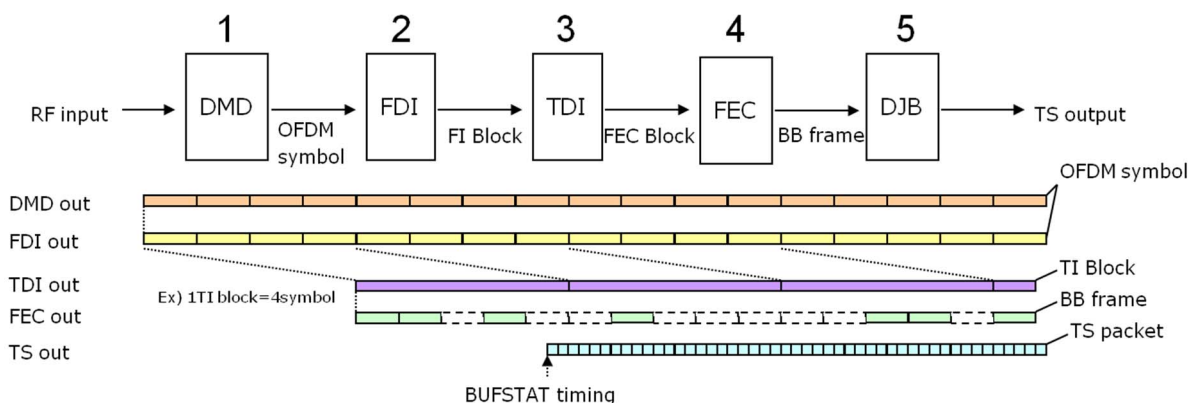
Figure 71 shows the semantic data flow within a **C2 system** chain with respect to ISSY data calculation and insertion.



**Figure 71: Semantic data flow with respect to ISSY calculations at the transmitter and receiver side**

At the transmitter side the input stream to the **C2 system**, a **Transport Stream** has a constant bit rate. After the application of the (optional) Null Packet Deletion to the **Transport Stream**, the resulting variable bit rate data (including additional signalling fields such as the ISSY data or the DNP counter) is merged into BaseBand frames and then processed by the Time Interleaver (amongst other modules).

At the receiver side the BaseBand frames are reconstructed as soon as all data to the Time Interleaver is written to TI memory. The content of the BaseBand frame, i.e. the variable bit rate **Transport Stream** packets (with deleted Null Packets), is queued in the DeJitter Buffer (DJB), i.e. the reception FIFO buffer. BUFSTAT should be used for controlling the buffer level at the receiver. BUFSTAT indicates the level of the buffer at the instant it is received.

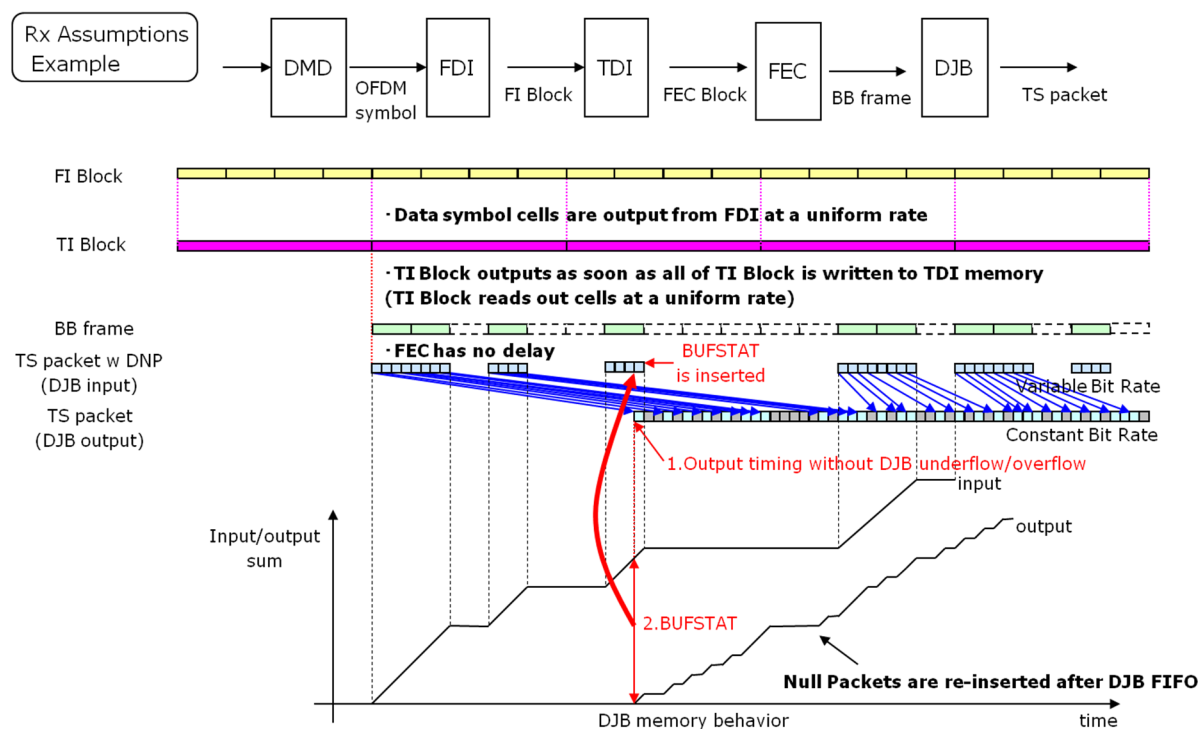


**Figure 72: Illustrative example explaining receiver assumptions regarding BUFSTAT calculation**

In the calculation of the value of BUFSTAT, the following assumptions should be made about the receiver:

- 1) The demodulation stages have no delay, and the cells carried in a particular symbol are written to the frequency de-interleaver at a rate of  $R_S$  cells per second starting from the moment symbol starts being received.  $R_S = 1/T$ , where  $T$  is as defined in clause 10.1.
- 2) The cells carried in a particular symbol are output from the frequency de-interleaver at a uniform rate during the time  $T_S$  that the symbol is being received.
- 3) The Time DeInterleaver (TDI) will read out the de-interleaved cells of that TI-block, starting as soon as all the cells of a TI-block have been written to the TDI memory. The TDI will read out the cells at a uniform rate and feed them to the FEC chain in the same time that the symbols of TI block were received.
- 4) The FEC chain has no delay and can process cells of selected **PLPs** continuously.
- 5) The data field bits of decoded **BBFrames** belonging to a **PLP** are then converted to canonical form. The canonical form is equivalent to Normal Mode with 3-byte ISSY and NPD enabled (see clause 5.1). The canonical form data is then written to the DJB. Bits are read out from the buffer at a constant rate calculated from the received ISCR values. Removed SYNC bytes and deleted Null Packets are re-inserted at the output of the DJB.

An illustrative example is shown below in figure 73.



**Figure 73: Example of system processing at the receiver: BUFSTAT timing and FIFO buffer level**

Following the receiver assumptions outlined above, the **Transport Stream** packets (with Null Packets deleted as necessary) are input to the DeJitter Buffer (DJB). In order to ensure the DJB does not underflow or overflow, a number of packets are stored in the buffer, and the **Transport Stream** is begun to output at a constant rate with the re-insertion of the deleted Null Packets. BUFSTAT is the value of the difference between the number of bits input and the number of bits output from the DJB. The exact timing of the insertion of BUFSTAT and BUFS should follow the directions given in clause 7.5 of the present document.

## 10.8 DVB-C2 FECFrame Header Detection

### 10.8.1 Overview of FECFrame Header Detection

The receiver needs to detect the **FECFrame** Header which is inserted in front of one or two **FECFrames** to support Adaptive Coding and Modulation (ACM). The **FECFrame** Header carries information of the **PLP\_ID**, the Coding and Modulation parameters of the following **FECFrame**, and the number of **FECFrames** following the information carried by this header.

### 10.8.2 FECFrame Header Detection

The application of the robust or the high efficiency **FECFrame** header is signalled within the Layer 1 - Part 2 signalling. The robust **FECFrame** header is taken here as an example. The **FECFrame** header detection can be performed by the following steps:

- 1) Assume that the 32-symbol complex sequence  $(s_0, s_1, \dots, s_{31}) = (r_i, r_{i+1}, \dots, r_{i+31})$  is the robust **FECFrame** header and demodulate them into a 64-bit sequence  $(a_0, a_1, \dots, a_{63})$  by a QPSK demapper. The complex symbol,  $r_i$  is a received **data symbol**. Note that the notation of  $(s_0, s_1, \dots, s_{31}) = (r_i, r_{i+1}, \dots, r_{i+31})$  is to describe the search process of **FECFrame** header over received **data symbols**. Thus,  $i = i+1$  means that the **data symbol**  $r_i$  is not the beginning symbol of the **FECFrame** header and the symbol index is advanced to see if  $r_{i+1}$  is the beginning symbol of the **FECFrame** header.
- 2) Compute the estimated 32-bit PN sequence  $\tilde{\mathbf{w}}^{RM} = (\tilde{w}_0^{RM}, \tilde{w}_1^{RM}, \dots, \tilde{w}_{31}^{RM})$   $\tilde{\mathbf{w}}^{RM} = (\tilde{w}_0^{RM}, \tilde{w}_1^{RM}, \dots, \tilde{w}_{31}^{RM})$  by  $\tilde{w}_{(k+2)_{32}}^{RM} = a_{2k} \oplus a_{(2k+5)_{64}}$   $w_{(k+2)_{31}}^{RM} = a_{2k} \oplus a_{(2k+5)_{64}}$ ,  $k=0,1,\dots,31$  where  $(x)_y$  is the result of  $x$  modulo  $y$ .
- 3) Compute the binary correlation of  $\tilde{\mathbf{w}}^{RM}$   $\tilde{\mathbf{w}}^{RM}$  and  $\mathbf{w}^{RM}$  by  $C_p = \sum_{k=0}^{31} (2 \cdot \tilde{w}_k^{RM} - 1)(2 \cdot w_k^{RM} - 1) C_p = \sum_{k=0}^{31} (2\tilde{w}_k^{RM} - 1)(2w_k^{RM} - 1)$ .
  - 3.1) If  $C_p < T_1$ , go to step 1 and advance symbol index by 1, e.g.  $i = i+1$ .
  - 3.2) If  $C_p \geq T_1$ , perform step 4.

The maximal value of the correlation is 32. The setting of  $T_1 = 20$  is good enough that most of the non-**FECFrame** header vectors will be identified in this step without missing the wanted **FECFrame** header vector.

- 4) Each bit of the 32-bit RM codeword is decoded for example by combining log-likelihood ratios of upper branch bit and its corresponding lower branch bit. After some straightforward simplifications, the estimated 32-bit RM codeword  $\tilde{\boldsymbol{\lambda}} = (\tilde{\lambda}_0, \tilde{\lambda}_1, \dots, \tilde{\lambda}_{31})$   $\tilde{\boldsymbol{\lambda}} = (\tilde{\lambda}_0, \tilde{\lambda}_1, \dots, \tilde{\lambda}_{31})$  is decoded by:

$$\tilde{\lambda}_k = \begin{cases} 0, & \text{Re}(s_k) + \text{Im}(s_{(k+2)_{32}}) \cdot (1 - 2w_{(k+2)_{32}}^{RM}) \geq 0 \\ 1, & \text{Re}(s_k) + \text{Im}(s_{(k+2)_{32}}) \cdot (1 - 2w_{(k+2)_{32}}^{RM}) < 0 \end{cases} \tilde{\lambda}_k = \begin{cases} 0, & \text{Re}(s_k) + \text{Im}(s_{(k+2)_{32}}) \cdot (1 - 2w_{(k+2)_{32}}^{RM}) \geq 0 \\ 1, & \text{Re}(s_k) + \text{Im}(s_{(k+2)_{32}}) \cdot (1 - 2w_{(k+2)_{32}}^{RM}) < 0 \end{cases}, k = 0, 1, \dots, 31.$$

- 5) The estimated 32-bit RM codeword  $\tilde{\boldsymbol{\lambda}} = (\tilde{\lambda}_0, \tilde{\lambda}_1, \dots, \tilde{\lambda}_{31})$   $\tilde{\boldsymbol{\lambda}} = (\tilde{\lambda}_0, \tilde{\lambda}_1, \dots, \tilde{\lambda}_{31})$  can be decoded by a 3-stage majority-logic decoding. The last 10 bits,  $(\tilde{b}_6, \tilde{b}_7, \dots, \tilde{b}_{15})$ , are decoded from the received code vector  $\tilde{\boldsymbol{\lambda}} = (\tilde{\lambda}_0, \tilde{\lambda}_1, \dots, \tilde{\lambda}_{31})$  in the first stage. These 10 bits are removed from  $\tilde{\boldsymbol{\lambda}}$  to form a modified code vector  $\tilde{\boldsymbol{\lambda}}^{(1)} = \tilde{\boldsymbol{\lambda}} - (0, 0, \dots, 0, \tilde{b}_6, \tilde{b}_7, \dots, \tilde{b}_{15}) \cdot \mathbf{G}$   $\tilde{\boldsymbol{\lambda}}^{(1)} = \tilde{\boldsymbol{\lambda}} - (0, 0, \dots, 0, \tilde{b}_6, \tilde{b}_7, \dots, \tilde{b}_{15}) \cdot \mathbf{G}$ .



- 6) The modified code vector  $\tilde{\lambda}^{(1)}$  has a symmetric structure and it can be used to double-check if the 32-symbol complex sequence  $(s_0, s_1, \dots, s_{31})$  is the **FECFrame** header. The RM autocorrelation of the received modified code vector is computed by:

$$R_{\text{RM}}(k) = \sum_{m=0}^{2^k-1} \sum_{n=0}^{2^{4-k}-1} (2 \cdot \tilde{\lambda}_{m \cdot 2^{5-k}+n}^{(1)} - 1) \cdot (2 \cdot \tilde{\lambda}_{m \cdot 2^{5-k}+2^{4-k}+n}^{(1)} - 1) R_{\text{RM}}(k) = \sum_{m=0}^{2^k-1} \sum_{n=0}^{2^{4-k}-1} (2 \cdot \tilde{\lambda}_{m \cdot 2^{5-k}+n}^{(1)} - 1) \cdot (2 \cdot \tilde{\lambda}_{m \cdot 2^{5-k}+2^{4-k}+n}^{(1)} - 1)$$

The RM symmetry measure is then computed by  $C_{\text{RM}} = \sum_{k=0}^4 |R_{\text{RM}}(k)|$ .

- 6.1) If  $C_{\text{RM}} < T_2$ , go to step 1 and advance symbol index by 1, *e.g.*  $i=i+1$ .  
 6.2) If  $C_{\text{RM}} \geq T_2$ , which means **FECFrame** header is detected, perform step 7.

From simulation results, the performance of setting of  $T_2 = 500$  has a miss-detection rate  $< 10^{-10}$  with respect to a false alarm rate  $< 10^{-10}$  for normal mode in a SNR of 10 dB and high efficiency mode in a SNR of 20 dB.

- 7) The first stage of majority-logic decoding is carried out in step 6. Perform the remaining 2 stages of majority-logic decoding procedure to obtain the 16 information bit.

### Supplementations

- 1) If high efficiency mode is assumed, the major difference is in decoding estimated 32-bit RM codeword  $\tilde{\lambda} = (\tilde{\lambda}_0, \tilde{\lambda}_1, \dots, \tilde{\lambda}_{31})$  in step 4. The estimated 32-bit RM codeword  $\tilde{\lambda} = (\tilde{\lambda}_0, \tilde{\lambda}_1, \dots, \tilde{\lambda}_{31})$  is decoded by computing  $k = 0, 1, \dots, 15$ :

$$\begin{aligned} x_k^{+1} &= \exp \left[ \frac{-\left(\frac{\text{Re}(s_k) - 1/\sqrt{10}}{\sigma^2}\right)^2}{\sigma^2} \right], & x_k^{+3} &= \exp \left[ \frac{-\left(\frac{\text{Re}(s_k) - 3/\sqrt{10}}{\sigma^2}\right)^2}{\sigma^2} \right] \\ x_k^{-1} &= \exp \left[ \frac{-\left(\frac{\text{Re}(s_k) + 1/\sqrt{10}}{\sigma^2}\right)^2}{\sigma^2} \right], & x_k^{-3} &= \exp \left[ \frac{-\left(\frac{\text{Re}(s_k) + 3/\sqrt{10}}{\sigma^2}\right)^2}{\sigma^2} \right] \\ y_k^{+1} &= \exp \left[ \frac{-\left(\frac{\text{Im}(s_k) - 1/\sqrt{10}}{\sigma^2}\right)^2}{\sigma^2} \right], & y_k^{-1} &= \exp \left[ \frac{-\left(\frac{\text{Im}(s_k) + 1/\sqrt{10}}{\sigma^2}\right)^2}{\sigma^2} \right] \\ y_k^{-3} &= \exp \left[ \frac{-\left(\frac{\text{Im}(s_k) + 3/\sqrt{10}}{\sigma^2}\right)^2}{\sigma^2} \right], & y_k^{+3} &= \exp \left[ \frac{-\left(\frac{\text{Im}(s_k) - 3/\sqrt{10}}{\sigma^2}\right)^2}{\sigma^2} \right] \end{aligned}$$

$$\tilde{\lambda}_{2k} = \begin{cases} 0, \log \frac{x_k^{-1} + x_k^{-3}}{x_k^{+1} + x_k^{+3}} + (1 - 2w_{(2k+2)_{32}}^{\text{RM}}) \cdot \log \frac{x_{(k+1)_{16}}^{-1} + x_{(k+1)_{16}}^{+1}}{x_{(k+1)_{16}}^{-3} + x_{(k+1)_{16}}^{+3}} < 0 \\ 1, \log \frac{x_k^{-1} + x_k^{-3}}{x_k^{+1} + x_k^{+3}} + (1 - 2w_{(2k+2)_{32}}^{\text{RM}}) \cdot \log \frac{x_{(k+1)_{16}}^{-1} + x_{(k+1)_{16}}^{+1}}{x_{(k+1)_{16}}^{-3} + x_{(k+1)_{16}}^{+3}} \geq 0 \end{cases}$$

$$\tilde{\lambda}_{2k+1} = \begin{cases} 0, \log \frac{y_k^{-1} + y_k^{-3}}{y_k^{+1} + y_k^{+3}} + (1 - 2w_{(2k+3)_{32}}^{RM}) \cdot \log \frac{y_{(k+1)_{16}}^{-1} + y_{(k+1)_{16}}^{+1}}{y_{(k+1)_{16}}^{-3} + y_{(k+1)_{16}}^{+3}} < 0 \\ 1, \log \frac{y_k^{-1} + y_k^{-3}}{y_k^{+1} + y_k^{+3}} + (1 - 2w_{(2k+3)_{32}}^{RM}) \cdot \log \frac{y_{(k+1)_{16}}^{-1} + y_{(k+1)_{16}}^{+1}}{y_{(k+1)_{16}}^{-3} + y_{(k+1)_{16}}^{+3}} \geq 0 \end{cases}$$

where  $\sigma^2$  is the estimated variance of noise.

- 2) The modified code vector  $\tilde{\lambda}^{(1)} = \tilde{\lambda} - (0, 0, \dots, 0, \tilde{b}_6, \tilde{b}_7, \dots, \tilde{b}_{15}) \cdot G$  has a symmetric structure because for the transmitted 32-bit RM code vector  $\lambda$ ,  $\lambda^{(1)}$  is a linear combination of the first 6 rows of the generator matrix and these 6 rows have a symmetric structure.

### 10.8.3 Alternative FECFrame Header Detection

The DVB-C2 preamble does not signal the start of the **Data Slice** Packets for the **Data Slice** Type 2. Hence, the preamble has to provide the synchronization capabilities on its own. This aim is reached by means of the transmission of the signalling data on the I- and Q-axes of the QPSK diagram (similar for 16-QAM) and the scrambling due to the PRBS sequence. The receiver knows the usage of the QPSK or 16-QAM header due to the Layer 1 -part 2 signalling. Figure 74 shows the synchronization to the headers for the robust mode. As the equalization of the channel has already been done in previous stages of the receiver, it is able to de-map the QPSK symbols. In contrast to the encoding of the header, the delay is now included in the in-phase (I) branch. When the demapped data of both branches is multiplied by each other, the output is exactly the Monic Polynomial Sequence (MPS) sequence for the error-free case. Naturally, the complete sequence is not obtained, as the delay does not completely remove the cyclic shift. However, the length should be sufficient to correlate the signal against the MPS sequence. The detected peak then allows estimating the beginning of a **Data Slice** Packet correctly.

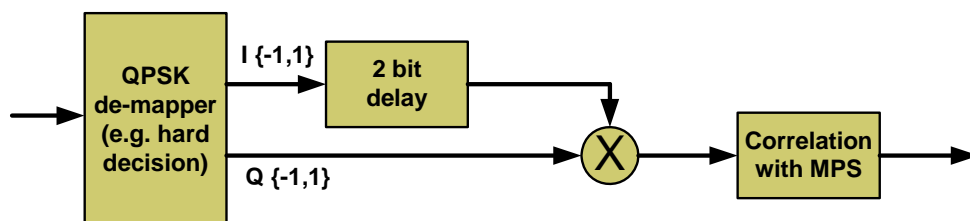


Figure 74: Synchronization scheme for robust header

After the synchronization to the header is reached, the signalling data has to be decoded. One approach is shown in figure 75. In order to achieve the highest performance, the cyclic shift and the MPS sequence has to be removed. For highest performance the complete processing should be done using soft bits (e.g. Log Likelihood Ratio values). This allows for combining the two branches in an optimal way to get optimum diversity. Afterwards, the Reed-Muller decoder removes the remaining errors.

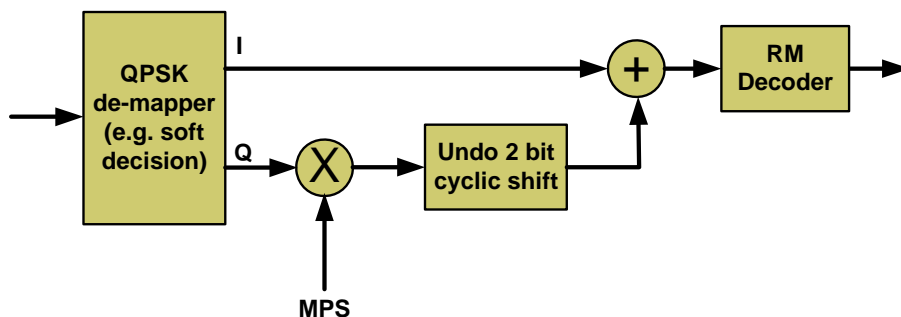


Figure 75: Decoding of the signalling data

Whilst the decoding of the **Data Slice** Packet Headers is required for every **Data Slice** Packet, the synchronization to the header is only required once. Afterwards, the position of the following header can be obtained by means of the signalling data within the header in addition to table 16(b) of ETSI EN 302 769 [i.1], as this allows for the calculation of the actual **XFECFRAME** length.

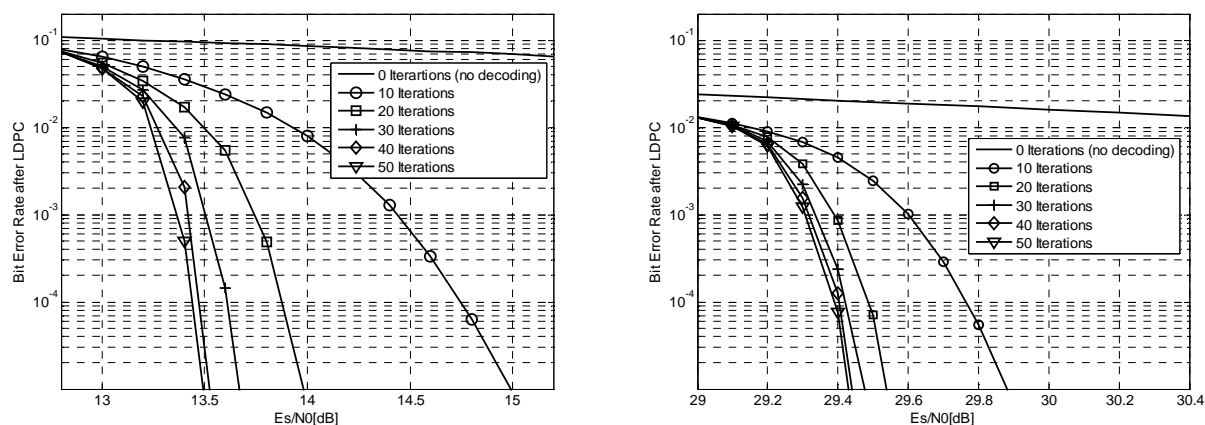


**Figure 76: Calculation of the position of the following Data Slice Packet Header by means of the header data**

## 10.9 LDPC Decoding

The most general decoding algorithm of LDPC code is an iterative message passing algorithm called belief propagation [i.13]. This algorithm is naturally suitable for parallel message computation. The structure of the parity check matrices of the LDPC codes adopted in DVB-C2 implies a partly-parallel decoder architecture.

The performance of LDPC codes can be improved by increasing the number of decoding iterations. Figure 77 shows the performance sensitivity, i.e. the variation of performance with the number of iterations for rate 2/3 with 64-QAM and rate 9/10 with 1 024-QAM, respectively.



**Figure 77: Performance sensitivity for rate 2/3 ( $N_{ldpc}=64\ 800$ ) at 64-QAM (left) and for rate 9/10 ( $N_{ldpc}=64\ 800$ ) at 1 024-QAM (right) on AWGN channel**

The number of decoding iterations per FECFRAME achievable at the receiver for a given hardware complexity can be estimated from the following equation:

$$I = (N_{ldpc} \cdot F_{clk} \cdot P_{dec}) / (C_{code} \cdot E_{mat} \cdot a_{dec}),$$

where:

$N_{ldpc}$ : code length of LDPC code (64 800 or 16 200);

$E_{mat}$ : number of '1's in parity check matrix;

$C_{code}$ : code bit rate (bits/sec);

$F_{clk}$ : decoder clock frequency (Hz);

$P_{dec}$ : parallel order of decoder (typically 360);

$a_{dec}$ : efficiency factor of decoder ( $a_{dec} = 2$  for variable- and check-node calculation with no overhead).

For example, if  $N_{ldpc}/E_{mat} = 0,3$  on average,  $C_{code} = 64$  Mbps,  $P_{dec} = 360$ ,  $a_{dec} = 4$ , and  $F_{clk} = 120$  MHz, then  $I = 50$  iterations/frame.

The maximum rate at which cells are delivered by the **C2 system** depends on the **Data Slice** bandwidth. The maximum **Data Slice** bandwidth of 7,61 MHz will lead to the maximum cell rate of 7,5 Mcell/s per **Data Slice**.

## 10.10 BCH Decoding

For the purpose of removing any possible error floor, the use of bounded-distance decoding [i.14] is sufficient.

## 10.11 Output processing

### 10.11.1 De-/Re-multiplexing

#### 10.11.1.1 Construction of output TS

Construction of the output TS will make use of the following pieces of information:

- SYNCD, SYNC and UPL (see clause 5 in [i.1]): Used to regenerate the user packets (where applicable). See clause 10.11.1.2.
- DNP: used to re-insert the null packets which were deleted in the modulator (see clause 10.11.1.5).
- ISCR: used to calculate the output bit-rate and for fine adjustment of the relative timing of data and common **PLP**s (see clause 10.11.1.6).

#### 10.11.1.2 Mode adaptation

Given the decoded baseband frames from the BCH decoder (see clause 10.10), the receiver should partially reverse the process of mode adaptation and so recover the payload data of the targeted service.

The receiver should first determine whether Normal or High Efficiency Mode is being used. This is indicated by performing an exclusive-OR of the CRC-8 field with 0 or 1. From a single BB-Frame the receiver cannot tell the difference between a correctly received BB-Header using HEM and an erroneous BB-Header using NM (and vice versa). Instead, the receiver should implement a confidence-count mechanism over a number of BB-Frames. If the received CRC consistently differs from the calculated value only in the LSB, then HEM is in use.

The process should only be partially reversed because some information needs to be retained:

- The ISCR values should be kept until the **Transport Stream** has been regenerated.
- The DNP fields should be kept until the deleted null packets have been reinserted, which is done at the output of the de-jitter buffer (see clause 10.11.1.5).

One possible approach is to remove the BB-Headers and reconstitute the stream in a canonical form for processing by the later stages, i.e. a form which does not depend on the particular mode adaptation options in use. An example format could be:

- Sync bytes reinserted;
- CRC-8 removed (if present);
- DNP fields retained, or inserted with a value of zero if DNP is not used;
- ISCR values retained. For packets with no associated field, or if ISSY is not used, these could either "freewheel" (see clause 10.11.1.4) or be set to a "unknown" value.

In principle the mode adaptation in use could change from one BB-Frame to the next, although this is unlikely to happen in practice. Reconstituting the stream in a canonical form provides a simple way to deal with such changes, since each BB-Frame can be processed individually and later processing stages need not be aware of the mode adaptation options that were used.

The only exception is the use of HEM, which should be assumed not to change from BB-Frame to BB-Frame, in order to allow the confidence-count mechanism described above to work correctly.

### 10.11.1.3 Determination of output-TS bit-rate

The receiver needs to know the exact bit-rate of the output **Transport Stream**, in order to be able to output the stream.

If ISSY is used, this can be calculated from the ISCR values. Following null-packet re-insertion, the bit stream will consist of a sequence of TS packets, some of which will have associated ISCR values.

The ISCR values are in units of the elementary time period  $T$ , which will have one of a limited number of values and will be known by the receiver based on the frequency band and channel bandwidth in use. The difference in time between the beginnings of the two packets is therefore known.

The number of packets, and therefore bits, between these two packets will also be known. The TS bit-rate  $R_{TS}$  can therefore be calculated simply by dividing the number of bits  $N_{bits}$  by the time:

$$R_{TS} = \frac{N_{bits}}{(ISCR_2 - ISCR_1) \times T}$$

The time between two packets might not be a whole number of cycles of  $T$ , and consequently there will be some rounding error in this calculation. The residual error can be removed using a feedback loop of the kind described in annex G of [i.2].

If ISSY is not used, the receiver should use a self-balancing buffer as described in annex G of [i.2]. An initial estimate for the TS bit-rate can be made by calculating the maximum useful bit-rate based on the maximum cell rate as described in clause 10.7. However, if null-packet deletion is used, the TS bit-rate may be significantly higher and the feedback loop would need to be able to adjust for this.

Where variable bit-rate **PLPs** with null-packet deletion are used, the buffer occupancy will be constantly changing, making it difficult to use the occupancy to drive a feedback loop reliably. Receiver implementers should assume that ISSY will be used in such cases and need not implement sophisticated strategies to deal with the possibility that it is not.

### 10.11.1.4 De-jitter buffer

The de-jitter buffer itself can be implemented using a FIFO more or less as described in annex C of [i.2].

The ISCR values are carried in the **BBHeader**; they are therefore prone to bit errors. The CRC allows bit errors to be detected, and when this happens the receiver should provide a "flywheel" mechanism to generate the missing values.

### 10.11.1.5 Re-insertion of deleted null packets

Conceptually, the deleted null packets are re-inserted before the de-jitter buffer as illustrated in annex G of [i.2]. The de-jitter buffer would then consist of a simple FIFO as shown. In practice, for reduced memory consumption, the deleted null packets may be re-inserted at the output of the de-jitter buffer. This can be done according to the DNP fields in the BB-Frame. On each TS clock pulse, data should either be read from the de-jitter buffer or regenerated by the null packet re-insertion process. This is the point in the chain at which the variable bit-rate stream carried in a **PLP** becomes a constant bit-rate TS, and the re-insertion of null packets is the mechanism which achieves this.

This implies that information about the deleted null packets should be stored either in the de-jitter buffer itself or in separate storage. One possibility is to retain the DNP fields from the Mode Adaptation. If the de-jitter buffer memory is used, note that this will not be taken into account by the modulator in managing the occupancy of the de-jitter buffer and so a small amount of extra memory may be needed.

### 10.11.1.6 Re-combining the Common and Data PLPs

When a particular input TS has been split and is sent in a data **PLP** and a common **PLP**, the original TS can be in principle recreated in the receiver by combining the contents of the data **PLP** and the common **PLP**, with the limitations given in annex D of [i.2] and described in this clause.

The operations required to re-combine the common and data **PLPs**, together with recommendations for receiver implementations including modification of the PSI/SI, are described in annex D of [i.2].

The DVB-C2 system does not guaranty that the receiver is able to reconstruct the identical input TS of the modulator in case a common **PLP** is used. TS packets carried in the common **PLP** may be time shifted in relation to the Data **PLP** by a few TS packets and additional Null Packets may be present, if the modulator has inserted Null Packet for reasons of reducing the buffering requirements of receivers.

### 10.11.2 Output interface

For TS input, the output of a demodulator shall be the relevant complete **Transport Stream**.

For generic streams, the output format will depend on the type of stream.

## 10.12 Power Saving

The configuration of the **C2 system** allows for different power-saving options. In general, the power consumption of digital circuits linearly depends on the bit-rate the receiver has to demodulate for receiving a specific service. This bit-rate can be reduced by means of the **PLP** concept and the **Data Slice** approach.

Transmitting a service offer in multiple **PLPs** instead of one big **PLP** reduces the overall bit-rate for the LDPC decoder, which is one of the most power-demanding parts in a receiver chip. However, the flexibility in terms of statistical multiplexing between the services (**PLPs**) remains possible, as the different **PLPs** can be statistically multiplexed in a single **Data Slice** while Null Packet Deletion is applied.

## 10.13 Basic considerations for an initial network scan

Many cable networks provide a cable related DVB-SI signalling. Any DVB compliant signalling includes a Network Information Table (NIT) (see ETSI EN 300 468 [i.9]) transmitted in any digital broadcasting **Transport Stream**. A receiver connected to such type of network for the first time is able to perform a very easy installation procedure by looking for a NIT (type cable) and creating the list of available services on basis of the information delivered by that NIT.

In case the delivered NIT is not consistent or even not of the right type (e.g. a satellite type NIT), the receiver will have to perform network scan. All details how to use the DVB SI-System for the installation of a digitalTV receiver with a DVB-C2 frontends are discussed in the Guidelines on implementation and usage of services information (SI), ETSI TS 101 211 [i.26].

# 11 Theoretical Performance

## 11.0 Introduction

This clause presents simulation results indicating the performance of DVB-C2 in a range of cable channels [i.16]. The results are based on the simulation and validation effort carried out during and following the development of the DVB-C2 standard [i.2].

The following information will be presented:

- description of the channel models used for simulations;
- performance simulation results for the whole chain, showing the trade-off between performance and data capacity.

## 11.1 Channel Models

### 11.1.0 Introduction

During the development of the DVB-C2 specification [i.2], different channel models have been used to provide simulated performance results. The channels have been chosen to verify the performance in a wide range of reception conditions.

#### 11.1.1 Additive White Gaussian Noise (AWGN)

In this channel model only white Gaussian noise (AWGN) is added to the signal, and there is only one path.

## 11.1.2 Echo Channel

This model includes two cases of an echo distribution based on the HFC Channel Model as in [i.20] with adapted delay values, the second case being defined as a worst case scenario.

The model is defined by:

$$y(t) = 1 + k \sum_{i=1}^N \rho_i e^{-j\theta_i} x(t - \tau_i)$$

where:

- $x(t)$  any  $y(t)$  are input and output signals, respectively,
- $k = \frac{1}{\sqrt{\sum_{i=1}^N \rho_i^2}}$  and
- the values of the relative power  $\rho_i$ , the delay  $\tau_i$  and the phase  $\theta_i$  are given in table 21.

**Table 21: Relative power, delay and phase values of the two cases of the echo model**

	Power	Delay	Phase
	[dB]	[ns]	[rad]
<b>Case 1</b>	-11	38	0,95
	-14	181	1,67
	-17	427	0,26
	-23	809	1,20
	-32	1 633	1,12
	-40	3 708	0,81
<b>Case 2</b>	-11	162	0,95
	-14	419	1,67
	-17	773	0,26
	-23	1 191	1,20
	-32	2 067	1,12
	-40	13 792	0,81

## 11.2 Simulated System Performance

### 11.2.0 Performance of DVB-C2 over a different channel models

Table 22 shows the simulated performance, assuming perfect channel-estimation, perfect synchronization and without phase noise, of channel coding and modulation combinations, and are subject to confirmation by testing.

Results are given for the Gaussian channel and the echo channel.

To keep simulation times reasonable, most of the results are for a bit error rate of  $10^{-4}$  after LDPC decoding. Some results are also given for the Gaussian channel at a rate of  $10^{-7}$  after LDPC, corresponding to approximately  $10^{-11}$  after BCH. Simulation times are much longer at the  $10^{-7}$  level and results will be added as they become available. Generally the difference between the two is around 0,2 dB in the Gaussian channel, but this difference would probably be greater in the other channels.

To ensure reliable results, the simulations at  $10^{-4}$  were run until at least 50 erroneous FEC blocks had been simulated at the target bit error rate. Furthermore, the LDPC decoding has been configured to apply 100 iterations. For the simulations at  $10^{-7}$ , the following two conditions had to be fulfilled:

- minimum of 100 erroneous FEC blocks; and

- minimum of 1 000 erroneous bits detected.

The DVB-C2 OFDM parameters used for these simulations were chosen as follows: a Guard Interval of 1/64 and the bandwidth 8 MHz. Reserved Tones were not applied. The simulations assumed ideal conditions, i.e. ideal synchronization and ideal channel estimation.

In the simulations, the transmitted signal includes no pilots at all and only OFDM **data symbols** have been considered omitting the Layer 1 Signalling symbols. The values of  $(C/N)_0$  should therefore be corrected for the **FFT size** and pilot pattern in use as described in clauses 9.3 and 9.6 in [i.2].

NOTE: This correction is separate from the penalty for real channel estimation discussed in clause 11.2 and both should be applied.

The simulations assume ideal demapping. However, iterative or "Genie-Aided" demapping as applied for the simulations of DVB-T2 [i.4] has not been used. Thus, deviations of the values for the Gaussian channel with respect to equivalent DVB-T2 channel coding and modulation combinations arise from this fact.

**Table 22: Required raw  $(C/N)_0$  to achieve a given bit error rate of  $10^{-4}$ :  
GI 1/64, bandwidth 8MHz, LDPC block length: 64 800 and 16 200**

	QAM	Code Rate	Effect. CR	Spectral Efficiency	Gaussian Channel	Echo Case 1	Echo Case 2
				[Bit/s/Hz]	[dB]	[dB]	[dB]
<b>LDPC 64 800 (long)</b>	16	4/5		3,19	10,70	11,10	11,60
		9/10		3,59	12,80	13,40	14,00
	64	2/3		3,99	13,40	13,70	14,10
		4/5		4,78	16,00	16,40	16,80
		9/10		5,39	18,40	19,00	19,60
	256	3/4		5,98	19,90	20,20	20,60
		5/6		6,65	21,90	22,30	22,70
		9/10		7,18	23,90	24,50	25,00
	1 024	3/4		7,47	24,60	25,00	25,30
		5/6		8,31	27,10	27,50	27,90
		9/10		8,98	29,40	29,90	30,50
	4 096	4/5		9,56	31,20	31,60	32,00
		5/6		9,97	32,20	32,60	33,20
		9/10		10,78	34,90	35,40	36,30
<b>LDPC 16 200 (short)</b>	16	4/5	7/9	3,07	10,80	11,30	11,70
		9/10	8/9	3,51	12,60	13,30	13,90
	64	2/3	2/3	3,94	13,60	13,90	14,20
		4/5	7/9	4,60	16,10	16,50	16,90
		9/10	8/9	5,27	18,30	18,90	19,40
	256	3/4	11/15	5,78	20,10	20,40	20,80
		5/6	37/45	6,49	22,10	22,50	22,90
		9/10	8/9	7,03	23,80	24,30	24,80
	1 024	3/4	11/15	7,23	24,90	25,20	25,50
		5/6	37/45	8,12	27,30	27,70	28,10
		9/10	8/9	8,79	29,30	29,80	30,40
	4 096	4/5	7/9	9,21	31,60	32,00	32,40
		5/6	37/45	9,74	32,40	32,70	33,30
		9/10	8/9	10,54	34,80	35,30	36,10



Figure 78 shows the results given in table 22.

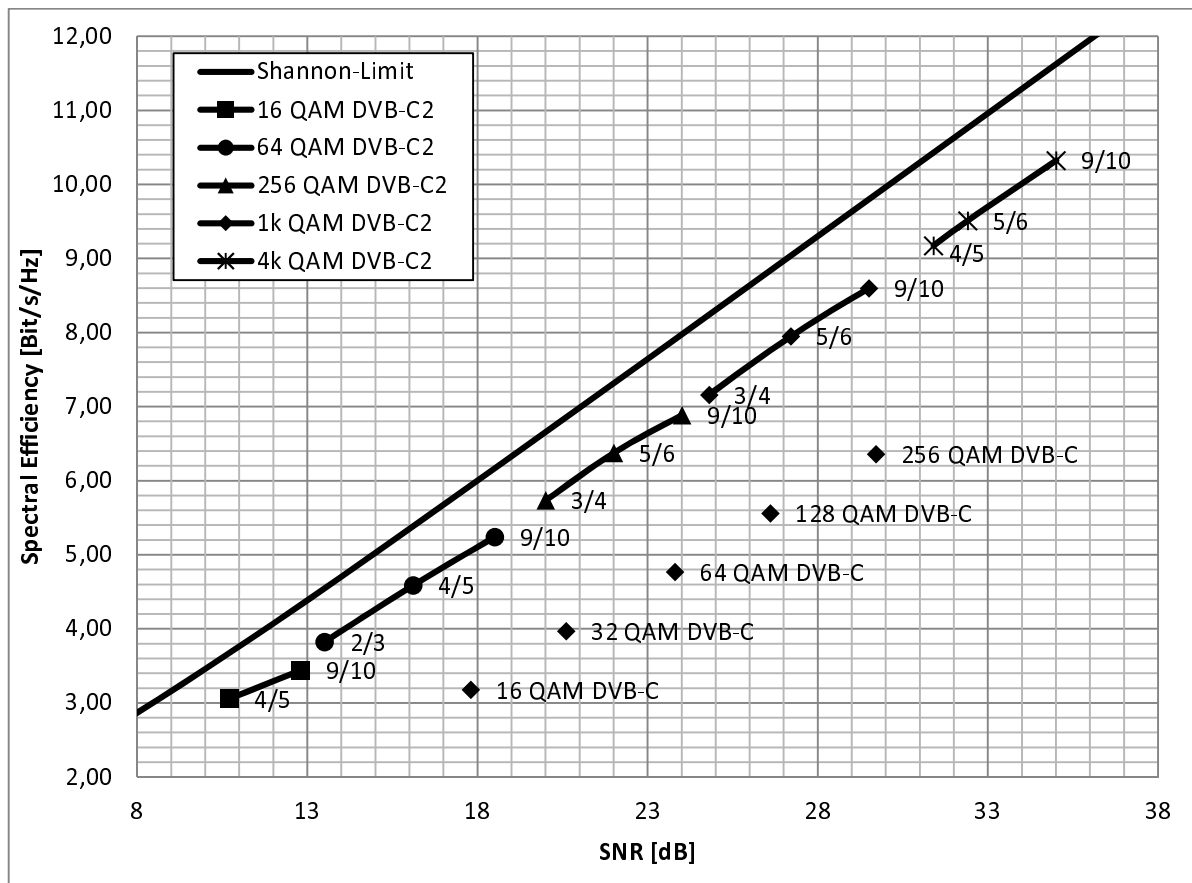


Figure 78: Graphical presentation of the simulation results according to table 21

### 11.2.1 Performance of L1-part2 signalling over an AWGN channel

Simulation was done to analyze the coded performance of L1 signalling part 2 over an AWGN channel. By using the shortening and puncturing procedures described in clause 8.4.5, the L1 signalling bits are first BCH-encoded, and then the BCH encoded bits are LDPC-encoded. Note that the 16K LDPC code with effective code-rate 4/9 is used for encoding of L1 signalling part 2, furthermore, the coded L1 signalling bits are modulated by 16-QAM.

Figure 79 shows the performance of the coded **L1-part2** signalling for a range of BCH information sizes. As shown in figure 79 the performance is almost invariant and stable because of the flexible adaptation of LDPC code rates (see clause 8.4.5.6). Note that the effective LDPC code-rates for **L1-part2** signalling vary as a function of the size of BCH information, as shown in figure 80.

The simulation parameters are as follows:

- size of BCH information: 272, 440, 816, 1 144, 1 470, 1 906, 2 452, 4 758 (bits);
- modulation order: 16-QAM;
- floating-point LDPC decoding (Iteration = 50);
- BCH emulation (error correction capability = 12 bits);
- no time interleaver.

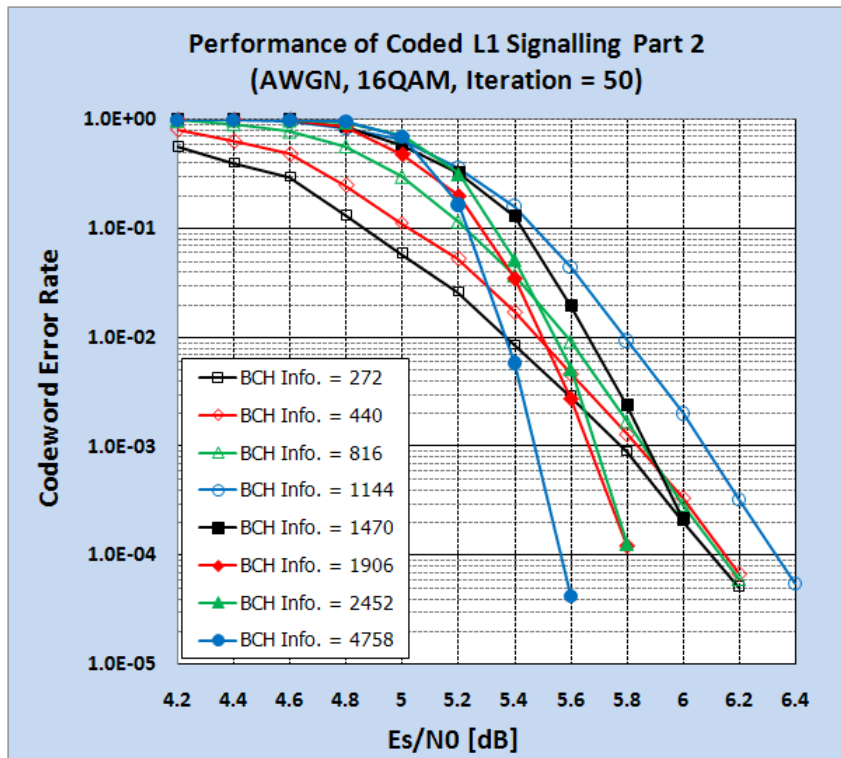


Figure 79: Performance of coded L1-part2 signalling

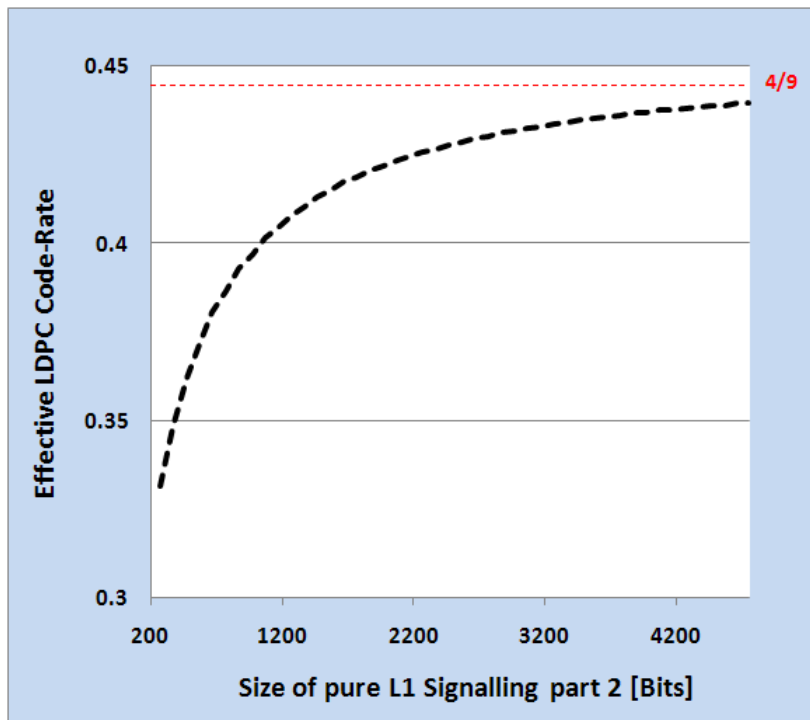


Figure 80: Effective LDPC code-rates for L1-part2 signalling

## 11.2.2 Correction values for pilot boosting

The values of required  $(C/N)_0$  given in table 21 are raw and do not taken into account the reduction in data  $C/N$  resulting from the presence of boosted pilots, since this depends on the pilot pattern in use, i.e. the actual Guard Interval length. Net values of  $C/N$  can be derived from the raw values  $(C/N)_0$  by computing a correction factor  $\Delta_{BP}$  such that:

$$\frac{C}{N} = \left( \frac{C}{N} \right)_0 + \Delta_{BP}$$

This correction factor is calculated by the following formula:

$$\Delta_{BP} = 10 \cdot \log_{10} \left( \frac{K_{\text{data}} + K_{\text{CP}} \cdot A_{\text{CP}}^2 + K_{\text{SP}} \cdot A_{\text{SP}}^2}{K_{\text{data}} + K_{\text{CP}} + K_{\text{SP}}} \right)$$

Where:

- $K_{\text{data}}$  Number of **data cells** per OFDM symbol;
- $K_{\text{CP}}$  Number of Continual Pilots per OFDM symbol;
- $K_{\text{SP}}$  Number of Scattered Pilots per OFDM symbol;
- $A_{\text{CP}}$  Amplitude of the continual pilot cells;
- $A_{\text{SP}}$  Amplitude of the scattered pilot cells.

The exact value of  $\Delta_{BP}$  depends on the number of pilots, but will not exceed 0,5 dB.

## 11.3 Receiver Performance Requirements

### 11.3.0 Introduction

This clause expands on the discussion of theoretical system performance in clause 11.2 to consider the target receiver performance requirements for AWGN, echoes and sensitivity

### 11.3.1 Receiver Performance Requirements in an Ideal System

Table 23 provides a more direct comparison between 'ideal' performance and required receiver performance than shown earlier in clause 11. This assumes:

$$C/N = C/N(\text{raw}) + A + B \text{ [dB]}$$

- $C/N(\text{raw})$  = required  $C/N$  for  $\text{BER} = 10^{-6}$  before BCH decoding
- $A = 0,1\text{dB}$  assumed additional  $C/N$  to achieve the  $\text{BER} = 10^{-7}$  before BCH decoding (assumed QEF **Transport Stream** after BCH decoding)
- $B$  = correction for pilot boosting (see clause 11.2.2)

This allows comparison to be made at the same  $\text{BER} = 10^{-7}$  and compensates for pilot boosting so that the theoretical values for each guard interval can be included for comparison.

Table 23: SNR Receiver Performance Requirements @ BER =  $10^{-7}$  assuming an Ideal System

	QAM	Code Rate	Gaussian Channel		EchoCase 1		EchoCase 2	
			GI 1/128	GI 1/64	GI 1/128	GI 1/64	GI 1/128	GI 1/64
			dB	dB	dB	dB	dB	dB
LDPC 64 800 (long)	16	4/5	11,2	11,3	11,6	11,7	12,1	12,2
		9/10	13,3	13,4	13,9	14,0	14,5	14,6
	64	2/3	13,9	14,0	14,2	14,3	14,6	14,7
		4/5	16,5	16,6	16,9	17,0	17,3	17,4
		9/10	18,9	19,0	19,5	19,6	20,1	20,2
	256	3/4	20,4	20,5	20,7	20,8	21,1	21,2
		5/6	22,4	22,5	22,8	22,9	23,2	23,3
		9/10	24,4	24,5	25,0	25,2	25,5	25,6
	1 024	3/4	25,1	25,2	25,5	25,6		
		5/6	27,6	27,7	28,0	28,1		
		9/10	29,9	30,0	30,4	30,5		
	4 096	4/5	31,7	31,8	32,1	32,2		
5/6		32,7	32,8	33,1	33,2			
9/10		35,4	35,5	35,9	36,0			
LDPC 16 200 (short)	16	4/5	11,3	11,4	11,8	11,9	12,2	12,3
		8/9	13,1	13,2	13,8	13,9	14,4	14,5
	64	2/3	14,1	14,2	14,4	14,5	14,7	14,8
		4/5	16,6	16,7	17,0	17,1	17,4	17,5
		8/9	18,8	18,9	19,4	19,5	19,9	20,0
	256	3/4	20,6	20,7	20,9	21,0	21,3	21,4
		5/6	22,6	22,7	23,0	23,1	23,4	23,5
		8/9	24,3	24,4	24,8	24,9	25,3	25,4
	1 024	3/4	25,4	25,5	25,7	25,8		
		5/6	27,8	27,9	28,2	29,3		
		8/9	29,8	29,9	30,3	30,4		
	4 096	4/5	32,1	32,2	32,5	32,6		
5/6		32,9	33,0	33,2	33,3			
8/9		35,3	35,4	35,8	35,9			

NOTE: The combinations of 1 024-QAM/4 096-QAM with echo case 2 are considered to be mutually exclusive for minimum performance requirements and therefore no figures are specified for these combinations in the above table and other performance tables in this clause.

### 11.3.2 Backstop Noise Definition

Backstop noise is the logarithmic sum of all the noise components in a C2 transmission system up to the output of the A/D converter in the C2 demodulator. The present document proposes two levels of backstop noise to accommodate current and future networks respectively.

For current networks supporting modulation and coding formats up to 1 024-QAM 9/10 code rate:

**BackStop 1: 35,8 dB SNR**

For future networks / advanced systems optimized for the maximum modulation order in DVB-C2 of 4 096-QAM.

**BackStop 2: 38,1 dB SNR**

The following tables show the assumptions made for the types of contributing components that make up backstop noise and their relative level to the wanted OFDM signal. The logarithmic (power) sum of items (1), (2), (3) & (4) is used to define the total backstop noise levels.

**Table 24: Backstop 1 Components**

Item	Backstop Component	Units	Value
1	ADC Quantization Noise	dB	50,7
2	Transmitter Thermal Noise	dB	45
3	Transmitter Phase Noise	dB	48
4	Receiver phase noise and thermal noise	dB	36, 8
NOTE: In (1) - ADC is assumed to be 12-bit resolution with some headroom allowed for the OFDM input signal (2), (3) are figures from available test equipment (4) Example receiver performance known from prototypes.			

**Table 25: Backstop 2 Components**

Item	Backstop Component	Units	Value
1	ADC Quantization Noise	dB	50,7
2	Transmitter Thermal Noise	dB	45
3	Transmitter Phase Noise	dB	48
4	Receiver phase noise + thermal noise	dB	40,1
NOTE: As for Table 24 description above with the following changes: (4) A possible future receiver performance with improvements to noise performance from (2).			

### 11.3.3 Performance Calculation Method

The method of calculation for the C2 performance figures is based around the NorDig receiver specification [i.20] method for T2 with some minor modifications.

$C/N = C/N(\text{raw}) + A + B + C + D + E$  [dB], where:

- $C/N(\text{raw})$  = required C/N for BER =  $10^{-6}$  before BCH decoding
- A = 0.1dB assumed additional C/N to achieve the BER= $10^{-7}$  before BCH decoding (assumed QEF **Transport Stream** after BCH decoding)
- B = correction for pilot boosting
- C = 1,0 dB assumed loss due to real channel estimation, imperfect LDPC decoding and other imperfections not considered part of the back-stop noise
- D = 0,5 dB assumed loss due to non-ideal demodulator performance across range of code rates.
- E = additional C/N term corresponding to a back-stop noise level (see clause 11.3.2). This term is derived by first calculating the sum of all terms except E and then checking how much degradation is caused by the backstop noise level. The term E is identical to this degradation

Note for DVB-C2 the required C/N for QEF and for error-free video is expected to be virtually identical due to the sharp waterfall characteristic of the LDPC+BCH decoding. Note also there is no separate definition for distortion components in this calculation method.

## 11.3.4 Performance Requirements for AWGN and Echoes

Table 26: Receiver Performance Requirements for AWGN and Echoes

	QAM	Code Rate num/denom	Gaussian Channel		EchoCase 1		EchoCase 2	
			GI 1/128	GI 1/64	GI 1/128	GI 1/64	GI 1/128	GI 1/64
			dB	dB	dB	dB	dB	dB
<b>LDPC 64 800 (long)</b>	16	4/5	12,7	12,9	13,1	13,3	13,6	13,8
		9/10	14,8	15,0	15,4	15,6	16,0	16,2
	64	2/3	15,4	15,6	15,7	15,9	16,1	16,3
		4/5	18,4	18,2	18,4	18,6	18,8	19,0
		9/10	20,5	20,7	21,1	21,3	21,7	21,9
	256	3/4	22,0	22,2	22,4	22,5	22,8	23,0
		5/6	24,1	24,3	24,6	24,8	25,0	25,2
		9/10	26,3	26,5	27,0	27,2	27,6	27,8
	1 024	3/4	27,1	27,3	27,6	27,8		
		5/6	30,1	30,3	30,6	30,8		
		9/10	33,3	33,6	34,1	34,4		
	4 096	4/5	35,4	35,7	36,1	36,4		
		5/6	36,4	36,7	37,1	37,4		
		9/10	42,8	43,6	45,2	46,5		
	<b>LDPC 16 200 (short)</b>	16	4/5	12,8	13,0	13,3	13,5	13,7
8/9			14,6	14,8	15,3	15,5	15,9	16,1
64		2/3	15,6	15,8	15,9	16,1	16,2	16,4
		4/5	18,1	18,3	18,5	18,7	18,9	19,1
		8/9	20,4	20,6	21,0	21,2	21,5	21,7
256		3/4	22,2	22,4	22,6	22,8	23,0	23,2
		5/6	24,4	24,6	24,8	25,0	25,2	25,4
		8/9	26,2	26,6	26,8	27,0	27,3	27,6
1 024		3/4	27,5	27,7	27,8	28,0		
		5/6	30,4	30,6	30,9	31,1		
		8/9	33,2	32,4	34,0	34,3		
4 096		4/5	35,9	36,2	36,6	36,9		
		5/6	36,7	37,0	37,3	37,6		
		8/9	42,4	43,2	44,4	45,8		

NOTE: Table 26 assumes ideal demapping and no iterative ('Genie-Aided') demapping in an 8 MHz reception bandwidth as defined in clause 11.2.

The figures in Table 26 can be used as performance targets for C2 receiver evaluations.

### 11.3.5 Performance Requirements for Sensitivity

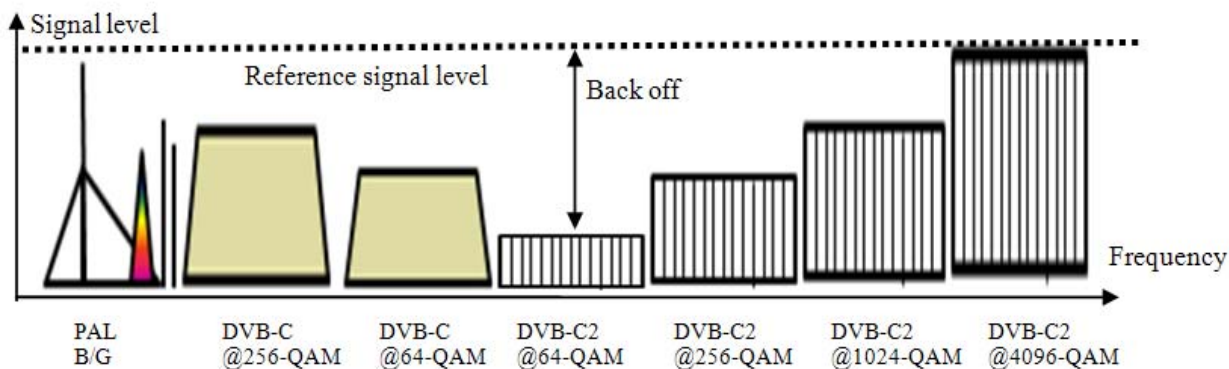
**Table 27: Receiver Performance Requirements for Sensitivity**

	QAM	Code Rate num/denom	Gaussian Channel	
			GI 1/128	GI 1/64
			dB	dB
<b>LDPC 64 800 (long)</b>	16	4/5	24,3	24,5
		9/10	26,4	26,8
	64	2/3	27,0	27,2
		4/5	29,6	29,8
		9/10	32,1	32,3
	256	3/4	33,6	33,8
		5/6	35,7	35,9
		9/10	37,9	38,1
	1 024	3/4	38,7	38,9
		5/6	41,7	41,9
		9/10	44,9	33,6
	4 096	4/5	47,0	47,3
		5/6	48,0	48,3
		9/10	54,4	55,2
	<b>LDPC 16 200 (short)</b>	16	4/5	24,4
8/9			26,2	26,4
64		2/3	27,2	27,4
		4/5	29,7	29,9
		8/9	32,0	32,2
256		3/4	33,8	34,0
		5/6	35,9	36,1
		8/9	37,8	38,0
1 024		3/4	39,0	39,3
		5/6	41,9	42,2
		8/9	44,7	45,0
4 096		4/5	47,5	47,8
		5/6	48,3	48,6
		8/9	54,0	54,7

Impedance of 75 ohm and tuner noise figure of 8 dB [i.24] is assumed in table 27. Figures are defined as the RF signal level at the input to the receiver RF tuner. The receiver shall receive error free video at these input signal levels to meet the defined requirements.

### 11.3.6 Requirements for Adjacent Channel Interference (ACI)

Due to the different SNR requirements of the different analogue and digital modulation schemes used in CATV networks the requirements for adjacent channel interference performance are usually defined as additional protection ratio with respect to the nominal signal levels used in the network. Figure 81 shows the principle of the different signal levels and the so-called backoff with respect to a reference signal level. A receiver meeting the adjacent performance requirements shall be able to decode a wanted signal in a configuration where the signal levels of the adjacent channels are increased by the additional required protection margin with respect to the nominal adjacent signals level settings.



**Figure 81: Example for a nominal setting of signal levels for different types of signals**

The adjacent channel performance requirements of digital receivers are usually defined in CPE specifications by the cable operators, who are purchasing CPEs compliant with the characteristics of their network. In Annex E of the present document the methodology for adjacent channel performance measurements performed during a DVB-C2 Plug Fest with first prototype receiver implementations is presented.

The nominal levels of DVB-C2 signals for ACI measurement are chosen to both meet the minimum required level according to the IEC 60728-1 [i.18] specification at the wall outlet and to meet the minimum required SNR for the specific modulation and coding selected for the wanted channel. The example signal levels in table 28 below meet these requirements.

**Table 28: Example Signal Levels for Wanted and Adjacent Channels**

Standard	Modulation	Signal Level (dB $\mu$ V)	Backoff to Analogue PAL (dB)
DVB-C2	4 096-QAM	66	-2
PAL-B/G	Analogue	64	-
DVB-C2	1 024-QAM	60	4
DVB-C	256-QAM	60	4
DVB-C2	256-QAM	54	10
DVB-C	64-QAM	54	10
DVB-C2	64-QAM	48	16
DVB-C2	16-QAM	42	22

Baseline assumptions for signal levels:

- 256-QAM DVB-C is set to 4 dB below PAL system level.
- 1 024-QAM DVB-C2 is set to the same level as 256-QAM DVB-C.
- 6 dB change in backoff is used for each increment/decrement of QAM modulation order

The test procedure assumed for ACI measurements is as follows:

- Configure signals with a centre DVB-C2 signal of desired modulation and coding and an upper and lower adjacent signal of the same type - e.g. DVB-C2, DVB-C or PAL B/G
- Set the levels of the wanted and adjacent signals to the nominal levels defined in the above table
- Adjust the level of the adjacent signals together keeping the wanted signal level fixed to find the maximum possible level of the adjacent signals before video/packet errors occur in the received signal
- The ACI protection ratio for the given test configuration is calculated as the maximum dB difference between the wanted and adjacent signal levels before video/packet errors occur. A dB margin value can also be calculated as the difference between the nominal ACI signal level and the maximum ACI level before video/packet errors occur.



Measurements have been performed based on the above signal level assumptions and test procedure for adjacent channel signals.

## 12 Examples of Possible Use of the System

### 12.0 Introduction

The examples described below present practical implementation scenarios which should demonstrate the plurality of functionalities the DVB-C2 standard [i.2] supports including applications for multiplexing of **PLPs** and **Data Slices**, **PLP** bundling, adaptive/variable coding, and modulation etc. Although the DVB-C2 physical layer provides a transparent mechanism for an end-to-end delivery of all input data stream formats, the examples describe only applications for a transmission based on MPEG2 **Transport Streams** and IP streams, which both constitute the data transmission formats usually used in cable networks.

### 12.1 Network Scenarios

#### 12.1.0 Introduction

The network scenarios introduced hereafter explain how DVB-C2 can be implemented to support different kind of signal processing (so-called transmodulation) from satellite and terrestrial networks to cable. Furthermore it is explained how the new technology could be introduced in cable networks and how the migration from DVB-C to DVB-C2 could be performed.

#### 12.1.1 Methods of signal conversion in cable head ends

##### 12.1.1.0 Introduction

This clause describes methods of signal processing, which are typically implemented in regional cable head ends applying the conversion of satellite signals into - in this case DVB-C2 based - cable signals. Two scenarios are subsequently introduced: (see clause 12.1.1.1) the general scenario of an efficient and very flexible signal conversion and (see clause 12.1.1.2) the scenario of the transparent conversion of a complete **Transport Stream**.

##### 12.1.1.1 Efficient signal conversion from satellite or terrestrial link to DVB-C2

This clause describes how signals which are received at a head end either via DVB-S or DVB-S2 transponders or via DVB-T or DVB-T2 channels or via any combination of both can be converted efficiently to a DVB-C2 signal.

Table 29 shows relevant technical parameters of an example for a conversion of six DVB-S2 satellite signals received via six individual satellite transponders to a single DVB-C2 signal of 32 MHz bandwidth. Four of the six DVB-S2 transponders provide a bit rate of 42,66 Mbit/s each and two of the six transponders provide a bit rate of 49,39 Mbit/s each. The four DVB-S2 signals are transmitted with 22 MBaud (Mega Symbols per second) and use an 8-PSK combined with a FEC code rate of 2/3 resulting in 42,66 Mbit/s. For the two satellite signals with the slightly higher bit rate of 49,39 Mbit/s the following technical parameters are applied: 27,5 MBaud, QPSK, FEC code rate of 9/10.

**Table 29: Signal parameters exemplarily used for the conversion of signals received via six DVB-S2 transponders to a single DVB-C2 signal**

Satellite transponders			DVB-C2 channel		
Transponders	6		<b>Data Slices</b>	6	
Transponder bit rate	4 x 42,66	2 x 49,39	Slice bit rate (Mbit/s)	4 x 42,66	2 x 49,39
			Total C2 channel bit rate (Mbit/s)	4 x 42,66+ 2 x 49,39 = 269,42	
Transponder bandwidth (MHz)	5 x 27	33	C2 channel bandwidth (MHz)	32	
Modulation	8PSK 3 bit/symbol	QPSK 2 bit/symbol	Modulation (CCM)	1 024-QAM 10 bit/symbol	1 024-QAM 10 bit/symbol
Code rate	2/3	9/10	Code rate	9/10	9/10
Symbol rate	22	27,5			

In the following some calculations are explained more detailed. Because in this example the **Transport Streams** from satellite transponders are retransmitted by equivalent number of **Data Slices** first the number of necessary OFDM carriers  $N_c$  per **Data Slice** should be calculated by the formula given below:

$$N_c = \frac{R_{Sat}}{\Delta f * N * G * C1 * C2 * SP * CP} \quad (26)$$

- $R_{Sat}$  = input **Transport Stream** data rate, in the example above 4 DVB-S2 transponders with 42,66 Mbit/s and two DVB-S2 transponders with 49,39 Mbit/s. According to [i.6]  $R_{Sat}$  is calculated in the formulas shown below;

$R_{Sat}$  = Symbolrate x M x LDPC-code rate x BCH-code rate x pilot loss (if used) from DVB-S2 transponder (2);

$R_{Sat}$  = 22 MBaud x 3 x 2/3 x (43 040/43 200) x 0,973 = 42,66 Mbit/s in case of the DVB-S2/8PSK transponders in table 1 in [i.6];

$R_{Sat}$  = 27,5 MS/s x 2 x 9/10 x (58 192/58 320) = 49,39 Mbit/s in case of the DVB-S2/QPSK transponders in table 1 in [i.6];

The values (43 040/43 200) and (58 192/58 320) are the BCH rates due to the code rates of 2/3 and 9/10. The figure 0,973 assigns the reduction factor due to pilots, which are recommended to be transmitted in the DVB-S2 transponder, if certain modulation schemes and code rates are applied (also for 8 PSK, code rate 2/3).

- $\Delta f$ : Carrier spacing (1/448 us for 8 MHz raster);
- $N$ : bits/s/Hz, in the example 10 bit/s/Hz for 1 024-QAM;
- $C1$ : Code rate of LDPC FEC encoder, in the example 9/10;
- $C2$ : Code rate of outer BCH coder, in the example (58 128/58 320), if  $C1=9/10$  for LDPC coder;
- $G$  = (1-GI) x GI, where  $G=128/129$  for  $GI=1/128$ ;
- $SP$  = (1-PP) x PP, where  $PP=1/96$  ( $GI=1/128$ ), which results in  $SP=95/96$ ;
- $CP$  = 0,99 with 1 % allotment of continual pilots;
- $PO$  = 448/449 for preamble overhead.

Due to formula (26) and the values mentioned above the number of carriers  $N_c$  is:

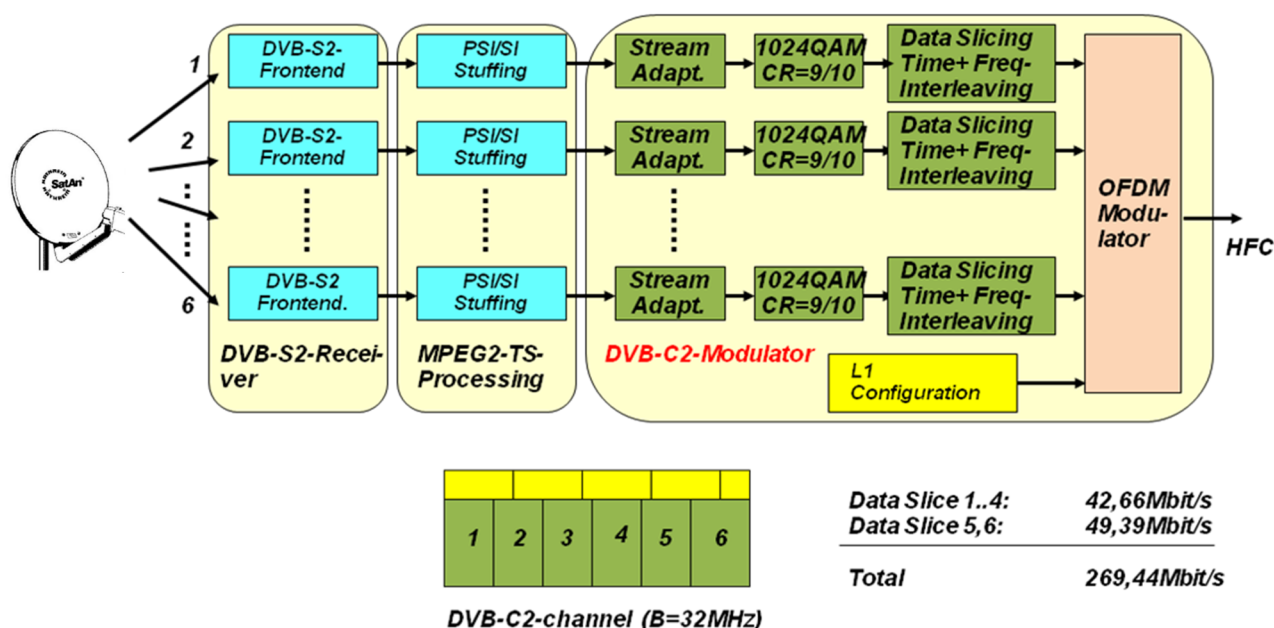
$$N_c = \frac{42.7\text{Mbit/s} * 448\text{us}}{10 * (128/129) * (9/10) * (58128/58320) * (95/96) * 0.99 * (448/449)} = 2208 \quad (27)$$

and  $N_c = 2 544$  carriers for  $R_{Sat} = 49,39$  Mbit/s.

If the Guard Interval is equal to 1/128 of the symbol length, the total number of OFDM carriers calculates to:

$$k_{SC,total} = 4 * 2 208 + 2 * 2 544 = 13 920.$$

The spectral efficiency achieved by this signal constellation is equal to 8,4 bit/s/Hz, which corresponds to a bit rate of 269,44 Mbit/s within 32 MHz channel bandwidth. The difference to the maximal possible figure of 8,6 bps/Hz corresponding to a bit rate of 274,76 Mbit/s within 32 MHz channel bandwidth is very small and due to the reduced number of used carriers as described above. It should be noted that the example examined does not consider the feature zero frame deletion which could be applied in the input stream adaption block. A block diagram of a signal conversion unit is shown in figure 82.



**Figure 82: Block diagram of a headend processing unit applying an efficient signal conversion of 6 DVB-S2 signals to a 32 MHz wide DVB-C2 signal**

In table 30 details of **Data Slice** configuration are shown. The 32 MHz C2 channel is inserted between 302 MHz and 334 MHz and centred to 318 MHz to achieve equal frequency distance to lower and higher adjacent channels.

**Table 30: Configuration of start frequency and Data Slices of the 32 MHz C2 channel for retransmission of satellite transponders**

Parameter	Value	Comment
Centre frequency [MHz]	318	
OFDM carrier spacing	1/448 us	Due to 8 MHz channel raster
$D_x$	24 OFDM carriers	Due to $GI=1/128$
Centre frequency (multiple of $D_x$ ) [MHz]	318	Aligned to $D_x$ raster
Number of OFDM carriers	13 920 + 1	Within the 32 MHz C2 channel, including the upper edge pilot
Start frequency [MHz]	302,46 MHz	$318 \text{ MHz} - (13\ 920/2) \times (1/448 \text{ us})$
Number of carriers		
Data slice 1 to 4	2 208	Width 1 to 4 = $2\ 208/448 \mu\text{s} = 4,93 \text{ MHz}$
Data slice 5 to 6	2 544	Width 5 to 6 = $2\ 244/448 \mu\text{s} = 5,68 \text{ MHz}$
<b>Data Slice</b> tuning positions [MHz]		
<b>Data Slice</b> 1 to 4	306,3 309,9 314,8 319,7	The tuning positions of <b>Data Slices</b> 1 and 6 shall have a distance of at least 1704 OFDM carriers from start and end of the <b>C2 system</b> , the tuning positions of <b>Data Slices</b> 2 to5 are in the middle of <b>Data Slice</b> 2...5
<b>Data Slice</b> 5 to 6	325 329,7	

Table 31 shows all **L1-part2** signalling parameters. The coding schemes of the different parameters are according to clause 8.3 of [i.2].

**Table 31: L1 signalling for the DVB-S2/DVB-C2 retransmission example**

Parameter	Value (transmission format)	Comment
NETWORK_ID	0000 0000 0000 0000	Network identifier
C2_SYSTEM_ID	0000 0000 0000 0000	<b>C2 system</b> identifier
START_FRQUENCY	000000100001000101010000	135 504: value of table 29 multiplied with 448 us (1/carrier spacing)
C2_BANDWIDTH	0000001001000100	$13\ 920/24 = 580$ due to calculated number of carriers. Total $580 \times 24 + 1 = 13\ 921$ carriers due to insertion of the highest carrier frequency edge pilot

Parameter	Value (transmission format)	Comment
GUARD_INTERVAL	00	GI = 1/128
C2_FRAME_LENGTH	0111000000	448 <b>data symbols</b> in one <b>C2 frame</b>
L1_PART_2_CHANGE_COUNTER	0000 0000	No changes between <b>C2 frames</b> , because one constant data rate multiplex (MPEG2-TS) in each <b>Data Slice</b> . Bit rate of those MPEG2-TS remain constant over time
NUM_DSLICE	0000 0110	6 <b>Data Slices</b> within the C2 channel
NUM_NOTCH	0000	No notch
DATA_SLICE_ID ( <b>Data Slice</b> 1 to 6)	0000 0000.....0000 0101	identifier for <b>Data Slice</b> ("0"...."5")
DSLICE_TUNE_POS ( <b>Data Slice</b> 1 to 6)	0000001000111 0000010001010 0000011100110 0000101000010 0000110100101 0000111111101	71 (3 408/2/24); tune pos = 1 704 <sup>th</sup> OFDM carrier 138 (3 x 2 208/2/24); tune pos. = 3 312 <sup>th</sup> OFDM carrier 230 (5 x 2 208/2/24); tune pos = 5 520 <sup>th</sup> OFDM carrier 322 (7 x 2 208/2/24); tune pos = 7 728 <sup>th</sup> OFDM carrier 421 (4 x 2 208+2 544/2)/24; tune pos = 10 104 <sup>th</sup> OFDM carrier 509 (13 920-1 704)/24; tune pos (13 920-1 704) <sup>th</sup> OFDM carrier
DSLICE_OFFSET_LEFT/RIGHT		
<b>Data Slice 1</b>	10111001 / 00010101	-71/+21 for the first <b>Data Slice</b> (min 1704 OFDM carrier distance from the start of the <b>C2 system</b> )
<b>Data Slice 2 to 4</b>	11010010 / 00101110	-46/+46 ( $\pm 2$ 208/2/24); $\pm 1$ 104 carriers left and right from tune pos
<b>Data Slice 5</b>	11001011 / 00110101	-53/+53 ( $\pm 2$ 544/2/24); $\pm 1$ 272 carriers left and right from the tuning position
<b>Data Slice 6</b>	11011101 / 01000111	-35/+71 for the last <b>Data Slice</b> ; minimum 1 704 OFDM carriers distance from the end of the <b>C2 system</b>
DSLICE_TI_DEPTH( <b>Data Slice</b> 1 to 6)	01	4 OFDM symbols
DSLICE_TYPE ( <b>Data Slice</b> 1 to 6)	0	Type 1; single <b>PLP</b> (constant bitrate MPEG2-TS) per <b>Data Slice</b>
DSLICE_CONST_CONF ( <b>Data Slice</b> 1 to 6)	1	<b>Data Slice</b> does not change
DSLICE_LEFT_NOTCH ( <b>Data Slice</b> 1 to 6)	0	No notch
DSLICE_NUM_PLP	000000001	1 <b>PLP</b> per <b>Data Slice</b>
PLP_ID ( <b>Data Slice</b> 1 to 6)	0	<b>PLP</b> identifier ( "0")
PLP_BUNDLED ( <b>Data Slice</b> 1 to 6)	0	Not bundled with <b>PLPs</b> of other <b>Data Slices</b>
PLP_TYPE ( <b>Data Slice</b> 1 to 6)	10	Normal data <b>PLP</b>
PLP_PAYLOAD_TYPE ( <b>Data Slice</b> 1 to 6)	00011	MPEG2-TS
PLP_START ( <b>Data Slice</b> 1 to 6)	00000000000010	First complete <b>XFECFrame</b> starts from 2 <sup>nd</sup> <b>data cell</b> of this <b>Data Slice</b>
PLP_FEC_TYPE	1	This <b>PLP</b> uses 64 K LDPC
PLP_MOD	100	This <b>PLP</b> uses 1 024-QAM modulation
PLP_COD	101	This <b>PLP</b> uses code rate 9/10
PSI/SI_REPROCESSING	1	Cable NIT, inserted in every MPEG2-TS, therefore PSI/SI reprocessing

### 12.1.1.2 Transparent signal conversion and MPEG2 Transport Stream processing

A very typical scenario which is currently implemented with DVB-C systems arranges for a complete and transparent signal conversion without any MPEG2 **Transport Stream** processing applied between receiver and transmitter unit. The transparent conversion can easily be achieved by replacing the DVB-C transmitter unit by a corresponding DVB-C2 unit. This simple and straightforward approach however does not exploit the full potential of DVB-C2 since the same constraints applicable to the traditional DVB-C signal conversion unit are also applied to the modern DVB-C2 based unit. For example, it is expected that the time for channel search required by the receiver cannot be shortened, although the tuning procedure itself has been optimized in the **C2 system**.

However, more feasible in practice is the scenario of an efficient signal conversion as referred to above. The block diagram of a conversion unit is shown in figure 78. As already for DVB-C, also for DVB-C2 a PSI/SI processing is primarily necessary to generate the cable NIT which contains the DVB-C2 delivery system descriptor. The complete cable NIT should be transmitted in each **PLP** in order to avoid, for instance, a long-lasting channel search of the receivers, an effect already known from the DVB-C system.

In a scenario considering a mixed channel occupation of DVB-C and DVB-C2 signals in the single cable network, it may be advantageous to incorporate the complete cable NIT in a so called "home channel" or "Barker channel" to which the cable STB always tunes first. If the signal transmitted through this channel complies with DVB-C, the utilization of such a method can prevent a DVB-C CPE from tuning to a DVB-C2 channel which may cause malfunctions.

Bit-rate stuffing at MPEG2 **Transport Stream** level is not necessary in the application described above, if the original MPEG2 **Transport Stream** received via a satellite transponder or a terrestrial channel will be converted to a DVB-C2 signal as a whole. The reason for this phenomenon is that the **Data Slice** bandwidth can be flexibly adapted to the input bit rate as shown in clause 12.1.1.1. However, if some program filtering is applied in the receiver unit to suppress programs which should not be transmitted in the cable network, stuffing is foreseen to get a constant bit rate for the MPEG2 **Transport Stream** at the DVB-C2 modulator input as required by the DVB-C2 standard [i.2]. Therefore, in this case zero frame deletion within the DVB-C2 modulator may be useful in terms of the highest efficiency of the utilization of available capacity. Of course, zero frame deletion can also be applied for **Transport Streams** coming unchanged from a satellite or terrestrial channel. But the number of zero frames is expected to be rather low for those streams. Therefore, the possible gain in transmission capacity may be neglected.

### 12.1.1.3 Example for a configuration with a narrowband notch within a DVB-C2 signal

This example gives the signalling parameters of DVB-C2 signal with a narrowband notch, where those subcarriers which transmit the **preamble symbols** and which are located in the range of the narrowband notch are switched off. In such a configuration the notch weakens the error protection of the preamble as the information carried by the switched off subcarriers is not available for the receiver.

In table 32 details of **Data Slice** configuration are shown. The 32 MHz C2 channel is inserted between 302 and 334 MHz and centred to 318 MHz to achieve equal frequency distance to lower and higher adjacent channels. One narrowband notch is inserted at the high frequency end of data Slice 5 at 327,86 MHz, with a notch width of 23 subcarriers. Apart from this, the satellite retransmission example of clause 12.1.1 is reused.

**Table 32: Configuration of start frequency and Data Slices of the 32 MHz C2 channel including one narrowband notch**

Parameter	Value	Comment
Centre frequency [MHz]	318	
OFDM carrier spacing	1/448 us	Due to 8 MHz channel raster
$D_x$	24 OFDM carriers	Due to $GI=1/128$
Centre frequency (multiple of $D_x$ ) [MHz]	318	Aligned to $D_x$ raster
Number of OFDM carriers	13 944+1	Within the 32 MHz C2 channel Number without notch
Start frequency [MHz]	302,44 MHz	318 MHz - (13 944/2) x (1/448 us)
Number of carriers <b>Data Slice</b> 1 to 4 <b>Data Slice</b> 5,6	2208 2544	Width 1 to 4 = 2 208/448 us = 4,93 MHz Width 5 = 2 554/448 us = 5,70 MHz
<b>Data Slice</b> tuning positions [MHz] <b>Data Slice</b> 1 to 4          <b>Data Slice</b> 5 to 6	306,2 MHz 309,8 MHz 314,8 MHz 319,7 MHz          325,0 MHz 329,7 MHz	The tuning positions of <b>Data Slices</b> 1 and 6 shall have a distance of at least 1704 OFDM carriers from the start or the end of the <b>C2 system</b> , the tuning positions 2 to 5 are in the middle of <b>Data Slice</b> 2 to 5
Narrowband notch Start Frequency	327,83 MHz	Notch starts at high frequency end of <b>Data Slice</b> 5
Width (carriers)	23	

Table 33 shows all **L1-part2** signalling parameters of the narrowband notch example.

**Table 33: L1 part 2 signalling for the narrowband notch example**

Parameter	Value (transmission format)	Comment
NETWORK_ID	0000 0000 0000 0000	Network identifier
C2_SYSTEM_ID	0000 0000 0000 0000	<b>C2 system</b> identifier
START_FRQUENCY	000000100001000101000100	135 492: value of table 31 multiplied with 448 us (1/carrier spacing)
C2_BANDWIDTH	0000001001000100	13 944/24 = 581 due to calculated number of carriers. Total 581 x 24 + 1 = 13 945 carriers due to insertion of the highest carrier frequency edge pilot
GUARD_INTERVAL	00	$GI = 1/128$
C2_FRAME_LENGTH	0111000000	448 <b>data symbols</b> in one <b>C2 frame</b>
L1_PART 2_CHANGE_COUNTER	0000 0000	No changes between <b>C2 frames</b> , because one constant data rate multiplex (MPEG2-TS) in each <b>Data Slice</b> . Bit rate of those MPEG2-TS remain constant over time
NUM_DSLICE	0000 0110	6 <b>Data Slices</b> within the C2 channel
NUM_NOTCH	0001	1 narrowband notch within the <b>C2 system</b>
DATA_SLICE_ID ( <b>Data Slice</b> 1 to 6)	0000 0000.....0000 0101	identifier for <b>Data Slice</b> ("0"....."5")

Parameter	Value (transmission format)	Comment
DSLICE_TUNE_POS (Data Slice 1 to 6)	0000001000111 0000010001010 0000011100110 0000101000010 0000110100101 0000111111110	71 (3 408/2/24); tune pos = 1 704 <sup>th</sup> OFDM carrier 138 (3 x 2 208/2/24); tune pos = 3 312 <sup>th</sup> OFDM carrier 230 (5 x 2 208/2/24); tune pos = 5 520 <sup>th</sup> OFDM carrier 322 (7 x 2 208/2/24); tune pos = 7 728 <sup>th</sup> OFDM carrier 421 (4 x 2 208 + 2 544/2)/24; tune pos = 10 104 <sup>th</sup> OFDM carrier 510 (13 944 - 1 704)/24; tune pos (13 944 - 1 704) <sup>th</sup> OFDM carrier
DSLICE_OFFSET_LEFT/RIGHT		
Data Slice 1	10111001 / 00010101	-71/+21 for the first Data Slice (min 1 704 OFDM carrier distance from the start of the C2 system )
Data Slice 2 to 4	11010010 / 00101110	-46/+46 ( $\pm 2$ 208/2/24); $\pm 1$ 104 carriers left and right from tune pos (Data Slice 1-4)
Data Slice 5	11001100 / 00110101	-53/+53 ( $\pm 2$ 544/2/24); $\pm 1$ 272 carriers left and right from tune pos (Data Slice 5)
Data Slice 6	11011101 / 01000111	-35/+71 for the last Data Slice (min 1 704 OFDM carriers distance from the end of the C2 system + (13 920 - 13 632) carriers)
DSLICE_TI_DEPTH(Data Slice 1 to 6)	01	4 OFDM symbols
DSLICE_TYPE (Data Slice 1 to 6)	0	Type 1, Single PLP (constant bitrate MPEG2-TS) per Data Slice
DSLICE_CONST_CONF (Data Slice 1 to 6)	1	Data Slice does not change
DSLICE_LEFT_NOTCH (Data Slice 1 to 6)	1	Left notch of Data Slice 6
DSLICE_NUM_PLP	00000001	1 PLP per Data Slice
PLP_ID (Data Slice 1-6)	0	PLP identifier ( "0" )
PLP_BUNDLED (Data Slice 1-6)	0	Not bundled with PLP's of other Data Slices
PLP_TYPE (Data Slice 1-6)	10	Normal data PLP
PLP_PAYLOAD_TYPE (Data Slice 1 to 6)	00011	MPEG2-TS
PLP_START (Data Slice 1 to 6)	00000000000010	First complete XFECFrame starts from 2 <sup>nd</sup> data cell of this Data Slice
PLP_FEC_TYPE	1	This PLP uses 64 K LDPC
PLP_MOD	100	This PLP uses 1 024-QAM modulation
PLP_COD	101	This PLP uses code rate 9/10
PSI/SI_REPROCESSING	1	Cable NIT, inserted in every MPEG2-TS, therefore PSI/SI reprocessing
Notch_Start	0000111011010	At the low frequency end of Data Slice 6 421 (tune pos) + 53 (offset) = 474
Notch_Width	1	Lowest possible narrowband notch: 23 subcarriers
RESERVED_3	00000000	Reserved for future use
RESERVED_TONE	0	There is no reserved tone in this C2 frame
EARLY WARNING SYSTEM (EWS)	0	There is no emergency event
C2 VERSION	0000	The C2 system ID is 0x0
RESERVED_4	000000000000	Reserved for future use

#### 12.1.1.4 Example for a configuration with a broadband notch within a DVB-C2 signal

This example gives the signalling parameters of DVB-C2 signal with two broadband **notches**, which takes into account the protection of two wireless BOS services located in the frequency ranges of 443,59375 - 444,96875 MHz and 448,59375 - 449,96875 MHz (example from Germany according to figure 87 in clause 12.1.6). The **preamble symbols** located in the range of the narrowband **notches** are switched off. In such a configuration the **notches** do not weaken the error protection of the preamble as adjacent to the broadband notch in the regular case a complete preamble is located at least one side (lower or higher frequency part) of the broadband notch. Apart from this, the retransmission example of clause 12.1.1 is reused.

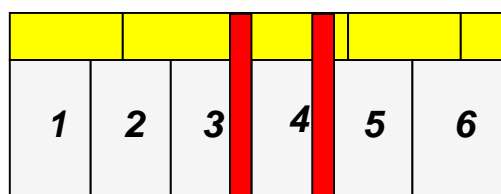
In table 34 details of **Data Slice** configuration are shown. A 32 MHz C2 channel is inserted between 430 MHz and 462 MHz and centred to 446 MHz to achieve equal frequency distance to lower and higher adjacent channels. Two broadband **notches** are inserted at the high frequency end of **Data Slice** 3 at 443,59 MHz, with a notch width of 672 carriers and at the high frequency end of **Data Slice** 4 with a notch width of 720 carriers. The minimum C2 channel bandwidth for this notch configuration is 24 MHz, if existing 8 MHz channel raster is used (the two broadband **notches** are in two different channels) and 16 MHz, if the C2 channel can be positioned out of the regular channel raster.

**Table 34: Configuration of start frequency and Data Slices of the 32 MHz C2 channel including a broadband notch**

Parameter	Value	Comment
Centre frequency [MHz]	446	
OFDM carrier spacing	1/448 us	Due to 8 MHz channel raster
$D_x$	24 OFDM carriers	Due to GI=1/128
Centre frequency (multiple of $D_x$ ) [MHz]	445,98	Aligned to $D_x$ raster
Number of OFDM carriers	13,920	Within the 32 MHz C2 channel number without <b>notches</b>
Start frequency [MHz]	430,45 MHz	445,98 MHz - (13 920/2) x (1/448 us)
Number of carriers <b>Data Slice</b> 1 and 2	2 208	Width 1 and 2 = 2 208/448 us = 4,93 MHz
<b>Data Slice</b> 3	1 440	Width 3 = 1 440/448 us = 3,21 MHz
<b>Data Slice</b> 4	1 584	Width 4 = 1 584/448 us = 3,54 MHz
<b>Data Slice</b> 5 and 6	2 544	Width 5 and 6 = 2 544/448 us = 5,68 MHz
<b>Data Slice</b> tuning positions [MHz] <b>Data Slice</b> 1 to 4	434,25 MHz 437,8 MHz 439,77 MHz 446,8 MHz	The tuning positions of <b>Data Slice</b> 1 and <b>Data Slice</b> 6 shall have a minimum distance of 1704 OFDM carriers from the start or the end of the <b>C2 system</b> . Furthermore the tuning position of <b>Data Slice</b> 3 shall have a minimum distance of 1704 OFDM carriers from the lower edge of broadband notch 1 and the tuning position of <b>Data Slice</b> 5 shall have a minimum distance of 1704 OFDM carriers from the upper edge of broadband notch 2.
<b>Data Slice</b> 5 and 6	453,0 MHz 457,7 MHz	
Broadband notch 1	443,52 MHz	Notch starts at upper frequency end of <b>Data Slice</b> 3
Broadband notch 2	448,55 MHz	Notch starts at upper frequency end of <b>Data Slice</b> 4

According to table 34 the **notches** are positioned in the 32 MHz C2 channel example as shown in figure 83.





**DVB-C2-channel (B=32MHz)**

NOTE: The 32 MHz DVB-C2 channel is located between 430 and 462 MHz; the both notch positions start at 443,52 MHz (width=1,45 MHz) and 448,55 MHz (width=1,61 MHz) according to table 33.

**Figure 83: Example for Broadband Notches in the area of BOS services (Germany)**

The bandwidth of the two broadband **notches** (1,45 MHz and 1,67 MHz) slightly differs from the nominal bandwidth of the two BOS services to be protected. The reason is that notch bandwidth shall be a multiple of the pilot raster (24 carriers for GI=1/128). Furthermore notch 2 is slightly wider than notch 1, because the very small frequency gap to the upper adjacent **Data Slice** of notch 2 is included into notch 2. The **Data Slices** are chosen according to the example in table 34. But **Data Slices** 3 and 4 cannot be used for retransmission of the satellite transponders given in table 33 due to notch location (possible data rate for **Data Slice** 3 and 4 reduced). Therefore for these two **Data Slices** other content sources shall be used. Additionally **Data Slice** 4 only can be used for data transmission if L1 signalling remains static (no changes will happen during transmission), because a receiver, which has been already tuned to **Data Slice** 4, cannot anymore detect a complete L1 signal.

Table 35 shows all **L1-part2** signalling parameters of the broadband notch example.

**Table 35: L1-part2 signalling for the DVB-C2 broadband notch example**

Parameter	Value (transmission format)	Comment
NETWORK_ID	0000 0000 0000 0000	Network identifier
C2_SYSTEM_ID	0000 0000 0000 0000	<b>C2 system</b> identifier
START_FRQUENCY	000000100001100001000000	137 280: value of table 33 multiplied with 448 us (1/carrier spacing)
C2_BANDWIDTH	0000001001000100	13 920/24 = 580 due to calculated number of carriers. Total 580 x 24 + 1 = 13 921 carriers due to insertion of the highest carrier frequency edge pilot  Total 580 x 24 + 1 = 13 921 carriers due to insertion of the highest carrier frequency edge pilot
GUARD_INTERVAL	00	GI=1/128
C2_FRAME_LENGTH	0111000000	448 <b>data symbols</b> in one <b>C2 frame</b>
L1_PART 2_CHANGE_COUNTER	0000 0000	No changes between <b>C2 frames</b> , because one constant data rate multiplex (MPEG2-TS) in each <b>Data Slice</b> . Bit rate of those MPEG2-TS remain constant over time
NUM_DSLICE	0000 0110	6 <b>Data Slices</b> within the C2 channel
NUM_NOTCH	0010	2 broadband <b>notches</b> within the C2_system
DATA_SLICE_ID (Data Slice 1 to 6)	0000 0000.....0000 0101	identifier for <b>Data Slice</b> ("0"...."5")
DSLICE_TUNE_POS (Data Slice 1 to 6)	0000001000111 0000010001010 0000011010110 0000100110001 0000110100101 0000111111101	71 (3 408/2/24); tune pos = 1 704 <sup>th</sup> OFDM carrier 138 (3 x 2 208/2/24); tune pos = 3 312 <sup>th</sup> OFDM carrier 2 142 x (2 208+(1 440/2))/24; tune pos = 5 136 <sup>th</sup> OFDM carrier 305 (2 208+1 440+672+(1 584/2))/24; tune pos = 7 320 <sup>th</sup> OFDM carrier 421 (4 x 2 208 + 2 544/2)/24; tune pos = 10 104 <sup>th</sup> OFDM carrier 509 (13 920 - 1 704)/24; tune pos (13 920 - 1 704) <sup>th</sup> OFDM carrier

Parameter	Value (transmission format)	Comment
DSLICE_OFFSET_LEFT/RIGHT <b>Data Slice 1</b>	10111001 / 00010101	-71/+21 for the first <b>Data Slice</b> (minimum 1704 OFDM carrier distance from the start of the <b>C2 system</b> )
<b>Data Slice 2</b>	11010010 / 00101110	-46/+46 ( $\pm 208/2/24$ ); $\pm 1$ 104 carriers left and right from tune pos
<b>Data Slice 3</b>	00001010/001000110	-30/+30 ( $\pm 1$ 440/2/24); $\pm 720$ carriers left and right from tune pos ( <b>Data Slice 1-4</b> )
<b>Data Slice 4</b>	10111110 / 01000010	-33/+33 ( $\pm 1$ 584/2/24); $\pm 792$ carriers left and right from tune pos ( <b>Data Slice 1-4</b> )
<b>Data Slice 5</b>	10111001 / 00100011	-71/+35 (minimum 1 704 carriers distance from upper edge of broadband notch 2)
<b>Data Slice 6</b>	11011101 / 01000111	-35/+71 for the last <b>Data Slice</b> ; minimum 1 704 OFDM carriers distance from the end of the <b>C2 system</b>
DSLICE_TI_DEPTH( <b>Data Slice 1 to 6</b> )	01	4 OFDM symbols
DSLICE_TYPE ( <b>Data Slice 1 to 6</b> )	0	Type 1, Single <b>PLP</b> (constant bitrate MPEG2-TS) per <b>Data Slice</b>
DSLICE_CONST_CONF ( <b>Data Slice 1 to 6</b> )	1	<b>Data Slice</b> does not change
DSLICE_LEFT_NOTCH ( <b>Data Slice 1 to 6</b> )	1	Left notch of <b>Data Slice 4</b> and <b>Data Slice 5</b>
DSLICE_NUM_PLP	000000001	1 <b>PLP</b> per <b>Data Slice</b>
PLP_ID ( <b>Data Slice 1 to 6</b> )	0	<b>PLP</b> identifier ("0")
PLP_BUNDLED ( <b>Data Slice 1 to 6</b> )	0	Not bundled with <b>PLPs</b> of other <b>Data Slices</b>
PLP_TYPE ( <b>Data Slice 1 to 6</b> )	10	Normal data <b>PLP</b>
PLP_PAYLOAD_TYPE ( <b>Data Slice 1 to 6</b> )	00011	MPEG2-TS
PLP_START ( <b>Data Slice 1 to 6</b> )	00000000000010	First complete <b>XFECFrame</b> starts from 2 <sup>nd</sup> <b>data cell</b> of this <b>Data Slice</b>
PLP_FEC_TYPE	1	This <b>PLP</b> uses 64 K LDPC
PLP_MOD	100	This <b>PLP</b> uses 1 024-QAM modulation
PLP_COD	101	This <b>PLP</b> uses code rate 9/10
PSI/SI_REPROCESSING	1	Cable NIT, inserted in every MPEG2-TS, therefore PSI/SI reprocessing
Notch1_Start	0000011110100	At the high frequency end of <b>Data Slice 3</b> (starting position 443,42 MHz)
Notch 2_Start	0000101010010	At the high frequency end of <b>Data Slice 4</b> (starting position 448,55 MHz)
Notch1_Width	00011100	672 subcarriers ( $(672/24) = 28$ )
Notch2_Width	00011110	720 subcarriers ( $(720/24) = 30$ )
RESERVED_3	00000000	Reserved for future use
RESERVED_TONE	0	There is no reserved tone in this <b>C2 frame</b>
EARLY WARNING SYSTEM (EWS)	0	There is no emergency event
C2 VERSION	0000	The <b>C2 system</b> ID is 0x0
RESERVED_4	000000000000	Reserved for future use

## 12.1.2 Further application for Common **PLP**'s and their efficient transmission

The propagation delay of data carried in common **PLPs** may differ from data carried in data **PLPs**, as there is no time synchronization mechanism available in the receiver to re-establish the original sequence of e.g. TS packet. However, there are many applications for carrying data in common **PLPs**, such as:

- EPG data, e.g. encoded in the DVB SI EIT table format.

- Entitlement Management Messages (EMMs) of the Conditional Access system.
- DSMCC data carousels, where no strict time correlation to other components of a service is required.

### 12.1.3 Video On Demand and other applications of personalized services

VOD services are 'unicast-type' of services, as a customer is requesting an individual communication link between the VOD server and his VOD enabled CPE. Therefore VOD services could be delivered via cable networks by means of both transmission systems DVB-C and DVB-C2, either using them separately or in parallel. The receiver of an individual customer could send the information whether it supports DVB-C and/or DVB-C2 reception capabilities together with the request for viewing of a VOD-event. The VOD backend-system will ensure that the event is transmitted with the proper format to the individual customer. The penalty a cable operator would have to pay for implementing the possibilities of a mixed DVB-C/DVB-C2 VOD scenario is the deployment of headroom for peak usage for both systems since traffic peaks will happen in both the DVB-C and the DVB-C2 system.

### 12.1.4 Utilization for interactive services (IPTV, Internet)

#### 12.1.4.0 Introduction

The following system explanations are based on the simplified block diagram depicted in figure 80 which shows two basic downstream delivery scenarios for interactive services implemented in the last mile of a cable network. The first scenario describes an IP delivery via the DVB-C2 downstream pipe, which is operated in parallel of a DOCSIS<sup>®</sup> system, whereas the second scenario shows how DVB-C2 is used as an integral part of DOCSIS<sup>®</sup>.

##### 1) Hybrid scenario

In this scenario programs are delivered in MPEG2 **Transport Stream** format via separately operated DVB-C2 cable channels. Examples for services mentioned in this context are TV, VOD etc. DVB-C2 is configured to work in usual broadcast mode at physical layer. Service request by the user is performed via upstream by a DOCSIS<sup>®</sup> cable modem.

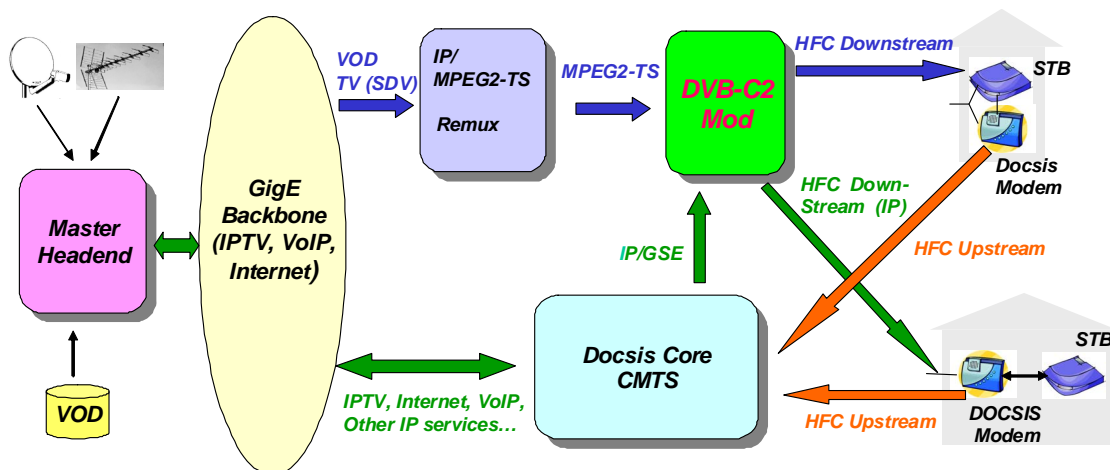
- A separate transmission of video content in parallel to the IP based system makes available more capacity for services which need to be transmitted by means of an IP stream (Internet, VoIP, High Speed business data). For example a universal Edge QAM device can schedule a higher number of cable channels for IP during the day (when business applications need more capacity) which then will be reduced during the evening, when more channels are required for TV and VoD.

Furthermore the hybrid scenario reduces equipment requirement on user side, if only TV and VoD is required (only up to DOCSIS<sup>®</sup> 2.0 modem needed for upstream).

##### 2) IP over HFC scenario

In this scenario programs are delivered as IP streams via DVB-C2 cable channels configured to work as an integral part of the downstream delivery media of the DOCSIS<sup>®</sup> system. The Encapsulation of IP packets is done by Generic Stream Encapsulation (GSE). GSE is an IP encapsulation method especially optimized for the Baseband Frame format as used in DVB second generation transmission standards as DVB-S2, DVB-T2, and DVB-C2. IP over HFC is useful for:

- a mixed IP service delivery including various services such as IPTV, VoIP, High speed business data, Internet, etc.;
- low latency applications like Network PVR based VOD, gambling, etc.



**Figure 84: Scenarios for interactive service transmission with DVB-C2 in a stand-alone configuration**

NOTE: Figure 84 shows the scenarios for interactive service transmission with DVB-C2 in a stand-alone configuration (in the figure referred to as HFC Downstream) on the one hand and an integral part of DOCSIS® (in the figure referred to as HFC Downstream (IP)).

#### 12.1.4.1 Example for IP transmission

Below a more detailed IP transmission example is shown.

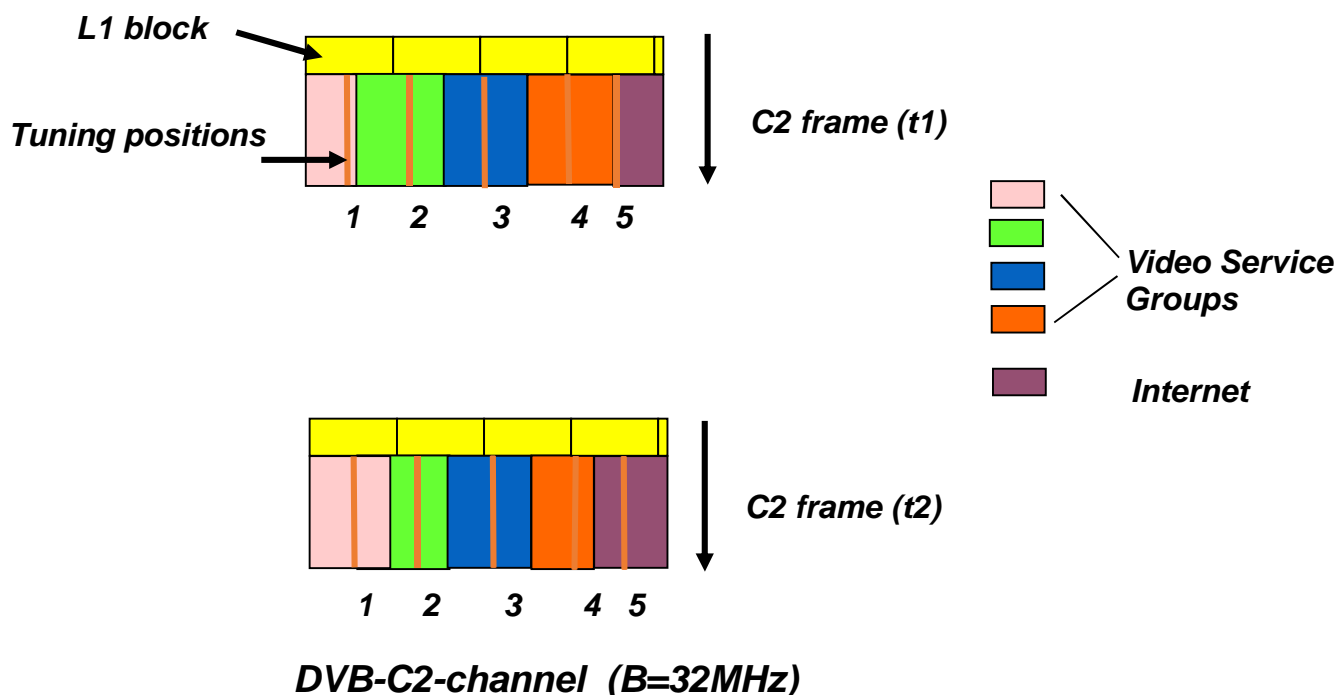
In a 32 MHz channel the maximum transmission capacity for 1 024-QAM and code rate 8/9 is:

$$R = \frac{14\,160}{14\,336} * \frac{448\,us}{10 * (128/129) * (8/9) * (14\,232/14\,400) * (95/96) * 0.99 * (448/449)} = 269.3\,Mbit/s \quad (28)$$

Where the maximum number of 14 160 active OFDM carriers (B=31,607 MHz) within a 32 MHz channel (14 336 nominal OFDM carriers) is used. This data rate value still includes the overhead to be taken into account due to GSE and IP transmission (values typical about 10 % for high data rate videos).

The 32 MHz channel is divided into 4 video **Data Slices** and 1 **Data Slice** for internet and other general purpose data.

In each video **Data Slice** 6 HD Video streams (MPEG4 advanced video coding, VBR, 10 Mbit/s max) and 2 SD Video streams (MPEG4 advanced video coding, VBR, 2 Mbit/s max) are transmitted. Due to VBR video streams the transmission rate of the video **Data Slices** are varying, the difference between the maximum possible data rate (269,3 Mbit/s) and the actual data rate of the 4 video **Data Slices** is used for internet and general purpose data in a fifth **Data Slice**. The tuning positions of the five **Data Slices** are simply equidistant distributed within the C2 channel, starting from the 1 704<sup>th</sup> carrier. The 5<sup>th</sup> tuning position deviates from equidistant tuning position raster, because a distance of 1 704 OFDM carriers from the end of the **C2 system** is required. Figure 85 shows how the width and the offset values of the **Data Slices** vary between different **C2 frames** (t1 and t2) due to the variable bit rate of the video streams.



**Figure 85: C2 channel configuration for variable bit rate input video and data streams during different C2 frame periods t1 and t2, block diagram of the headend**

Figure 85 shows the configuration for variable bit rate input video and data streams and table 36 shows the characteristics (**Data Slices**, start frequency, GI etc.) of the C2 channel shown in figure 85.

**Table 36: Configuration of start frequency and Data Slices of the 8 MHz C2 channel for IPTV and data transmission**

Parameter	Value		Comment
OFDM carrier spacing	1/448 us		Due to 8 MHz channel raster
$D_x$	24 OFDM carriers		Due to GI = 1/128
Number of OFDM carriers	14 160		Within the 32 MHz C2 channel
Start frequency [MHz]	302,20 MHz		Multiple of OFDM carrier spacing and $D_x$ raster
Number of carriers, bandwidth	Carr. Number	Bandwidth [MHz]	Data rates [Mbit/s]
<b>Data Slice 1 (8 PLPs)</b>	2 424(t1), 3 384(t2)	5,41(t1); 7,55(t2)	Video group 1: 46(t1); 64(t2)
<b>Data Slice 2 (8 PLPs)</b>	3 384(t1), 2 904(t2)	7,55(t1); 6,48(t2)	Video group 2: 64(t1); 55(t2)
<b>Data Slice 3 (8 PLPs)</b>	2 904(t1), 2 904(t2)	6,48(t1); 6,48(t2)	Video group 3: 55(t1); 55(t2)
<b>Data Slice 4 (8 PLPs)</b>	3 384(t1), 2 424(t2)	7,55(t1); 5,41(t2)	Video group 4: 64(t1); 46(t2)
			8 MPEG4 advanced video coding Videos in each Video Group (6 x HD 10 Mbit/s and 2 x SD 2 Mbit/s)
<b>Data Slice 5 (2 PLPs)</b>	2 064(t1); 2 544(t2)	4,61(t1); 5,68(t2)	Data group: 40,3(t1); 49,3(t2) One high speed Internet Service
<b>Data Slice tuning positions [MHz]</b>			The tuning position of <b>Data Slice</b> 1 and 5 have a distance of minimum 1 704 OFDM carriers from end of the <b>C2 system</b> . The tuning positions 2-4 have equal OFDM carrier distance to each other
<b>Data Slice 1 to 4</b>	306,0 MHz		
	312,3 MHz		
	318,6 MHz		
	325,0 MHz		
	330,0 MHz		

The calculation of OFDM carrier numbers for the different **Data Slices** is done according to equation (26), but with LDPC code rate 8/9 and the corresponding BCH code rate of 14 232/14 400 (short LDPC frames for IP transmission, length 14 400 bit).

Table 37 shows all **L1-part2** signalling parameters.

**Table 37: L1 signalling for the IP transmission**

Parameter	Value (transmission format)	Comment
NETWORK_ID	0000 0000 0000 0000	Network identifier
C2_SYSTEM_ID	0000 0000 0000 0000	<b>C2 system</b> identifier
START_FRQUENCY	000000100001000011011000	135384: value of Tab 35 multiplied with 448 us (1/carrier spacing), which shall also be a multiple integer value of $D_x$ (24)
C2_BANDWIDTH	0000001001001110	14 160/24 = 590 due to calculated number of carriers Total 590 x 24 + 1 = 14 161 carriers due to insertion of the highest carrier frequency edge pilot
GUARD_INTERVAL	00	GI = 1/128
C2_FRAME_LENGTH	0111000000	448 <b>data symbols</b> in one <b>C2 frame</b>
L1_PART 2_CHANGE_COUNTER	0000 0001	Changes between <b>C2 frames</b> because of varying data rate of <b>Data Slices</b>
NUM_DSlice	0000 0101	5 <b>Data Slices</b> within the C2 channel
NUM_NOTCH	0000	No notch within the C2 system
DATA_SLICE_ID ( <b>Data Slice</b> 1 to 6)	0000 0000.....0000 0100	identifier for <b>Data Slice</b> ("0"...."5")
DSLICE_TUNE_POS ( <b>Data Slice</b> 1 to 5)	0000001000111, 0000010111101, 0000100110011, 0000110101001, 0001000000111	71; 189; 307; 425; 519 The first and the last tuning positions are located minimum 1 704 carriers from start and end of the <b>C2 system</b> , the other three tuning positions have an equal carrier distance to each other
DSLICE_OFFSET_LEFT/RIGHT <b>Data Slice</b> 1	10111001/00011110(t1); 10111001/01000110(t2)	-71/30 (t1); -71/70(t2)
<b>Data Slice</b> 2	01011000/00110101(t1); 11010000/01001001(t2)	-88/53(t1); -48/73(t2)
<b>Data Slice</b> 3	10111111/00111000(t1); 11010011/01001100(t2)	-65/56(t1); -45/76(t2)
<b>Data Slice</b> 4	11000010/01001111(t1); 11010110/00111011(t2)	-62/79(t1); -42/59(t2)
<b>Data Slice</b> 5	11110001/01000111 (t1); 11011101/ 01000111 (t2)	-15/71(t1); -35/71(t2) Calculation as described in the other examples above
DSLICE_TI_DEPTH( <b>Data Slice</b> 1 to 5)	01	4 OFDM symbols
DSLICE_TYPE ( <b>Data Slice</b> 1 to 5)	1	Type 2
FEC_HEADER_TYPE	1	High Efficiency Mode
DSLICE_CONST_CONF ( <b>Data Slice</b> 1 to 5)	0	<b>Data Slice</b> changes between <b>C2 frames</b>
DSLICE_LEFT_NOTCH ( <b>Data Slice</b> 1 to 4)	0	No notch
DSLICE_NUM_PLP	0000 1000 0000 0010	8 <b>PLPs</b> per <b>Data Slice</b> in <b>Data Slice</b> 1 to 4 2 <b>PLPs</b> per <b>Data Slice</b> in <b>Data Slice</b> 5
<b>PLP_ID</b> ( <b>Data Slice</b> 1 to 4), 8 <b>PLPs</b> <b>PLP_ID</b> ( <b>Data Slice</b> 5), 2 <b>PLPs</b>	0000 0000....0000 0111 0000 00000/0000 0001	<b>PLP</b> identifier ("0"/"1"/"2"/"3"/"4"/"5"/"6"/"7") <b>PLP</b> identifier ("0"/"1")
<b>PLP_BUNDLED</b> ( <b>Data Slice</b> 1 to 4)	0	Not bundled with <b>PLPs</b> of other <b>Data Slices</b>
<b>PLP_TYPE</b> ( <b>Data Slice</b> 1 to 6)	10	Normal data <b>PLP</b>
<b>PLP_PAYLOAD_TYPE</b> ( <b>Data Slice</b> 1 to 5)	00010	GSE
<b>PLP_GROUP_ID</b>	00	No association with common <b>PLP</b>
PSI/SI_REPROCESSING	0	IP data, Data Service, signalling on IP level (DVB-IPI)

## 12.1.4.2 Recommendations for arrangement of PLP's and Data Slices

### 12.1.4.2.0 Introduction

The following description gives a guidance how **PLP's** and **Data Slices** may be arranged dependent on the conditions within a certain cable network.

#### 12.1.4.2.1 Robust mode

In this mode the **Data Slice** width should be as high as possible (upper limit 7,61 MHz) to support optimal frequency deinterleaver performance in the receiver. The reason for that is that frequency interleaving is only performed within a **Data Slice** but not within the whole C2 channel. In case of rather discrete disturbances or so called "cluster beats", as they occur for example within cable networks with mixed occupations of DVB and PAL signals, the "Robust mode" should be the preferred choice. This obviously depends also from the level and therefore relevance of those interferers. Within a **Data Slice** of maximum 7,61 MHz some buffer bandwidth shall be foreseen because of VBR rate of incoming data streams/**PLP's**. The amount of buffer overhead depends on the number of services transmitted within a **Data Slice** (see also figures 86 and 87). Therefore for HD services (higher VBR rate) a higher buffer bandwidth for a **Data Slice** shall be foreseen as for SD services (lower VBR rate). But the necessary buffer overhead usually is minimum 10%. Because this overhead shall be foreseen in every **Data Slice** which carries VBR payload within the same C2 channel, this mode does not achieve optimum transmission capacity ("statmux gain") for the complete C2 channel. Figure 88 shows the "robust mode" configuration.

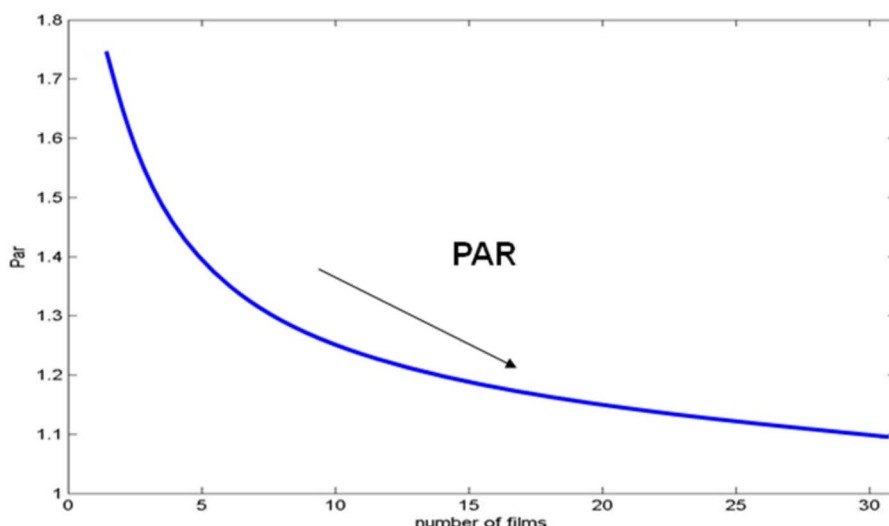


Figure 86: Dependency of Peak to Average data rate ratio ("Par") on the number of transmitted video streams ("films"); effect of "statmux gain"

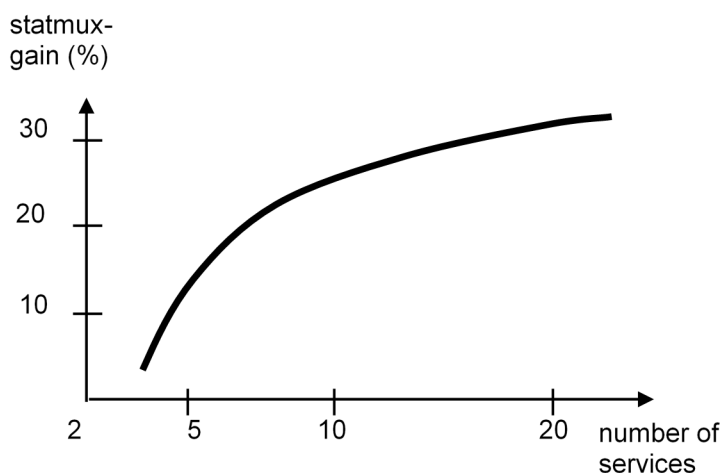
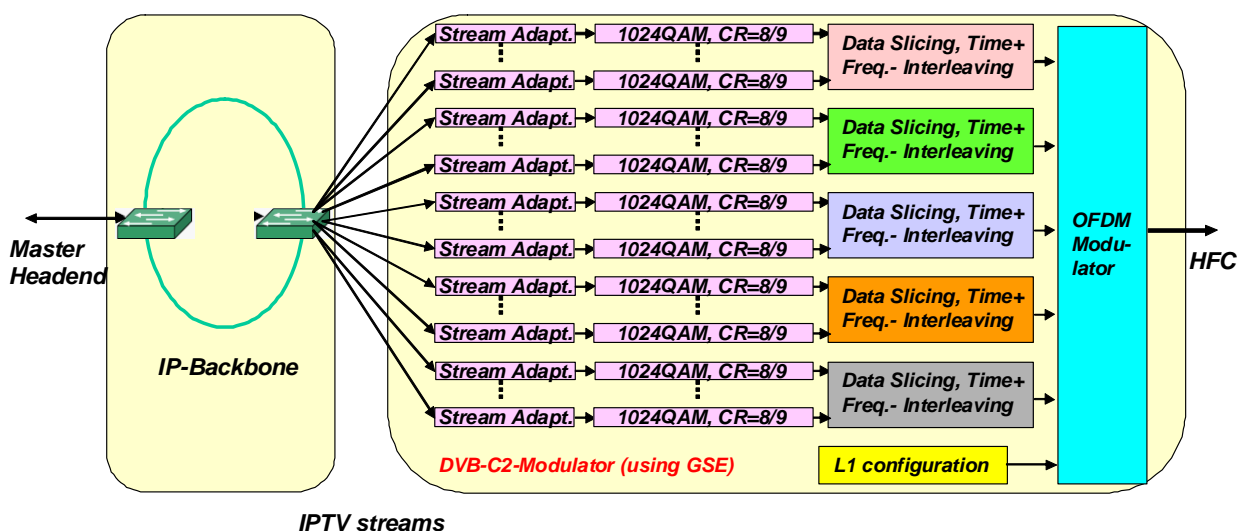


Figure 87: Gain of statistical multiplexing versus number of services integrated in the statistical multiplex, schematic representation



NOTE: This configuration gives best frequency interleaver performance for CSO/CTB (for example mixed occupation with Pal).

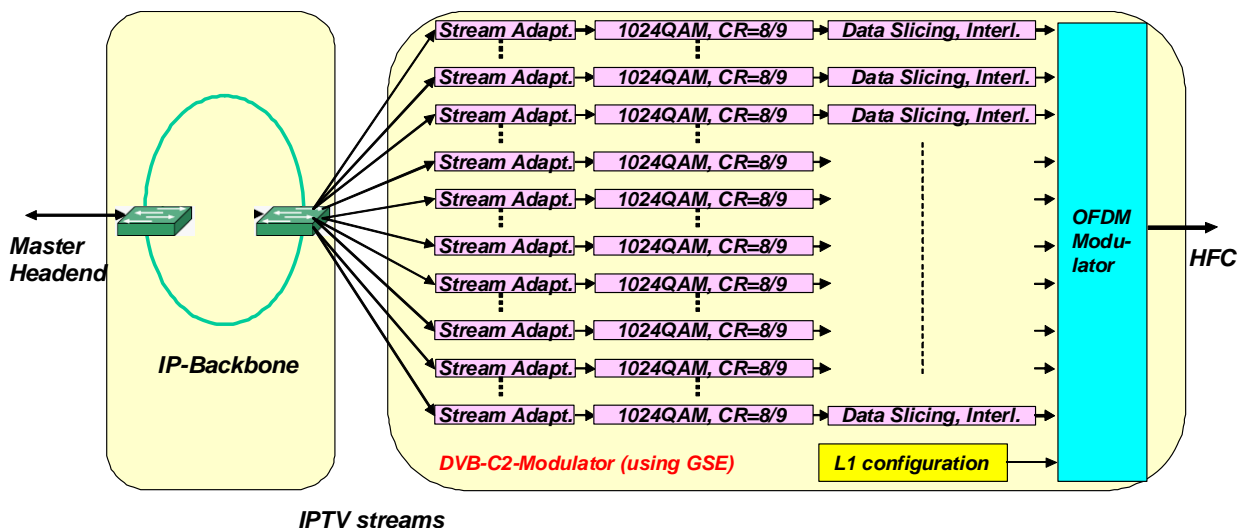
**Figure 88: Configuration of PLP's and Data Slices for robust configuration (Data Slice width as close as possible to the maximum allowed bandwidth of 7,61 MHz)**

12.1.4.2.2 High Efficiency mode

Where level of discrete interferences within a **Data Slice** can be neglected, this mode should be chosen. In this mode for every input stream /PLP a single **Data Slice** is foreseen (see figure 89).

Buffer overhead shall be foreseen only once for the complete C2 channel. This buffer overhead is always in the range of the necessary minimum of 10 % if the C2 bandwidth is sufficient large.

Stuffing to a constant output transmission rate can be performed by stuffing **Data Slices** as described in clause 9.5 of [i.2] and clause 7.6 of the present document.



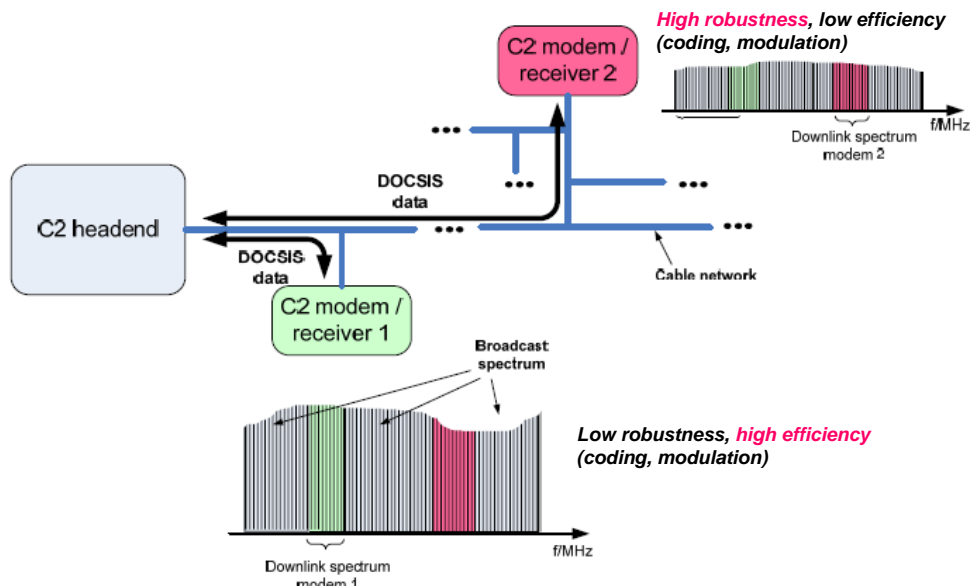
NOTE: One PLP per **Data Slice** results in optimum statmux gain, because peak to average data rate ratio is lower for the complete DVB-C2 channel than for a single **Data Slice**.

**Figure 89: High Efficiency mode for single service transmission**



### 12.1.4.3 Adaptive Coding and Modulation (ACM)

Cable networks with bi-directional communication capabilities allow the implementation of variable coding and modulation for the purpose of adapting the parameters for these elements in accordance with the channel conditions such as the CNR value occurring at a dedicated user outlet. This feature is referred to as Adaptive Coding and Modulation. The CNR value and other channel impairments measurable are reported to the headend via the return path.



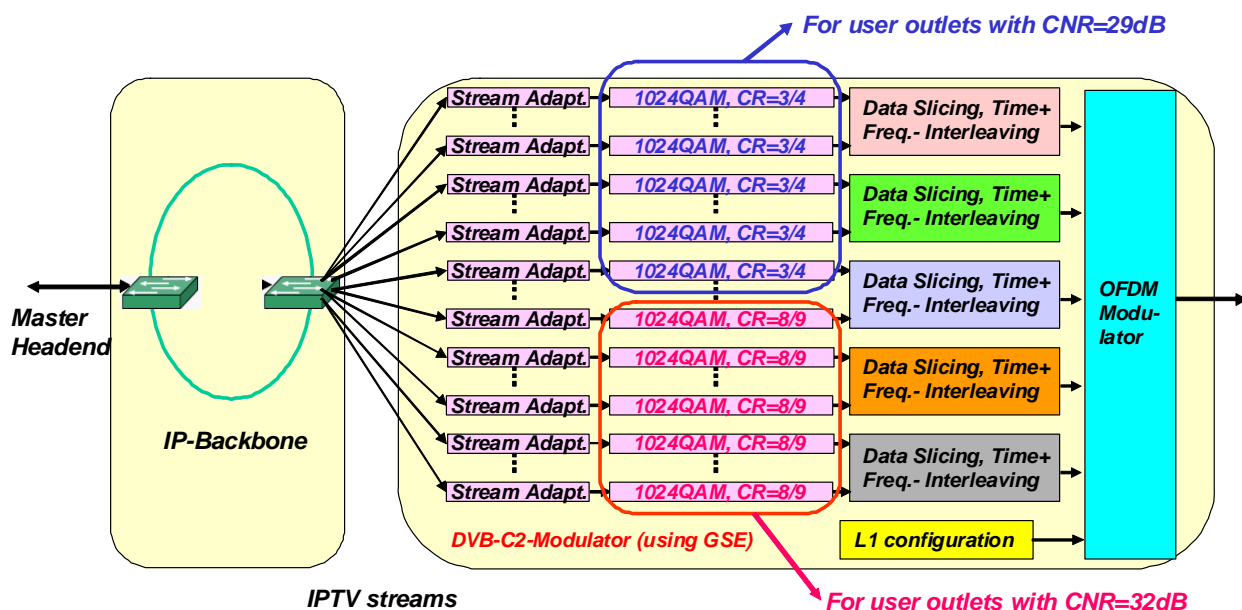
**Figure 90: Principle of Adaptive Coding and Modulation (ACM)**

Figure 90 visualizes such a scenario. User outlets with high CNR values receive the signal configured for lower robustness and higher efficiency. In contrary the signal is transmitted to user outlets with low CNRs by means of a higher robustness and consequently a lower efficiency.

The application of ACM seems to provide particular benefits in networks with varying channel conditions per service area. DVB-C2 could optimize the spectral efficiency for each individual unicast stream taking account of, for instance, the signal-to-noise ratio measured at the corresponding user outlet and adapting the channel coding and modulation appropriately. In this example it is obvious that the implementation of Variable Coding and Modulation (VCM) provides great benefits compared to an application of a schema using Constant Channel Coding and Modulation (CCM).

A concrete example for a unicast IPTV application (e.g. VOD) is given in table 33. It is assumed that a CNR lower than 29 dB is measured at a certain percentage of user outlets of a service area. This value includes a 3 dB implementation margin leading to the selection of a 1 024-QAM combined with a FEC code rate of 3/4. The remaining outlets support a higher SNR value of 32 dB including a 2,5 dB margin, allowing an application of 1 024-QAM and FEC code rate of 8/9.

Figure 91 shows the block diagram of the DVB-C2 downstream part for ACM.



**Figure 91: Block Diagram of the downstream configuration for Adaptive Coding and Modulation (ACM) using Variable Coding and Modulation (VCM) in the DVB-C2 Modulator**

For the IP transmission example given in clause 12.1.5.1 the following results are relevant:

- 1) If all services within the C2 channel are supplied to all users of the network, CCM (Constant Coding and Modulation) shall be applied. In this case generally the robust modulation and coding scheme (1 024-QAM, CR = 3/4) shall be chosen to assure QoS for the complete network. Therefore the possible data rate is reduced according to the equation below:

$$R = \frac{14160}{14336} * \frac{448us}{10 * (128/129) * (3/4) * (11712/11880) * (95/96) * 0.99} = 226.65 \text{ Mbit / s} \quad (29)$$

- 2) If not all services within the C2 channel are required by users with the lower CNR values at their outlet, VCM (Variable Coding and Modulation) can be applied. If for example only 50 % of the services within the C2 channel are required by users with lower outlet CNR, the resulting overall data rate is calculated as follows:

$$R_{robust} = 0,5 * \frac{14160}{14336} * \frac{448us}{10 * (128/129) * (3/4) * (11712/11880) * (95/96) * 0.99} = 113.32 \text{ Mbit / s} \quad (30)$$

for the services with robust coding and modulation, and

$$R_{eff} = 0,5 * \frac{14160}{14336} * \frac{448us}{10 * (128/129) * (8/9) * (14232/14400) * (95/96) * 0.99} = 134.65 \text{ Mbit / s} \quad (31)$$

for the services with efficient coding and modulation.

The total bit rate therefore is  $R_{tot} = 113,32 \text{ Mbit / s} + 134,65 \text{ Mbit / s} = 248 \text{ Mbit / s}$

In this case the capacity increase compared to CCM is about 21 Mbit/s.

Table 38 summarizes the results above.

**Table 38: Data rates of the IP transmission example for different operation modes (CCM, VCM/ACM)**

	100 % high efficient Mod/Cod (CCM) (1 024-QAM, CR = 8/9)	100 % robust Mod/Cod (CCM) (1 024-QAM, CR = 3/4)	VCM/ACM with 50 % robust Mod/Cod 50 % high eff Mod/Cod
Data Rate (32 MHz C2 channel)	269,3 MBit/s	226,65 Mbit/s	248 MBit/s

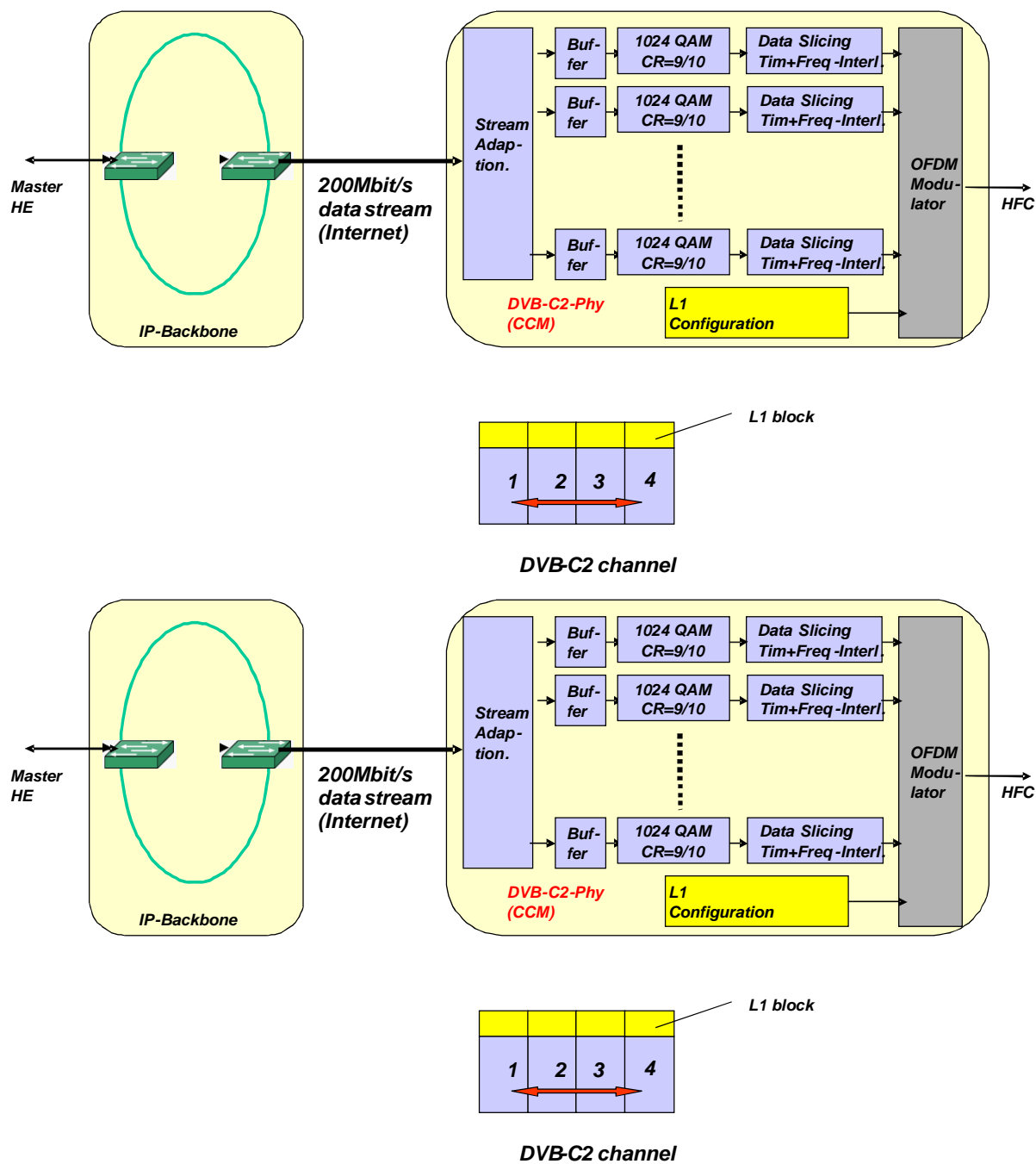
The requirement for Adaptive Coding and Modulation (ACM) occurs, if for example a certain service, which has been only transmitted to users with better CNR values before, is then also requested by users with lower CNR. At this moment the transmission parameters shall be changed from "high efficient" (CR = 8/9) to "robust" (CR = 3/4) for this service.

#### 12.1.4.4 PLP bundling

**PLP** bundling allows the transmission of a high bit rate stream. The bit rate per **PLP** is limited by the maximum **Data Slice** bandwidth of 7,61 MHz. By bundling of several **PLPs** and **Data Slices**, respectively, payload capacity can be significantly increased. The following example is given to elaborate this application in more detail.

The most likely application is the transmission of high speed internet with very high data rate requirement of up to several 100 Mbit/s. Figure 92 shows the basic principle. Calculation for **Data Slice** parameters (width, tuning frequency) basically follows the example given in clause 12.1.5.1, in **L1-part2** signalling bundled **PLP** shall be indicated (BUNDLED\_PLP="1") in table 37. The remaining structure of table 37 is also applicable for this bundled **PLP** application, except from some of the figures.

DVB-C2 compliant receivers may be fixed bandwidth, and therefore not supporting **PLP** bundling. However, DVB-C2 supports this functionality.



**Figure 92: Application example for PLP bundling; transmission of a high data rate MPEG2 Transport Stream via several bundled Data Slices in a DVB-C2 channel**

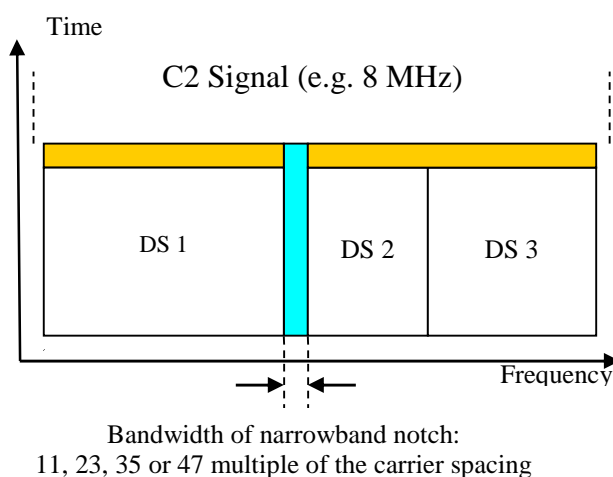
Of course, the application of channel bundling requires special CPEs with higher tuner bandwidth or multiple tuners. For more details of the receiver implementation see also clause 10.11.1.7.

It is not mandatory to transport bundled PLPs within one C2 system, Bundled PLPs may be transported in different C2 systems. Details are discussed in Annex C.

### 12.1.5 Handling of Interference scenarios

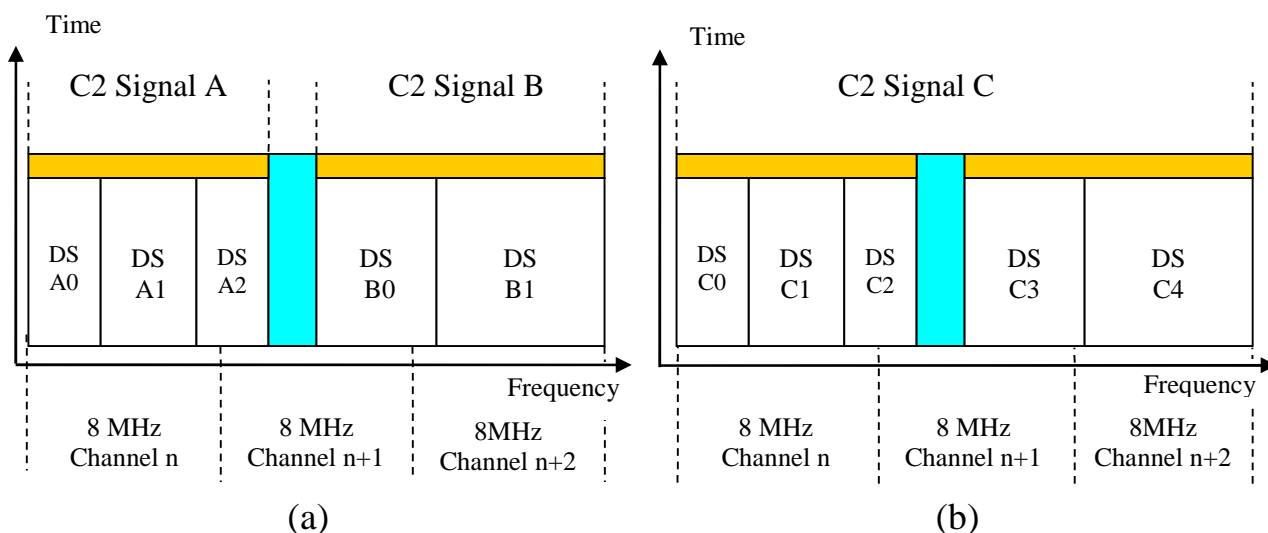
In several countries different frequency bands used in cable networks are allocated for radio service as well, including safety related services such as air traffic control. For this reason, interference scenarios, both cable signals interfering radio services and terrestrial services interfering cable signals are discussed in the present document. In some cases it is necessary to locally switch off (notch) certain frequency bands used e.g. for air traffic control services. Current aircraft radio services use the 25 kHz channel raster.

The DVB-C2 system allows to reduce power of sets of subcarriers or to switch off sets of OFDM carriers where interference scenarios cannot be overcome by other means. In many cases those problems are local problems only. Due to the strong error protection of the DVB-C2 preamble, narrowband **notches** may have up to  $47 \times 2,232 \text{ kHz} = 104,9 \text{ kHz}$  within 8 MHz channel bandwidth. Those narrow **notches** are called 'Narrowband **notches**' in the DVB-C2 system.



**Figure 93: Narrowband Notch with a maximum bandwidth of 104,9 kHz**

Figure 93 shows schematically a cable TV channel (8 MHz). The blue bar is a notched frequency band with a maximum bandwidth of 104,9 kHz.



**Figure 94: Two options for the implementation of Broadband Notches in a DVB-C2 signal**

NOTE 1: Figure 94 shows two options for the implementation of Broadband **notches** in a DVB-C2 signal:

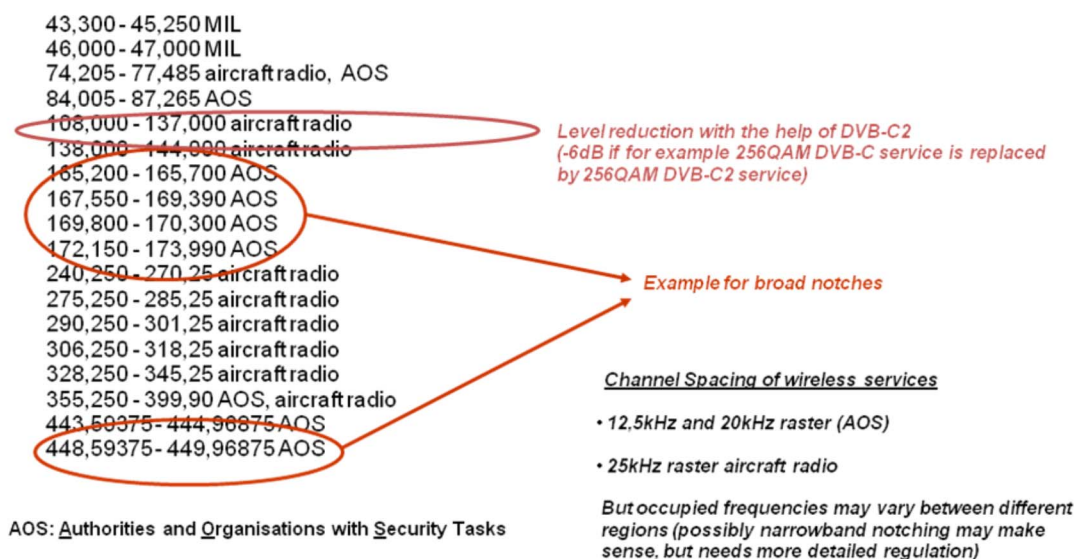
- Broadband **notch** placed between 2 adjacent DVB-C2 signal spectra; and
- Broadband **notch** placed within a single DVB-C2 signal spectrum.

Figure 94 schematically shows a C2\_System with a broadband notch and with two adjacent preambles and totally 5 **Data Slices**. The broadband notch is shown by the blue bar.

In case a Broadband **notch** (bandwidth >105 kHz) has to be implemented, there are generally two options possible with DVB-C2 system. The first option is to extend the frequency bands of the adjacent channels up to the lower and upper boarder of the broadband notch. This constellation is shown in figure 94 (a). With the second solution the broadband notch is accommodated within a C2\_System, which needs to have an overall bandwidth, which provides an overall bandwidth of the C2-System with at least a complete preamble in both the lower and the upper adjacent band to the broadband notch. This configuration is shown in figure 94 (b).

NOTE 2: Only in case of **Dependent Static Data Slices**, there may be less than a complete preamble bandwidth at the lower or the upper part of the C2-System adjacent to a broadband notch. For further details see also clause 9.3.5.2 in [i.2].

It is worth mentioning, that besides the options given with narrowband or broadband notch features, the fact, that DVB-C2 is significantly more robust in relation to DVB-C, can be used to mitigate interference scenarios.



**Figure 95: Example of the used opportunities from DVB-C2 to reduce the interference to wireless services**

Figure 95 shows a frequency allocation list of a set of different terrestrial services. Several examples of the different channel spacing of terrestrial wireless services are shown. Some of those potential interference conflicts might require narrowband or broadband notching, others might be solved by signal level adjustment in the critical bands.

In cable networks there is not only the egress problem (where cable networks are radiating interfering signals), but there is an ingress problem as well, where external radio services are interfering with cable signals. In the latter case DVB-C2 offers generally higher robustness in relation to DVB-C, which generally mitigates the interference problems.

In conclusion, DVB-C2 does provide powerful means to overcome interference conflicts, where other methods like improving the shielding of cables or receiving devices or adaptation of frequency planning does not provide appropriate solutions. Whereas with DVB-C the only choice was to switch off a complete 8 MHz channel, there are more and very flexible solutions in relation to adjustment of signal levels in parts of the signal available with DVB-C2. It should be mentioned that all those means of course result in a penalty of reduced usage or restriction of usage of the scarce frequency resources of cable networks.

## 12.2 Migration Scenarios

### 12.2.0 Introduction

The introduction of DVB-C2 requires the implementation of a migration strategy which allows continuing the transmission of traditional signals for a dedicated period of time. The strategies may differ among cable operators since they are based on many aspects and thus a general all-embracing migration cannot be discussed in the present document. However, a few specific features of DVB-C2 may be helpful for a cable operator in defining his strategy and thus are introduced hereafter.

### 12.2.1 From fixed to flexible channel raster

During transition from traditional transmission systems such as analogue TV and DVB-C to the new technology, DVB-C2 will have to coexist in a cable network side by side with analogue TV- and DVB-C/DOCSIS® signals. This requirement has an impact on certain DVB-C2 parameters selected for a DVB-C2 transmission such as power level, modulation constellation, FEC code rate, and last but not least signal bandwidth. This clause highlights bandwidths constraints and opportunities for DVB-C2 transmitted in different channel scenarios.

The transmission of analogue TV signals in cable networks will be reduced in terms of numbers throughout the coming years. This means that the channel occupancy will change while new channels and frequency bands will become available for the injection of DVB-C2 signals. As pointed out already, the bandwidth of a DVB-C2 signal can be flexibly assigned and depends on following considerations:

- 1) DVB-C2 was designed for utilization in cable systems with basic channel spacing of 6 MHz and 8 MHz, respectively. Depending on the time duration selected for the **Elementary period**  $T$ , all timing and frequency parameters of the system scale accordingly. For the two channel raster alternatives, the **Elementary period** is chosen to  $7/48 \mu\text{s}$  (6 MHz) and  $7/64 \mu\text{s}$  (8 MHz), respectively (see according to [i.2], table 38). As it is assumed that equipment be available for both solutions, cable operators can decide which parameter sets fit best to his implementation strategy. The most important parameters are summarized in [i.2] table 39 which is also included in the present document as table 38.

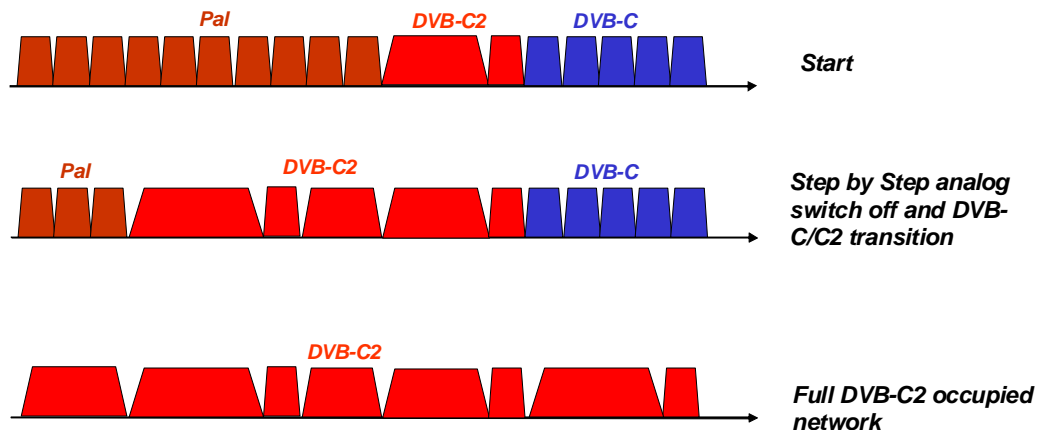
**Table 39: OFDM parameters for 8 MHz and 6 MHz systems**

Parameter	"6 MHz" 1/64	"6 MHz" 1/128	"8 MHz" 1/64	"8 MHz" 1/128
Number of OFDM carriers per <b>L1 block</b> $K_{L1}$	3 408	3 408	3 408	3 408
Bandwidth of L1 Signalling Block (see note)	<i>5,71 MHz</i>	<i>5,71 MHz</i>	<i>7,61 MHz</i>	<i>7,61 MHz</i>
Duration $T_U$	$4\,096T$	$4\,096T$	$4\,096T$	$4\,096T$
Duration $T_U \mu\text{s}$ (see note)	<i>597,3</i>	<i>597,3</i>	<i>448</i>	<i>448</i>
Carrier spacing $1/T_U$ (Hz) (see note)	<i>1 674</i>	<i>1 674</i>	<i>2 232</i>	<i>2 232</i>
Guard Interval Duration $\Delta T_U$	$64T$	$32T$	$64T$	$32T$
Guard Interval Duration $\Delta T_U \mu\text{s}$ (see note)	<i>9,33</i>	<i>4,66</i>	<i>7</i>	<i>3,5</i>
NOTE: Numerical values in italics are approximate values.				

- 2) The minimal bandwidth of a DVB-C2 signal is equal to the bandwidth of a **L1 part2** signalling block of some 5,71 MHz and 7,61 MHz for the respective cases of 6 MHz and 8 MHz channel spacing. This means for instance that an analogue TV signal transmitted in Europe within a 7 MHz channel in the VHF band cannot be replaced by the 8 MHz version but by the 6 MHz version.
- 3) The efficiency of a DVB-C2 increases with an increasing signal bandwidth.
- 4) It is necessary to inject in a single cable system DVB-C2 signals which all are based on one and the same **Elementary period**.

Taking these arguments into account, the cable operator has finally to decide which features meet their strategic requirements for the migration to DVB-C2 better. He has to make a trade-off between higher efficiency of the 8 MHz system and better frequency agility provided by the 6 MHz system. The different symbol durations  $T_S = T_U + \Delta T_U$  (see table 37) of the two system configurations will probably be taken into account as well.

Figure 96 depicts how DVB-C2 signals of varying bandwidths could be used to replace single signals as well as group of signals. More frequency spaces become available due to a smooth switch-off of analogue TV. Finally the entire frequency spectrum available may be filled with DVB-C2 signals.



**Figure 96: Spectrum occupation during the transition from analogue TV and DVB-C to DVB-C2**

## 12.2.2 Power level aspects

In addition to the flexible bandwidth allocation explained in the last clause, the correct adjustment of power levels is an important aspect to be considered for the introduction of DVB-C2. Two contradicting requirements have to be taken into account to identify the optimal power level for a signal.

- 1) The entire network load shall not increase by the introduction of new or different signals. This requirement needs to be strictly fulfilled at least as long as analogue TV signals are transmitted as part of the signal portfolio. Increasing network load would entail a reduction of signal-to-intermodulation ratio such as composite Triple Beat (CTB) and Composite Second Order (CTO) which will have a major negative impact on the technical picture quality provided by analogue TV signals.
- 2) The decrease of the DVB-C2 signal quality caused by an overlay of additive disturbances such as additive white Gaussian noise (AWGN) could be limited by increasing the level of the wanted signals - DVB-C2 in this case. This measure would result in high signal-to-noise ratios allowing the selection of higher order constellations and lower code rates which both have a positive impact on the actual spectral efficiency achieved.

Cable operators will seek to adjust the individual signal levels complying with both requirements. The first requirement could be met if DVB-C2 signals are injected in the network using the same level which was assigned to the replaced signal prior to its replacement. Figure 97 shows a comparison between power levels currently applied for DVB-C and possible levels applicable to DVB-C2 in the future. The power level of the analogue TV signal is shown as a reference. DVB-C signals using a 64-QAM schema are inserted in a cable network with a power back-off of 10 dB to 12 dB. The back-off of a DVB-C signal with a 256-QAM is 6 dB less and thus 4 dB to 6 dB below the reference level. A DVB-C 256-QAM signal could be replaced by a DVB-C2 signal using modulation constellation of a 1 024-QAM. The SNR requirements are the same for both signals while the DVB-C2 variant provides a higher spectral efficiency as explained in clause 11. The spectral efficiency can be increased further if the power budget allows an injection of DVB-C2 signals with a power level equal to the reference level since it would allow the adoption of a 4 096-QAM scheme. In case the power budget of future cable systems provides sufficient reserves, which may be possible for instance in networks with very deep fibre penetration with the utilization of state-of-the-art amplifier stages and after analogue switch-off, it could be conceivable that the power of some DVB-C2 signals would even exceed the reference level. The consequence may be that it could be possible to use even higher order constellations such as 16K or 64K in combination with the remaining DVB-C2 signal processing. However, these constellation schemes are not defined yet by the current version of the DVB-C2 standard [i.3] but may be subject to investigation for future updates.



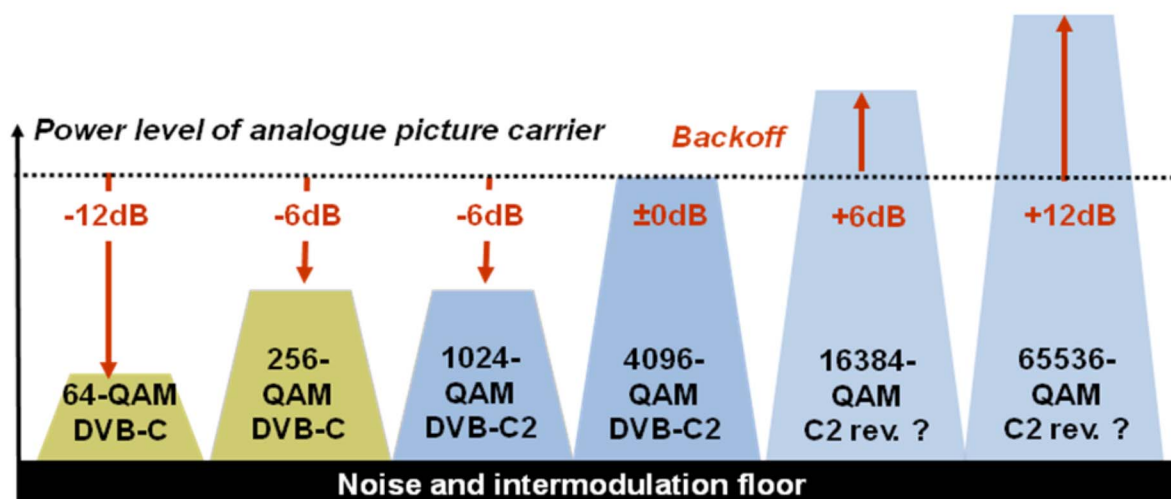


Figure 97: Examples for power levels applicable to DVB-C and DVB-C2 signals in cable networks

### 12.2.3 Non-backward compatibility to DVB-C

The signal processing defined by the DVB-C2 standard [i.4] does not provide any backward compatibility with DVB-C. A related commercial requirement was not agreed in the DVB Commercial Module. In fact, the lack of backward compatibility was accepted in order to allow the exploitation of the full potentials provided by the state-of-the-art transmission techniques adopted. Any backward compatibility requirement would diminish the performance or the flexibility of a DVB-C2 solution. In practice, however, backward compatibility will be established by equipping DVB-C2 compliant receivers with a DVB-C front-end in addition to the DVB-C2 front-end. Such devices will be capable of receiving and decoding both types of signals DVB-C and DVB-C2 while providing the means for a smooth migration from the traditional to the new technology.

### 12.2.4 Parameter Configurations Optimized for Data Transmission

Processing of bidirectional data services, e.g. in a DOCSIS<sup>®</sup> system, require C2 parameters to be optimized for low latency operation. The example six modes of operation defined in Table 38 meet the dual requirements of low latency and high bit-rate required of many applications. Time interleaving is provided in three of the modes to provide improved robustness with some tradeoffs on minimum achievable latency. Other low latency modes are possible in addition to those defined in Table 40, e.g. using guard interval = 1/64 for cable networks with a long delay spread but are not detailed further here.

Table 40: DVB-C2 Parameter Configurations for High Speed & Low Latency Data Transmission

Mode ID	Modulation	FEC Code Rate	Time Interleaving	Comment
1a	1 024-QAM	8/9	OFF	Direct replacement for DVB-C 256-QAM
1b	4 096-QAM	5/6	OFF	Better network SNR, maximum throughput
1c	4 096-QAM	8/9	OFF	Best network SNR, maximum throughput
2a	1 24-QAM	8/9	4 sym	As (1a) with improved robustness
2b	4 096-QAM	5/6	4 sym	As (1b) with improved robustness
2c	4 096-QAM	8/9	4 sym	As (1c) with improved robustness

Achievable latencies for each mode are dependent on a number of parameters and are to some extent implementation-dependent. The modes with the highest modulation and code rates provide the lowest latencies. As a general guide the latencies achievable in modes 1a, 1b & 1c are approximately 3- 4 ms and in modes 2a, 2b & 2c are 7-8 ms. This compares favourably with broadcast modes (e.g. annex B) where latencies can be typically 20-30ms.

Common parameters for each of the modes in table 40 are given in the following table 41.

**Table 41: Common Parameters for High Speed & Low Latency Data Transmission**

Parameter	Mode
PLP	Single
Data Slice	Type 1
Guard Interval	1/128
FEC Codeword Length	16K
ISSY	Not Used
Channel Bundling	Not Used

In Table 40 **PLP** bundling is not used. This is to simplify initial adoption of C2 as a new downstream PHY in cable networks supporting DOCSIS<sup>®</sup> by retaining the use of the DOCSIS<sup>®</sup> MAC layer for channel bonding of DVB-C2 channels.

## 12.2.5 Dynamic DVB-C2 Signalling Parameter Changes: Boundaries and Policies

The DVB-C2 system is designed to be flexible with respect to its parameterization. While the specification only describes the value semantics of the available parameters of the Layer 1 Signalling, this clause focuses on the parameter change boundaries and policies, i.e. the interaction with the demodulator.

**Table 42: Layer 1 Signalling parameters update boundaries and policies.**

Parameters	Boundary	Policy Seamless?
<b>L1 HEADER</b>		
L1_INFO_SIZE	CHGCNTR	YES
L1_TI_MODE	CHGCNTR	NO
<b>L1 SIGNALING PART 2 DATA</b>		
NETWORK_ID	STATIC	NO
C2_SYSTEM_ID	STATIC	NO
START_FREQUENCY	STATIC	NO
C2_BANDWIDTH	STATIC	NO
GUARD_INTERVAL	STATIC	NO
C2_FRAME_LENGTH	STATIC	NO
L1_PART2_CHANGE_COUNTER	DYNAMIC	YES
NUM_DSLICE	CHGCNTR	YES
NUM_NOTCH	STATIC	NO
<b>DATA SLICE</b>		
DSLICE_ID	STATIC	NA
DSLICE_TUNE_POS	CHGCNTR	NO
DSLICE_OFFSET_LEFT	CHGCNTR	YES
DSLICE_OFFSET_RIGHT	CHGCNTR	YES
DSLICE_TI_DEPTH	CHGCNTR	NO
DSLICE_TYPE	CHGCNTR	NO
FEC_HEADER_TYPE	CHGCNTR	NO
DSLICE_CONST_CONF	CHGCNTR	NO
DSLICE_LEFT_NOTCH	CHGCNTR	YES
DSLICE_NUM_PLP	CHGCNTR	YES
<b>PHYSICAL LAYER PIPE</b>		
PLP_ID	STATIC	NA
PLP_BUNDLED	CHGCNTR	NO
PLP_TYPE	CHGCNTR	NO
PLP_PAYLOAD_TYPE	CHGCNTR	NO
PLP_GROUP_ID	CHGCNTR	NO
PLP_START	DYNAMIC	YES
PLP_FEC_TYPE	CHGCNTR	NO
PLP_MOD	CHGCNTR	NO
PLP_COD	CHGCNTR	NO
PSI/SI_REPROCESSING	CHGCNTR	NO
TRANSPORT_STREAM_ID	CHGCNTR	NO
ORIGINAL_NETWORK_ID	CHGCNTR	NO
RESERVED_1	IGNORED	NA
<b>FEC FRAME HEADER</b>		
PLP_ID	STATIC	NA
PLP_FEC_TYPE	DYNAMIC	YES

PLP_MOD	DYNAMIC	YES
PLP_COD	DYNAMIC	YES
HEADER_COUNTER	DYNAMIC	YES
RESERVED_2	IGNORED	NA
<b>NOTCH</b>		
NOTCH_START	STATIC	NO
NOTCH_WIDTH	STATIC	NO
RESERVED_3	IGNORED	NA
RESERVED_TONE	STATIC	NO
EARLY_WARNING_SYSTEM (EWS)	DYNAMIC	YES
C2_VERSION	STATIC	NO
RESERVED_4	IGNORED	NA

The change boundaries impose the following semantics:

- [ STATIC]  
The value of the relevant parameter is constant throughout the lifetime of the particular **C2 system**.
- [DYNAMIC]  
The value of the relevant parameter is subject to changes on its enclosing boundary, i.e. a parameter of the Layer 1 Signalling Part 2 Data may be changed dynamically on an OFDM frame boundary while a parameter of the Layer 1 Signalling Part 1 Data may be changed dynamically on a FECFRAME boundary.
- [CHGCNTR]  
The value of the relevant parameter is subject to the boundary imposed by the semantics of the parameter L1\_PART2\_CHANGE\_COUNTER.
- [IGNORED]  
The value of the relevant parameter is ignored as defined in the specification.

The seamless change policy imposes the following semantics:

- [YES]  
The value of the relevant parameter shall be processed seamlessly within the imposed update boundary, i.e. no service interruption shall occur.
- [ NO]  
The value of the relevant parameter need not to be processed seamlessly.
- [ NA]  
Not Applicable

Special notes:

- L1\_INFO\_SIZE  
The value of L1\_INFO\_SIZE and thus the decoding of the Layer 1 signalling shall be processed seamlessly as long as the number of **preamble symbols** remain the same between updates.  
Nevertheless, L1\_INFO\_SIZE will change only on the CHGCNTR boundary, since all relevant parameters that influence the size of the Layer 1 Signalling are also defined to be within the CHGCNTR boundary.
- DSLICE\_ID, PLP\_ID  
The values of DSLICE\_ID and PLP\_ID are static within a **C2 system**, since no mechanisms exist to signal a change of an ID for the very same **Data Slice** or **PLP**. Nevertheless, new **Data Slices** and/or **PLPs** can dynamically and seamlessly be added to and deleted from the **DVB-C2 system**, the latter case resulting in undefined behaviour if the **Data Slice** or **PLP** is currently being decoded.
- Parameter subgroups for **Data Slices** and **PLPs**  
Reordering the parameter subgroups for **Data Slices** and **PLPs** within the Layer 1 Signalling is dynamic and shall be processed seamlessly.

## Annex A (informative): Examples for the calculation of payload capacity of the DVB-C2 system

### A.1 Introduction

Annex A gives four examples for calculated payload capacities of DVB-C2\_systems with different parameters.

### A.2 Examples for payload capacity using Guard Interval 1/128 using 2,232 kHz OFDM carrier spacing (for European type cable networks)

The following set of parameters applies for table A.1: one symbol for the preamble, single **PLP** and single **Data Slice** with **Data Slice** type 1.

**Table A.1: DVB-C2 Payload capacity for Guard Interval 1/128**

	OFDM 8 MHz channel bandwidth	OFDM 16 MHz channel bandwidth	OFDM 24 MHz channel bandwidth	OFDM 32 MHz channel bandwidth	OFDM 64 MHz channel bandwidth
Mode	Throughput (Mbit/s) per 8 MHz	Throughput (Mbit/s) per 8 MHz	Throughput (Mbit/s) per 8 MHz	Throughput (Mbit/s) per 8 MHz	Throughput (Mbit/s) per 8 MHz
16-QAM 4/5	23,56	24,16	24,36	24,46	24,61
16-QAM 9/10	26,51	27,19	27,42	27,53	27,70
64-QAM, 2/3	29,45	30,20	30,45	30,58	30,77
64-QAM 4/5	35,33	36,24	36,54	36,70	36,92
64-QAM 9/10	39,77	40,79	41,13	41,30	41,55
256-QAM 3/4	44,16	45,29	45,67	45,86	46,14
256-QAM 5/6	49,11	50,37	50,79	51,00	51,32
256-QAM 9/10	53,02	54,39	54,84	55,07	55,41
1 024-QAM 3/4	55,20	56,61	57,09	57,32	57,68
1 024-QAM 5/6	61,39	62,97	63,49	63,75	64,15
1 024-QAM 9/10	66,28	67,98	68,55	68,83	69,26
4 096-QAM 4/5	70,67	72,48	73,09	73,39	73,84
4 096-QAM 5/6	73,67	75,56	76,19	76,51	76,98
4 096-QAM 9/10	79,54	81,58	82,26	82,60	83,11

### A.3 Examples for payload capacity using Guard Interval 1/64 using 2,232 kHz OFDM carrier spacing (for European type cable networks)

The following set of parameters applies for table A.2: one symbol for the preamble, single **PLP** and single **Data Slice** with **Data Slice** type 1.

Table A.2: DVB-C2 Payload capacity for Guard Interval 1/64

	OFDM 8 MHz channel bandwidth	OFDM 16 MHz channel bandwidth	OFDM 24 MHz channel bandwidth	OFDM 32 MHz channel bandwidth	OFDM 64 MHz channel bandwidth
Mode	Throughput (Mbit/s) per 8 MHz	Throughput (Mbit/s) per 8 MHz	Throughput (Mbit/s) per 8 MHz	Throughput (Mbit/s) per 8 MHz	Throughput (Mbit/s) per 8 MHz
16-QAM 4/5	23,14	23,74	23,93	24,03	24,18
16-QAM 9/10	26,04	26,71	26,94	27,05	27,21
64-QAM, 2/3	28,93	29,67	29,92	30,04	30,23
64-QAM 4/5	34,71	35,60	35,90	36,05	36,27
64-QAM 9/10	39,07	40,07	40,40	40,57	40,82
256-QAM 3/4	43,38	44,49	44,86	45,05	45,33
256-QAM 5/6	48,25	49,48	49,90	50,10	50,41
256-QAM 9/10	52,09	53,43	53,87	54,09	54,43
1 024-QAM 3/4	54,22	55,62	56,08	56,31	56,66
1 024-QAM 5/6	60,31	61,86	62,37	62,63	63,02
1 024-QAM 9/10	65,11	66,78	67,34	67,62	68,04
4 096-QAM 4/5	69,42	71,21	71,80	72,10	72,54
4 096-QAM 5/6	72,37	74,23	74,85	75,16	75,62
4 096-QAM 9/10	78,13	80,14	80,81	81,14	81,64

## A.4 Examples for payload capacity using Guard Interval 1/128 using 1,674 kHz OFDM carrier spacing (for US type cable networks)

The following set of parameters applies for table A.3: one symbol for the preamble, single **PLP** and single **Data Slice** with **Data Slice** type 1.

Table A.3: DVB-C2 Payload capacity for Guard Interval 1/128

	OFDM 6 MHz channel bandwidth	OFDM 12 MHz channel bandwidth	OFDM 18 MHz channel bandwidth	OFDM 24 MHz channel bandwidth	OFDM 48 MHz channel bandwidth
Mode	Throughput (Mbit/s) per 6 MHz	Throughput (Mbit/s) per 6 MHz	Throughput (Mbit/s) per 6 MHz	Throughput (Mbit/s) per 6 MHz	Throughput (Mbit/s) per 6 MHz
16-QAM 4/5	17,67	18,12	18,27	18,35	18,46
16-QAM 9/10	19,89	20,40	20,57	20,65	20,78
64QAM, 2/3	22,09	22,65	22,84	22,94	23,08
64-QAM 4/5	26,50	27,18	27,41	27,52	27,69
64-QAM 9/10	29,83	30,59	30,85	30,98	31,17
256-QAM 3/4	33,12	33,97	34,25	34,40	34,61
256-QAM 5/6	36,84	37,78	38,10	38,25	38,49
256-QAM 9/10	39,77	40,79	41,13	41,30	41,56
1 024-QAM 3/4	41,40	42,46	42,82	42,99	43,26
1 024-QAM 5/6	46,05	47,23	47,62	47,82	48,11
1 024-QAM 9/10	49,71	50,99	51,41	51,63	51,95
4 096-QAM 4/5	53,00	54,36	54,82	55,04	55,38
4 096-QAM 5/6	55,25	56,67	57,15	57,38	57,74
4 096-QAM 9/10	59,66	61,19	61,70	61,95	62,34

## A.5 Examples for payload capacity using Guard Interval 1/64 using 1,674 kHz OFDM carrier spacing (for US type cable networks)

The following set of parameters applies for table A.4: one symbol for the preamble, single **PLP** and single **Data Slice** with **Data Slice** type 1.

**Table A.4: DVB-C2 Payload capacity for Guard Interval 1/64**

	OFDM 6 MHz channel bandwidth	OFDM 12 MHz channel bandwidth	OFDM 18 MHz channel bandwidth	OFDM 24 MHz channel bandwidth	OFDM 48 MHz channel bandwidth
Mode	Throughput (Mbit/s) per 6 MHz	Throughput (Mbit/s) per 6 MHz	Throughput (Mbit/s) per 6 MHz	Throughput (Mbit/s) per 6 MHz	Throughput (Mbit/s) per 6 MHz
16-QAM 4/5	17,36	17,80	17,95	18,03	18,14
16-QAM 9/10	19,53	20,04	20,20	20,29	20,41
64QAM, 2/3	21,70	22,25	22,44	22,53	22,67
64-QAM 4/5	26,04	26,70	26,93	27,04	27,20
64-QAM 9/10	29,30	30,05	30,30	30,43	30,62
256-QAM 3/4	32,54	33,37	33,65	33,79	34,00
256-QAM 5/6	36,18	37,12	37,43	37,58	37,81
256-QAM 9/10	39,07	40,07	40,41	40,57	40,82
1 024-QAM 3/4	40,67	41,71	42,06	42,24	42,50
1 024-QAM 5/6	45,23	46,39	46,78	46,98	47,27
1 024-QAM 9/10	48,84	50,09	50,51	50,72	51,03
4 096-QAM 4/5	52,07	53,40	53,85	54,07	54,41
4 096-QAM 5/6	54,28	55,67	56,14	56,37	56,72
4 096-QAM 9/10	58,60	60,11	60,61	60,86	61,24

## Annex B (informative): Example for the choice of a DVB-C2 parameter set for broadcasting application

This annex gives an example for a set of DVB-C2 parameters chosen for the introduction of DVB-C2 modulation planned in the cable network of Kabel Deutschland for Video on Demand (VOD) and High Definition TV applications. The VOD services and the HDTV services are delivered via a DVB transport stream and highest spectrum efficiency for the given guaranteed carrier to noise ratio (CNR) was the key decision criterion for that set of parameters.

- Modulation parameters
  - Modulator Bandwidth: 8 MHz
  - Carrier spacing: 2,232 kHz
  - Constellation: 1 024 QAM
  - Code rates: 9/10
  - Guard Interval: 1/128
  - Reserved Tones: None
  - Narrowband **notches**: None
  - Broadband **notches**: None
  - Constant Coding and Modulation: Yes
  - Variable Coding and Modulation: No
  - Adaptive Coding and Modulation: No
- Framing parameters
  - Number of **PLPs**: 1
  - Number of **Data Slices**: 1
  - **Data Slice** Type: 1
  - **Data Slice** stuffing: no
  - Preamble symbols: 1 symbol (due to the simple multiplexing structure)
  - **Data Slice** TI mode: 8 symbols
  - L1 TI mode: 0 symbols (no time interleaving, due to the given **preamble symbol** length)
- Physical Layer Pipe (**PLP**) parameters
  - FEC Frame length: 64 800 bit

- Pre-processing parameters
  - Input mode: Single DVB Transportstream
  - Generic Stream Encapsulation: no
  - High Efficiency Mode: yes
  - Input stream synchronization (ISSY): no
  - Additional Null Packet insertion: no
  - Null packet deletion: no
  - Baseband Frame stuffing: yes
  - Common **PLP**: no



## Annex C (informative): Use of PLP bundling

As specified in annex F in [i.2], DVB-C2 allows optional bundling of **PLPs** carried in different **Data Slices**. **PLP** bundling is possible within a single **C2 system** as well as across different **C2 systems**. **PLP** bundling across different **C2 systems** allows for larger frequency distances of the **Data Slices**. Within a **C2 system**, in **PLP** bundling mode a single **PLP** can allocate arbitrary bandwidths up to the whole **C2 system** bandwidth. Annex F in [i.2] mentions that in this case all data packets of a bundled **PLP** connection pass the same input processing block. Inserting the **ISSY** timestamp in the mode adaptation block of every **BBFrame** is mandatory for this operation mode in order to allow the correct reordering of the packets from different **Data Slices** on the receiver side. At the output of the input processing block the **BBFrames** of the bundled **PLP** are spread over the different **Data Slices**. Figure F.1 and figure F.2 [i.2], show the block diagram for the **PLP** Bundling operation mode within a single **C2 system** or across different **C2 systems**.

The basic transmitter side processing is shown in figure C.1:

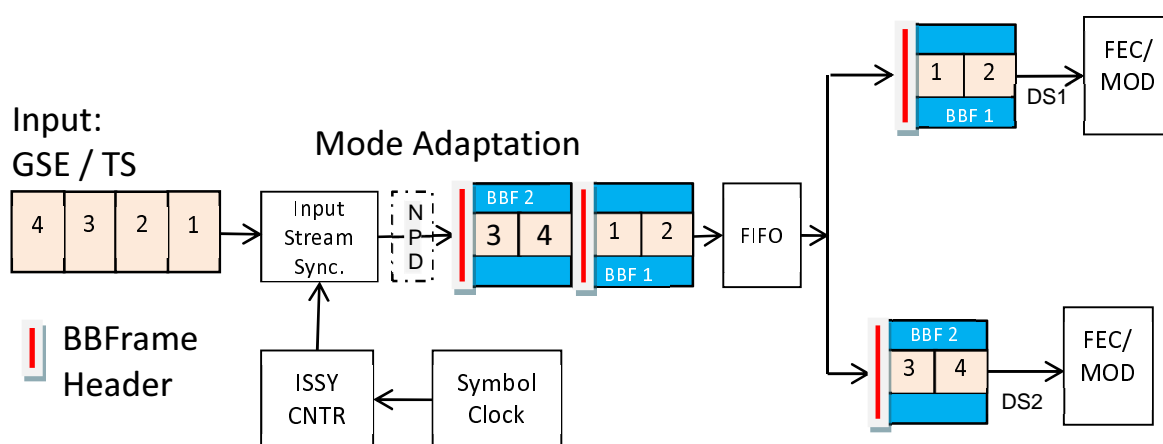


Figure C.1: Transmitter side processing for PLP bundling

As explained in [i.1], all data packets of a **PLP** transmitted in bundling mode pass the same input processing block, where the addition of the **ISSY** timestamp in the mode adaptation block takes place. This is required to allow the correct reordering of the packets from different **Data Slices** on at the receiver side. The High Efficiency Mode stream format is used for **TS** as well as for **GSE**. At the output of the input processing block the **BBFrames** of the bundled **PLP** are spread over the different **Data Slices**.

**BBFrames** are created for each **Data Slice** according to the ratio of the bitrate of each **Data Slice**. The following figure C.2 shows the load balancing principle for 2 **Data Slices**:

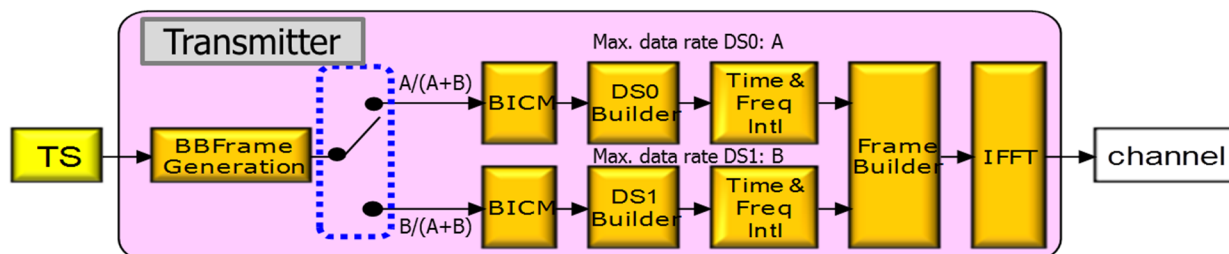


Figure C.2: Load balancing for bundled PLPs

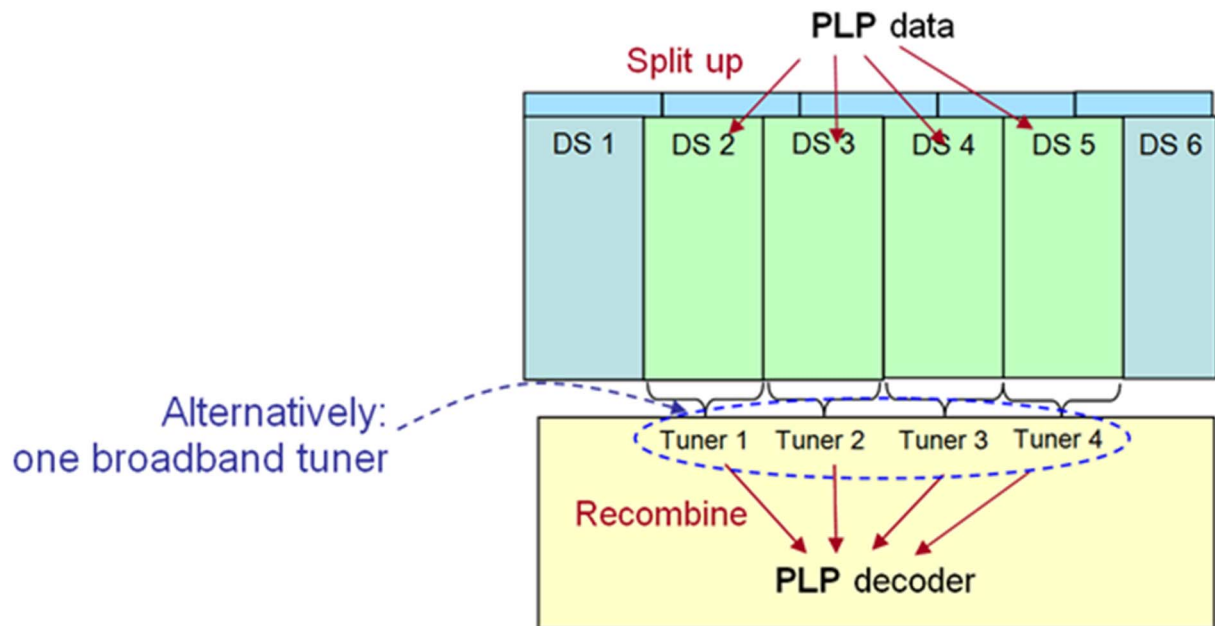
In this example for 2 **Data Slices**, let  $A$  be the maximum data rate of **Data Slice 1** and  $B$  the maximum data rate of **Data Slice 2**. Then  $A/(A+B)$  of **BBFrames** are assigned to **Data Slice 1** and  $B/(A+B)$  of **BBFrames** to **Data Slice 2**.

This allows for a minimum necessary and deterministic buffer size on receiver side. For example if the bitrates of two bundled **Data Slices** are equal, **BBFrames** for each **Data Slice** occur in alternating fashion.

The ratio of the highest and lowest payload capacity of **Data Slices** carrying bundled **PLPs** is not permitted to exceed the ratio of 5:1. For example, in case of **PLP** bundling across two **Data Slices** not more than 5 consecutive **BBFrames** has to be assigned to the same **Data Slice**.

A DVB-C2 demodulator, which supports the **PLP** bundling mode, therefore applies the inverse signal processing. All data packets marked to be bundled and the targeted **PLP\_ID** are demodulated and FEC processed. Data packets of the targeted **PLP** are multiplexed in the order given by the **ISSY** timestamp. As the **BUFSTAT** field is not available in case of **PLP** bundling, the demodulator recovers the output clock on basis of transmitted **ISCR** field of the master channel and has to establish an appropriate output buffer management, which does not require the **BUFSTAT** information.

NOTE: The master channel is signalled in the C2 bundled delivery system descriptor (see ETSI EN 300 468 [i.9]).



**Figure C.3: Bundled PLPs - Data Slice split up and recombining process at the receiver**

Figure C.3 shows the recombining process of the data of a bundled **PLP**. Different data packets of a bundled **PLP** that are split up into different **Data Slices** are collected on receiver side. After demodulation and FEC decoding the data of the **BB** packets can be resorted according to the increasing value of their **ISSY** timestamp in the **BB** header.

The demodulation of the different **Data Slices** can be realized in different ways and is implementer dependant: The first implementation example is that each **Data Slice** is decoded by a separate 8 MHz tuner. Another option is to apply a broadband tuner architecture that is capable to cover the required frequency range within its tuning window. In this case the selection of the appropriate subcarriers of each **Data Slice** is done after a larger reception FFT (e.g. the example given in figure C.1 would require four times the standard 4k **FFT size**: 16k **FFT**).

Note that data packets of a bundled **PLP** do not need to be located in neighboured **Data Slices** nor in the same **C2 system**, as long as the reception architecture can cover the applied split (i.e. sufficient 8 MHz tuners or a single broadband tuner covering the complete required frequency range).

## Annex D (informative): The concept of optimized frequency utilization

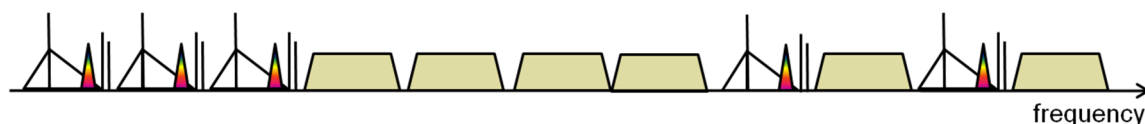
### D.1 Introduction

The following explanations of annex D are results of investigations carried out by the ReDeSign project. They analyze the possibilities for the utilization of DVB-C2 in HFC networks and the potentials provided by the new technology. In the first clause, state of the art RF characteristics of HFC networks are explained to an extent important to understand the impacts caused by and the limitations for DVB-C2. Three basic methods are explained which show how DVB-C2 signals could be transmitted in compliance with the existing RF spectrum. The physical characteristics of DVB-C2 can be configured in a very flexible manner, which allows deviating from the traditional cable channel pattern of 6 MHz and 8 MHz, respectively. Instead of a channel number, a new characteristic frequency value is introduced which takes account of the new approach implemented in DVB-C2 called Absolute OFDM. Further to requirements in the frequency domain, the evaluation of future usage scenarios for DVB-C2 requires the consideration of the actual network load in association with the different architectures implemented in the last mile of HFC networks. A more detailed investigation of which network load could be transmitted through which topology is described in the clause 9 of the present document.

### D.2 RF Spectrum Considerations

For the introduction of DVB-C2 signals in scenarios provided by today's existing HFC networks a number of technical requirements needs to be taken account. The most important requirement certainly is the demand to prevent any reduction of technical quality of the traditional signals. DVB-C2 signals fulfil this important requirement if the transition parameters applied comply with the respective parameters defined by the relevant standards published by IEC and CENELEC, see [i.9]. In particular channel frequencies and power levels are of interest meaning that DVB-C2 signals are to be injected in the existing channel raster at a defined power level. For the following discussion it is assumed that a sub-band of the RF frequency spectrum is occupied with signals as indicated in below figure a) by means of an example. A simple means introducing DVB-C2 is provided by a one-to-one exchange of an individual signal such as analogue TV or DVB-C with a DVB-C2 signal while keeping the signal power at the dedicated level. The resulting RF spectrum of such a simple example is depicted in figure b). Two DVB-C signals transmitted in adjacent channels and one analogue TV signal are replaced each by a DVB-C2 signal. Figure c) shows the possibility to transmit a wide-band DVB-C2 signal of a bandwidth equivalent to two cable channels (e.g. 16 MHz in Europe, 12 MHz in the U.S.). The benefit arising from the injection of wide-band signals in two or more adjacent channels consists of an increase of spectral efficiency caused by an active use for data transmission of the frequency Guard Band traditionally applied between two adjacent signals. In figure c) such frequency band is marked by the coloured ellipse. Compared to this example of a combination of two channels, the spectral efficiency can be further accelerated by either transmitting signals of even wider bandwidths and combining a higher number of channels or by injecting individual DVB-C2 signals adjacent to each other and without making use of any frequency Guard Band. Such a close packet signals line-up can be achieved if the signals comply with the pre-requisites defined for the special case of an Optimized Frequency Utilization. The pre-requisites are illustrated further down in this clause. An example of the resulting RF spectrum is depicted in figure D.1.

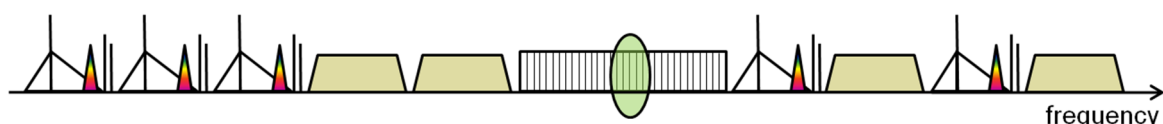
- a) Example of original RF spectrum prior to DVB-C2 introduction



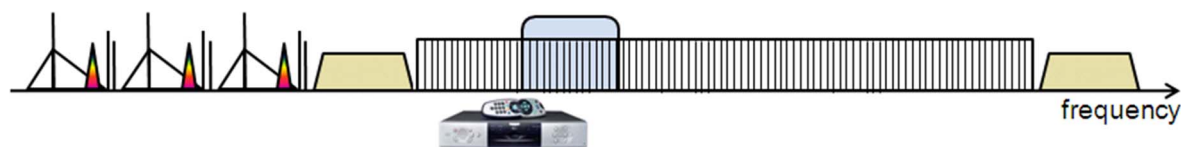
- b) Insertion of DVB-C2 according to IEC 60728 (traditional frequency raster and power levels)



c) Combination of more than one channel of the traditional frequency raster



d) Principle of optimized frequency utilization



Symbols used in this figure



Figure D.1: Opportunities for RF spectrum utilization

As described already in various clauses above, DVB-C2 receivers using tuners with a traditional receiving bandwidth (of 8 and 6 MHz, respectively) are capable of receiving individual **Data Slices** or bundles of **Data Slices** transmitted in related frequency bands. Guided through the DVB-C2 Signalling Information, the CPE will be tuned to the **Data Slice** which carries the service of interest, e.g. an HD program.

## D.3 Optimized Frequency Utilization

The pre-requisite for the application of an Optimized Frequency Utilization is the establishment of a mode requiring the full synchronization of all DVB-C2 signals being part of a DVB-C2 ensemble. The synchronization has to be created in both time and frequency domain. It is necessary to ensure that the physical behaviour of all synchronized DVB-C2 signals of the entire ensemble is identical with the behaviour of a single wide-band DVB-C2 signal. In fact the orthogonality of the OFDM sub-carriers applied in the frequency domain has to be guaranteed not only within an isolated **OFDM symbol** but also among the sub-carriers being part of each adjacent **OFDM symbol** and thus of each **OFDM symbol** contributing to the ensemble of DVB-C2 signals. This synchronization requirement needs to be fulfilled at least during the active symbol duration (excluding the temporal Guard Interval). Therefore the DVB-C2 signals of the ensemble have to be synchronized in time domain as well.

The spectrum of such an ensemble is shown in figure D.2. While the bandwidth of a single DVB-C2 signal can be assigned in a very flexible manner, the bandwidth of such an ensemble of DVB-C2 signals can vary significantly and has only a lower limit which is equal to the minimal bandwidth of a single DVB-C2 signal.

## D.4 Spectral implications of the Absolute OFDM mechanism

The introduction of the so-called Absolute OFDM concept (see clause 5.9) entails the definition of a new set of RF parameters corresponding to the traditional RF channel numbering. The OFDM Sub-carrier Index and the L1 Channel Number. As shown in figure D.2, the entire RF frequency band used for up- and downstream transmission in HFC networks is subdivided for this purpose in sub-bands of the OFDM sub-carrier bandwidth:

$$f_{SC} = 1/448 \text{ MHz} \approx 2,232 \text{ kHz}$$

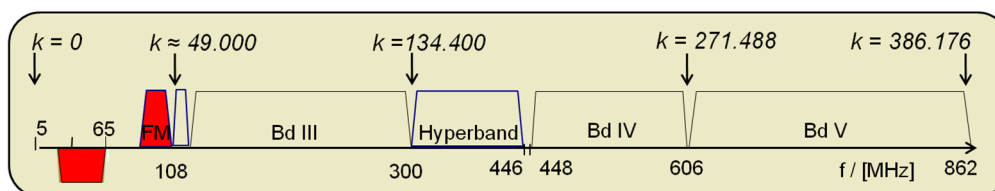
And parameterized with a running index  $k$ . The Sub-carrier Index  $k$  starts with a value of 0 indicating the frequency sub-band at DC (0 Hz) and ends at a value of 386,176 which corresponds with a frequency of 862 MHz. The interrelationship between the index  $k$  and the medium frequency of the related frequency sub-band is given by the formula:

$$f_{sc,k} = k/448 \text{ MHz} \text{ with } f_{sc,\min} = 0 \text{ MHz} \text{ and } f_{sc,\max} = 862 \text{ MHz}$$

As explained already in clause 8.4, **L1 blocks** are aligned to a fixed 7,61 MHz frequency grid which corresponds to 3 408 OFDM sub-carriers. The theoretical concept arranges for the first **L1 block** starting at 0 Hz whereas the subsequent **L1 blocks** starting at sub-carrier indexes of a multiple number of 3 408. They can be calculated as follows:

$$k_{L1-Start} = n \cdot 3,408 \text{ for } n = 0,1,2,\dots,113$$

Assuming that the FM radio band will be continuously used for analogue audio transmission, the first entire **L1 block** transmitted starts at  $k_{L1-Start, \min} = 15$  corresponding to 114,1 MHz. However, parts of the **L1 block** could also be transmitted below this frequency but above 108 MHz which is the upper frequency limit of FM radio. A protection distance between DVB-C2 and the FM signal needs to be taken into account when introducing DVB-C2 in these frequency bands. The required protection ratio is currently investigated by ReDeSign and under standardization by CENELEC and IEC.



**Figure D.2: Definition of Absolute OFDM sub-carrier grid: number of absolute carrier  $k$  and respective frequency in MHz**

NOTE: More details on the concept of optimized spectrum utilization are discussed in [i.24].

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# Annex E (informative): DVB-C2 system evaluation and performance testing

## E.1 Introduction

Already during the phase of development of the technical specification of the DVB-C2 system and during the system evaluation phase as well the DVB-C2 validation & verification experts teams has been very active in simulating the characteristics and performance of individual system building blocks and the whole DVB-C2 modulation system. Key findings of this experts group and a summary of the measurement configurations and methodologies used during the first two DVB-C2 Plug Fests are documented in this informative Annex E.

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## E.2 Interoperability Testing

The basic interoperability tests (functional tests) focus on the test of interoperability between modulators and demodulators. However, *the main goal of the Verification and Validation (V&V) group is to ensure that the specification is unambiguous and that all companies involved have the same understanding of it.*

The process of Validation & Verification may be subdivided into two different approaches that differ in the process of testing for standard conformity and interoperability, respectively. While the process of validating the standard purely focuses on the generation of DVB-C2 signals and its conformance to the DVB-C2 standard [i.5], the interoperability testing does not include the introspection of the signals but tests for an error free decoding of the signal where influences other than the coding and decoding process can be excluded.

Testing devices for interoperability can basically be interpreted as check for end-to-end link transparency when applied to a **C2 system**, i.e. testing the correct and error free transmission of a data stream. This also leads to possible circumstances in which two devices are interoperable since the streams can be decoded error free but are not necessarily standard conformant.

The DVB-C2 system includes a large set of features and configuration options that makes it virtually impossible to test DVB-C2 implementations using all possible/allowed parameter combinations. Therefore, grouping of parameters and features and thus reducing the number of the configurations of a **C2 system** by applying multiple **PLPs** may be of advantage when testing devices for interoperability.

### Basic parameters of the DVB-C2 system:

In principle, the set of parameters of the DVB-C2 system, i.e. the parameters of the layer 1 signalling part 1 and part 2 data as well as the **preamble header**, can be subdivided into enumerations and rational quantities. The parameters which provide a limited number of options, i.e. that can be enumerated, will be referred to as basic parameters. The basic parameters included are: `L1_TI_MODE`, `GUARD_INTERVAL`, `DSLICE_TI_DEPTH`, `DSLICE_TYPE`, `PLP_FEC_TYPE`, `PLP_MOD`, `PLP_COD`, `HEADER_COUNTER`, `RESERVED_TONE`. Furthermore, the parameters `NUM_DSLICE` and `NUM_PLP` (per **Data Slice**) are referred to as basic parameters as well. All these parameters have in common, that no special interpretation or semantics during the generation and decoding process are required, which would otherwise denote them as parameters used in conjunction with special features of the DVB-C2 system (see below).

During the first and second Plug-Fest the basic parameters and the (available and valid) combinations have been used to test the basic features of a DVB-C2 system and were referred to as the baseline Test Cases. The devices were tested using the minimum available bandwidth of a DVB-C2 signal of 7,61 MHz, i.e. 8 MHz operation. Testing each parameter, its options and combinations with other parameters in sequence is time-consuming and may be disregarded in favour of a more sophisticated approach. Thus, during the first and second Plug-Fest performed the approach of combining the basic parameters in dedicated Test Cases was applied. Here, multiple **PLPs** and multiple **Data Slices** were configured for a single DVB-C2 signal where each **PLP** and **Data Slice**, respectively, used different settings. To limit the impact of side effects e.g. due to narrowing the bandwidth of **Data Slices** below a certain threshold which might interfere with any stream data rate requirement the number of **PLPs** and/or **Data Slices** might be limited to four (or less), as done for the Plug-Fests.

### Extended features of the DVB-C2 system:

In addition to the baseline Test Cases further groups of tests were derived covering major features of the DVB-C2 system. The testing of these features or groups, respectively, may be performed orthogonal to the basic parameters, i.e. only one or a specific subset of the basic parameters need to be tested in conjunction with the extended features of the DVB-C2 system to provide a certain amount of confidence for interoperability.

The extended features/groups as derived for the second Plug-Fest are as follows:

- **Minimum Bandwidth of Data Slices:** This group includes Test Cases to test the minimum allowed bandwidth of **Data Slices** for both types of **Data Slices**.
- **Broadband Signals:** This group includes Test Cases for testing signal configuration exceeding the minimum bandwidth of a DVB-C2 signal of 7,61 MHz, e.g. 14 MHz, 16 MHz or 32 MHz.
- **Narrow Band Notches:** This group includes Test Cases for testing notch configuration solely deploying narrow band **notches** with DVB-C2 signals of minimum bandwidth. The **notches** are configured at different locations within the signal as well as between or overlapping **Data Slices**.
- **Broad Band Notches:** This group includes Test Cases for testing notch configurations deploying broadband **notches** for DVB-C2 signals exceeding 8 MHz bandwidth. The configurations include testing of broadband **notches** at different locations and bandwidth as well as testing Dependent Static **Data Slices** with minimum bandwidth.
- **Signal Parameter Changes:** This group includes Test Cases explicitly testing the change of layer 1 signalling parameters which should not result in a service interruption during the decoding process. (see clause 12.2.5)

These groups of Test Cases were tested in the course of the second DVB-C2 Plug-Fest resulting in a total of 35 Test Cases (per modulator), i.e. 7 baseline Test Cases + 28 Test Cases of extended features.

### Error indication and criterion:

Two systems can be referred to as interoperable with respect to the transmission of data via a **C2 system** if the input to the modulator and output of the demodulator device result in the very same data/stream, i.e. the transmission link is transparent and error free. To test for an error free decoding (generation, transmission and decoding) of the transmitted stream several means exist:

- **Using the video/audio content of the transmitted stream:** The video (and audio) content of a transmitted stream may be used to check for errors during the generation, transmission or decoding of the stream by evaluating any block artifact in the picture.
- **Using a transport stream analyser:** A **Transport Stream** analyser may be used to check for any errors in the transmitted stream.
- **Using an error indicator of Forward Error Correction:** The most valuable way of checking for errors during reception is the evaluation of the error indication of the forward error implementation in the decoder, as this block can also detect errors which might be hidden from the stream, e.g. if the errors occur within stuffing regions of the baseband frame.

### Special notes about interoperability tests with DVB-C2 systems:

**Signal quality and high constellation sizes:** DVB-C2 supports the 1 024-QAM and 4 096-QAM constellation sizes which demand high performance figures with respect to transmission conditions. When testing DVB-C2 configuration deploying these QAM modes an error free reception of the signal might be interfered by insufficient signal quality and thus lead to misinterpretations with respect to interoperability. Thus, it should be assured when performing interoperability tests with high constellation sizes (or features degrading the performance of the signal itself, such as narrow band **notches** located within **Data Slice**) that the signal quality is sufficiently high.

**Stream rate requirements:** When configuring a **C2 system** for interoperability testing care is taken about the input stream rate requirement, i.e. the data rate of the **PLP** provided for the embedded stream needs to be always greater than the required data rate of the stream to prevent loss of data and possible misinterpretations.

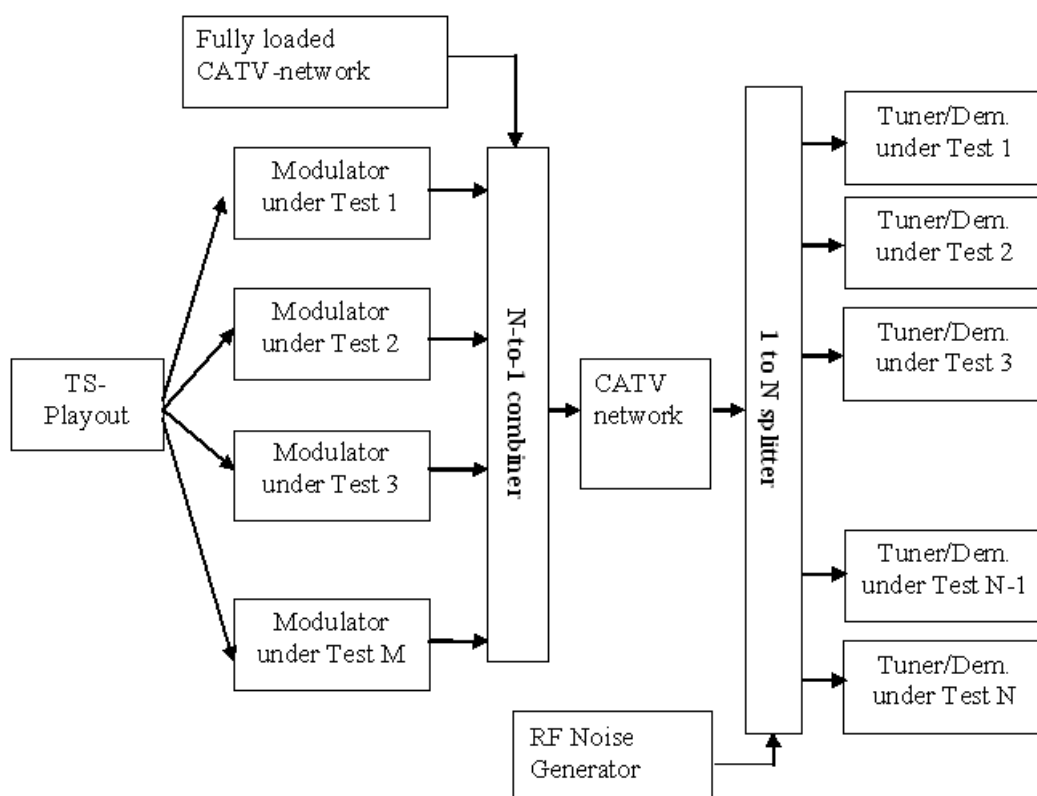
## E.3 DVB-C2 tuner/demodulator performance testing

This test session focused on the overall system performance, taking into account impairments caused by the modulator, the cable network and the demodulator. The test configuration should be a realistic model of the usage of the DVB-C2 system in a real cable environment.

On the transmitter side a fully loaded cable system was used with the channel D562 (562 MHz centre frequency) as the wanted channel and the two adjacent channels (D554 and D570) each carried a DVB-C Signal (DVB-C@256-QAM). The remaining channel capacity was fully filled with a mix of analogue and digital signals.

The basis parameters of the DVB-C2 wanted signal were chosen as given in Annex B of the present document. However, all different coding and modulation parameters of the DVB-C2 system have been tested during this session.

Figure E.1 gives the block diagram of the testing configuration



**Figure E.1: Block diagram of the performance testing configuration**

The used HFC cable network model consisted of an optical transmission system and a typical cascade of amplifiers and realistic coaxial cable segments.

Noise was added at the receiver side at RF level.

CNR was measured with a professional measurement equipment (Spectrum Analyser).

During the DVB-C2 Plug Fest the performance of the different prototype modulators and demodulators were measured for each of the available modulation and FEC code rate configuration of the DVB-C2 system. In a channel configuration with nominal signal levels noise was added step by step at the receiver side and the minimum CNR for error free reception was noted per prototype demodulator. On basis of the theoretical performance characteristics the overall implementation loss of each of the configurations could be calculated.

Such performance figures are not subject to standardization. Any operator of a digital TV platform using the DVB-C2 system has to define his reference chain and allocate a figure for implementation loss to the different components of his reference chain.



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## E.4 DVB-C2 adjacent channel protection ratio measurement

This test session evaluated the performance of the tuners/demodulators under test in presence of adjacent channel interference. For this test only the two neighbouring channels were transmitted in addition to the measurement channel.

The Adjacent Channel protection ratio is the ratio of the maximum signal level of the adjacent signals allowing an error free reception of the wanted channel to the nominal adjacent channel signal level.

All ACI measurements were performed with the same reference modulator.

For the adjacent channels identical modulation schemes were chosen for the lower and the upper adjacent channels.

Measurements started with nominal signal levels. DVB-C2 @1024-QAM was transmitted at the same signal level as used for DVB-C@256-QAM today.

In order to measure the available protection ratio, the signal levels of the adjacent channels were increased up to the maximum signal level (a level which is still appropriate for the tuner input circuits). In case an error free reception was still possible at maximum signal level, the signal level of the wanted channel was decreased in order to increase the amplitude difference between wanted and adjacent channels. Care was taken that the CNR of the wanted channel was still appropriate for error free reception.

Those measurements were performed for DVB-C2 as the wanted signal with all possible configurations of modulation and code rates.

For the adjacent channels the following modulation schemes have been tested:

- DVB-C2, @4096-QAM
- DVB-C2, @1024-QAM
- DVB-C @ 256-QAM
- PAL B/G

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## Annex F (informative): Bibliography

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- ETSI TR 102 376: "Digital Video Broadcasting (DVB) User guidelines for the second generation system for Broadcasting, Interactive Services, News Gathering and other broadband satellite applications (DVB-S2)".
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## History

<b>Document history</b>		
V1.1.1	August 2010	Publication
V1.2.1	June 2011	Publication
V1.3.1	January 2016	Publication