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Technical Specification

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 standardisation, interoperability and future proof specifications.

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 provides offering 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) compared to DVB-C.

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 standard. 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 standard.

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 standards [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 9 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.

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 standard document 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 standard 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 standard 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 invent 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 optimised 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 standard. The present document was prepared by these experts with the intention to provide know-how complementary to the explanations of the standard 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.

2 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 referenced 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.

2.1 Normative references

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

Not applicable.

2.2 Informative references

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: "Digital Video Broadcasting (DVB); Second generation framing structure, channel coding and modulation systems for Broadcasting, Interactive Services, News Gathering and other broadband satellite applications (DVB-S2)".

[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: " Digital Video Broadcasting (DVB);Generic Stream Encapsulation (GSE) Protocol".

[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. B¨orjesson, "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.

NOTE: Available from BBC website: http://www.bbc.co.uk/rd/pubs/papers/pdffiles/jsebu276.pdf.

[i.14] E.R.Berlekamp: "Algebraic Coding Theory", New York:McGraw-Hill, 1968.

[i.15] D. J. MacKay and R. M. Neal: "Near Shannon limit performance of low density parity check codes", Electronics Lett. Mar. 1997, vol. 33, no.6, pp. 457-458.

[i.16] Thomas J. Kolze, "HFC Channel Model Submission", IEEE 802.14a/98-12, May 1998.

[i.17] IEEE 802.14: "Broadband Cable Access Method and Physical Layer Specification".

[i.18] IEC 60728-1 (May 2008): "Cable networks for television signals, sound signals and interactive services - Part 1: System performance of forward paths".

[i.19] "A new cable frequency plan and power deployment rules", ReDeSign Deliverable D14, 2009.

NOTE: Available at: www.ict-redesign.eu.

[i.20] "The HFC channel model", ReDeSign Deliverable D8, 2008.

NOTE: Available at: 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.

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 Symbol.

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

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

div: integer division operator, defined as:

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

dummy cell: OFDM Cell carrying a pseudo-random value used to fill the remaining capacity not used for L1 signalling, PLPs or Auxiliary Data

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_{LDPC} (16 200 or 64 800) bits of one LDPC encoding operation

NOTE: In case of Data Slices 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.

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-part 2: 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-part 2 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:

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

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

nn_D: 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 Ts 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 Slices. Transmission parameters may change each XFECFrame.

PLP bundling: transmission of one PLP via multiple Data Slices

PLP_ID: this 8-bit field identifies uniquely a PLP within a C2 transmission signal

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" shall be set to "0" unless otherwise defined

START_FREQUENCY: Index of lowest used OFDM subcarrier of a C2 System. The value of START_FREQUENCY shall be a multiple of D_X

***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 [i.1] and the following apply:

$\oplus$	exclusive OR / modulo-2 addition operation
Δ	Absolute guard interval duration
Λ	LDPC codeword of size N_{ldpc}
λ_i	LDPC codeword bits
λ^{RM}	32 output bits of Reed-Muller encoder
λ_i^{RM}	Bit number of index *i* of 32 bit long output bits of Reed-Muller encoder
η_{MOD}, $\eta_{MOD}(i)$	Number of transmitted bits per constellation symbol (for PLP *i*)
π_{p}	Permutation operator defining parity bit groups to be punctured for L1 signalling
π_s	Permutation operator defining bit-groups to be padded for L1 signalling
$A_{m,l}$	Output vector of the frequency interleaver of OFDM Symbol *l* and C2 Frame *m*
A_{CP}	Amplitude of the continual pilot cells
A_{SP}	Amplitude of the scattered pilot cells
$a_{m,l,q}$	Frequency-Interleaved cell value, cell index *q* of symbol *l* of C2 Frame *m*
$B(n)$	Location of the first Data Cell of symbol l allocated to Data Slice *n* in the frequency interleaver
b	16 bit long FECFrame 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 FECFrame 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 C2 Frame *m*
$dB\mu V$	relative logarithmic signal level with reference to 1µV
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 ($0 \leq e < N_{substreams}$), 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

$g_1(x), g_2(x), \ldots, g_{12}(x)$	Polynomials to obtain BCH code generator polynomial
g_q	Complex cell of index q of a Data Slice Packet
$H(q)$	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_{bch}	Number of bits of BCH uncoded Block
K_{i}	L1 signalling part 2 parameter selected as $\mathrm{N}_{L1part\ 2}(\mathrm{K_i}) <= N_{L1part\ 2\_Cells} \times \eta_{\mathrm{MOD}}$
K_{ldpc}	Number of bits of LDPC uncoded Block
$K_{L1\_PADDING}$	Length of L1_PADDING field
$K_{L1part\ 2}$	Length of L1-part 2 signalling field including the padding field
$K_{L1part\ 2\_ex\_pad}$	Number of information bits in L1-part 2 signalling excluding the padding field.
$K_{N,min}$	Lowest frequency carrier index of a frequency Notch
$K_{N,max}$	Highest frequency carrier index of a frequency Notch
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 OFDM Symbols per C2 Frame (excluding Preamble)
L_{F}	Number of OFDM Symbols per C2 Frame including excluding preamble
L_P	Number of preamble OFDM Symbols within the C2 Frame
l	Index of OFDM Symbol within the C2 Frame (excluding preamble)
l_P	Index of preamble OFDM Symbol in C2 Frame
m	C2 Frame 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_{data}	Number of Data Cells in a Data Slice in frequency interleaver
N_{DP}	Number of complex cells per Data Slice Packet
N_{group}	Number of bit-groups for BCH shortening
$N_{L1part\ 2}$	Length of punctured and shortened LDPC codeword for L1-part 2 signalling
$N_{L1part\ 2\_Cells}$	Number of available cells for L1 signalling part 2 in one OFDM Symbol
$N_{L1part\ 2\_FEC\_Block}$	Number of LDPC blocks for the L1 signalling part 2
$N_{L1part\ 2\_max\_per\_Symbol}$	Maximum number of L1 information bits for transmitting the encoded L1 signalling part 2 through one OFDM Symbol
$N_{L1\_TI\_Depth}$	Time interleaving depth for L1 signalling part 2
$N_{L1part\ 2\_temp}$	Intermediate value used in L1 puncturing calculation
N_{ldpc}	Number of bits of LDPC coded Block
$N_{MOD\_per\_Block}$	Number of modulated cells per FEC block for the L1-part 2 signalling
N_{MOD_Total}	Total number of modulated cells for the L1-part 2 signalling
N_{pad}	Number of BCH bit-groups in which all bits will be padded for L1-part 2 signalling
Nplp	*Number of PLPs in a C2-system*
N_{punc}	Number of LDPC parity bits to be punctured
N_{punc_groups}	Number of parity groups in which all parity bits are punctured for L1 signalling
N_{punc_temp}	Intermediate value used in L1 puncturing calculation
N_r	Number of bits in Frequency Interleaver sequence
N_r	Number of rows of bit or time interleaver

N_{RT}	Number of reserved carriers
$N_{substreams}$	Number of substreams produced by the bit-to-sub-stream demultiplexer
n	Data slice number
$P_k(f)$	Power spectral density
p_i	LDPC parity bits
Q_{ldpc}	Code-rate dependent LDPC constant
q	Data Cell index within the OFDM Symbol prior to frequency interleaving and pilot insertion
$R_{eff\_16K\_LDPC\_1\_2}$	Effective code rate of 16K LDPC with nominal rate ½
$R_{eff\_L1part\ 2}$	Effective code rate of L1-part 2 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
S_0	List of reserved carriers
T	Elementary period
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 C2 Frame
T_P	Time interleaving period
T_S	Total OFDM Symbol duration
T_U	Useful OFDM Symbol 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 FECFrame header encoding
$u_{(i+2)mod32}{}^{RM}$	Output of the cyclic delay block for input bit *i* in the FECFrame header 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* (= Data slice start carrier) of the current OFDM Symbol *l* and C2 Frame *m*
v^{RM}	Scrambled output sequence in the lower branch of the FECFrame header encoder
$\mathrm{v}_i{}^{RM}$	Bit *i* of scrambled output sequence in the lower branch of the FECFrame header encoder
w_i	Bit *i* of the data scrambling sequence
w^{RM}	32 bit scrambling sequence in the lower branch of the FECFrame header encoder
$\mathrm{w}_i{}^{RM}$	Bit *i* of scrambling sequence in the lower branch of the FECFrame header encoder
w^{p}	Pilot synchronization sequence, build out of w_i and w'
$\mathrm{w}_k{}^{\mathrm{p}}$	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 Data Cells of the OFDM Symbol *l* and the C2 Frame *m*
x	Address of the parity bit accumulator according to i_{360} in LDPC encoder
$\mathrm{y}_{i,q}$	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 [i.1] and the following apply:

1024-QAM	1024-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
ACM	Adaptive Coding and Modulation
ATM	Asynchronous Transfer Mode
AWGN	Additive White Gaussian Noise
BB	BaseBand
BBFrame	BaseBand Frame
BCH	Bose-Chaudhuri-Hocquenghem multiple error correction binary block code
BUFSTAT	Actual status of the receiver buffer
C/N	Carrier-to-Noise ratio
C/N+I	Carrier-to-noise and intermodulation power ratio
CATV	Community Antenna Television
CCM	Constant Coding and Modulation
CINR	Carrier to Intermodulation Noise Ratio
CIR	Carrier to Intermodulation Ratio
CNR	Carrier to Noise Ratio
COFDM	Coded Orthogonal Frequency Division Multiplex
CP	Continual Pilot
CPE	Common Phase Error
CPE	Customer Premises Equipment
CRC	Cyclic Redundancy Check
CSO	Composite Second Order
CTB	Composite Triple Beat
CTO	Composite Second Order
D	Decimal notation
dB	decibel
DBPSK	Differential Binary Phase Shift Keying
DFL	Data Field Length
DFT	Discrete Fourier Transformation
DNP	Deleted Null Packets
DVB	Digital Video Broadcasting project
DVB-C	DVB System for cable transmission

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

DVB-C2 DVB-C2 System

NOTE: As specified in the present document.

DVB-S DVB System for digital broadcasting via satellites

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

DVB-S2 Second Generation DVB System for satellite broadcasting

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

DVB-T DVB System for Terrestrial broadcasting

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

DVB-T2	Second Generation DVB System for Terrestrial broadcasting

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

DVB-X2	Generic abbreviation for the group of 2nd generation DVB systems: DVB-C2, DVB-S2 and DVB-T2
EBU	European Broadcasting Union
EIT	Event Information Table (DVB SI Table)
EMM	Entitlement Management Message
FEC	Forward Error Correction
FFT	Fast Fourier Transformation
FIFO	First In First Out
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
IDTV	Integrated Digital TV
IF	Intermediate Frequency
IFFT	Inverse Fast Fourier Transform
IM	InterModulation
IPv4	Internet Protocol version 4
IPv6	Internet Protocol version 6
IS	Interactive Services
ISCR	Input Stream Time Reference
ISI	Input Stream Identifier
ISSY	Input Stream SYnchronizer
Kbit	2^{10} = 1 024 bits
LDPC	Low Density Parity Check (codes)
LSB	Least Significant Bit
LTE	Long Term Evolution
LUT	LookUp-Table
Mbit	2^{20} = 1 048 576 bits
Mbit/s	Mbit per second
Mega	1.000.000 (1 Million)
MBaud	1.000.000 Baud = 1.000.000 Symbols per second

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

MPE	Multi Protocol Encapsulation
MPEG	Moving Pictures Experts Group
MPS	Monic Polynomial Sequence
MPTS	Multi Program Transport Stream
MSB	Most Significant Bit

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

NA	Not Applicable
NIT	Network Information Table
NM	Normal Mode
NPD	Null Packet Deletion
OFDM	Orthogonal Frequency Division Multiplex
OSI	Open Systems Interconnection Reference Model
PAPR	Peak to Average Power Ratio
PER	(MPEG TS) Packet Error Rate
PDF	Probability Density Function
PLP	Physical Layer Pipe
PRBS	Pseudo Random Binary Sequence

PSD	Power Spectral Density
PSI/SI	Program Specific Information/Service Information
PVR	Personal Video Recorder
QAM	Quadrature Amplitude Modulation
QoS	Quality of Service
QPSK	Quaternary Phase Shift Keying
RF	Radio Frequency
SD	Standard Definition
SDTV	Standard Definition TV
SNR	Signal to Noise Ratio
SP	Scattered Pilot
SPTS	Single Program Transport Stream
STB	Set Top Box
TDM	Time Division Multiplex
TF	Time/Frequency
TI	Time Interleaver
TS	Transport Stream
TSPSS	Transport Stream Partial Stream Synchronized
UP	User Packet
UPL	User Packet Length
VBR	Variable Bit Rate
VCM	Variable Coding and Modulation
VOD	Video On Demand
XFECFrame	XFEC Frame

4 Overview of DVB-C2

4.1 DVB-C2 commercial requirements

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, 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 1024-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 max 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, Davic) and the associated signal backoff 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 must 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 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 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 systems currently using DVB-C for the EuroDOCSIS 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) 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 present DVB-C2 standard.

Although the DVB 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 optimise 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 Slices 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 DVB-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 µs).

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 optimised 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 mayor 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	1024-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 (4096-QAM)
8 MHz channel	50,9 Mbit/s	79,39 Mbit/s
16 MHz channel	50,9 Mbit/s	81,40 Mbit/s, 2,46 % (see note)
24 MHz channel	50,9 Mbit/s	82,08 Mbit/s, 3,41 % (see note)
32 MHz channel	50,9 Mbit/s	82,42 Mbit/s, 3,84 % (see note)
64 MHz channel	50,9 Mbit/s	82,93 Mbit/s, 4,49 % (see note)
NOTE: Relative gain with reference to 4096-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 a 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

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 Slices (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 Slices, prepends a preamble with the Level 1 signalling information, and forwards the resulting C2 frame to the OFDM generation stage.

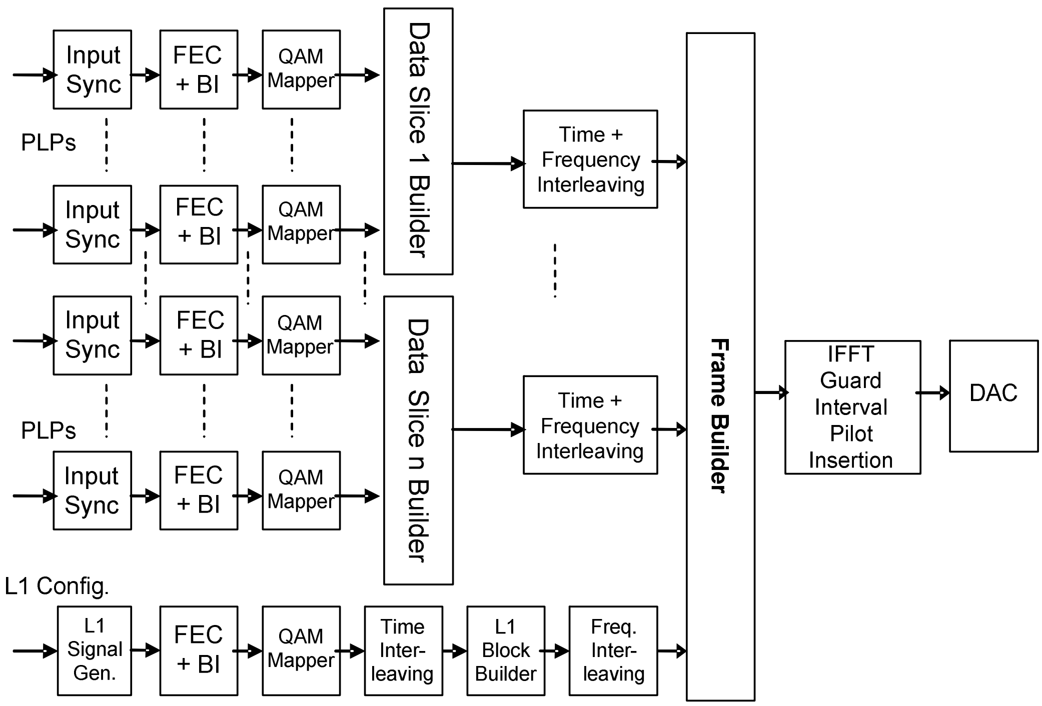

Figure 1: High-level block diagram of the signal processing defined for the DVB-C2 transmitting end

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 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 became possible in the recent years due to the progress in semiconductor manufacturing. The impressing 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 especially reached 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 1024-QAM and 4096-QAM. The possible combinations of modulation and coding schemes are given in table 1. 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	1024-QAM	4096-QAM
2/3	-	13,5 dB	-	-	-
¾	-	-	20,0 dB	24,8 dB	-
4/5	10,7 dB	16,1` 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

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 and refined and extended the parameters substantially in DVB-T2. 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.

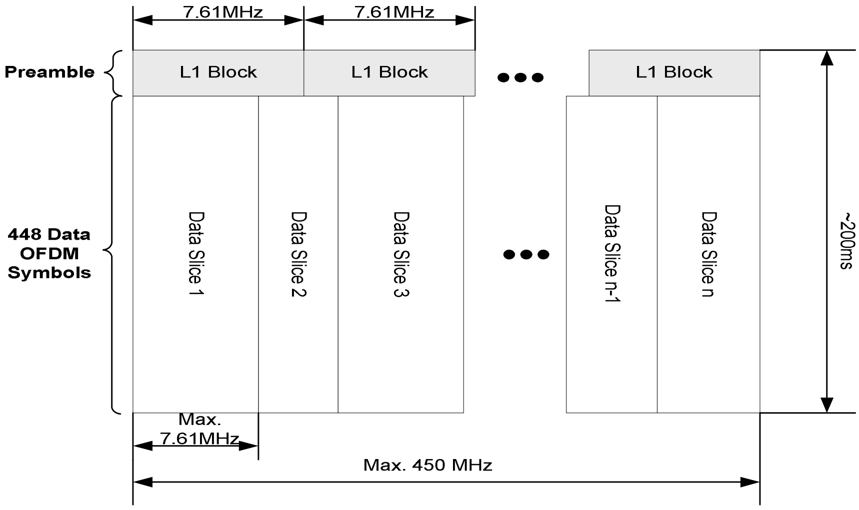

Figure 2: Time frequency diagram of a DVB-C2 frame

DVB-C2 reuses the OFDM subcarrier spacing of the 4K FFT mode from DVB-T2, i.e. the useful OFDM symbol duration is 448 µs. 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 a 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 Slices 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 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.

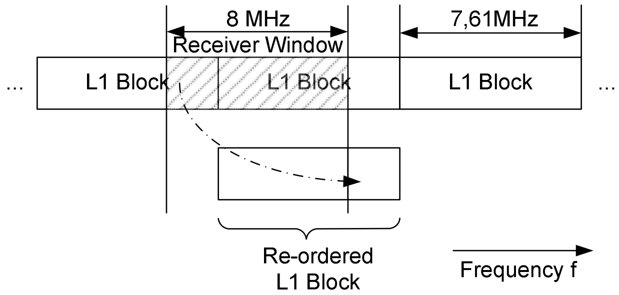

Figure 3: DVB-C2 preamble structure, re-ordering of the OFDM sub-carriers in the L1 Blocks to extract the L1 signalling information

In contrast to the L1 part 2 blocks, the Data Slices 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 part 2 signalling data at any possible tuning position. The only requirement for the Data Slices is that each Data Slice must 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 Slices can be accumulated until the overall number of OFDM sub-carriers of the C2 signal is reached. Both position and bandwidth of the Data Slices may vary between different DVB-C2 frames as this does not require any re-tuning of the receiver. The signalling inside the L1 part 2 Blocks does not only contain the start and the end frequency of the Data Slices, 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

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 part 1) 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 part 2) 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 7.5

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.

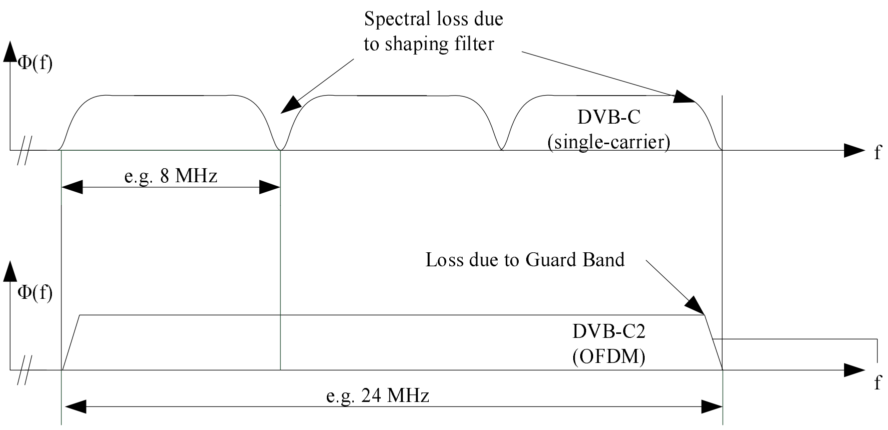

Figure 4: Occupation of frequency spectrum by DVB-C today (upper diagram) and by DVB-C2 in the future (lower diagram)

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 Slices of one OFDM signal, i.e. Data Slices 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 Slices of 6,4 MHz) is only 3,25 %, while it is 15 % for DVB-C.

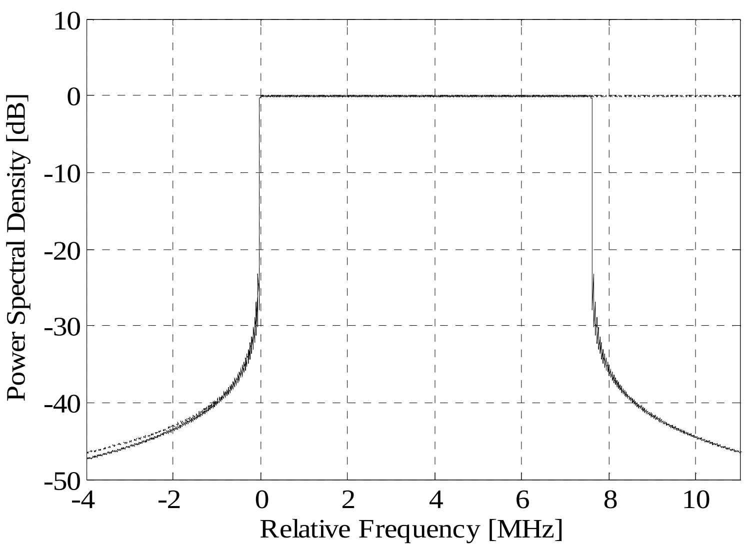

Figure 5: Ideal power spectral density of DVB-C2 signal with Guard Interval 1/128 for 7,61 MHz and 450 MHz

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.

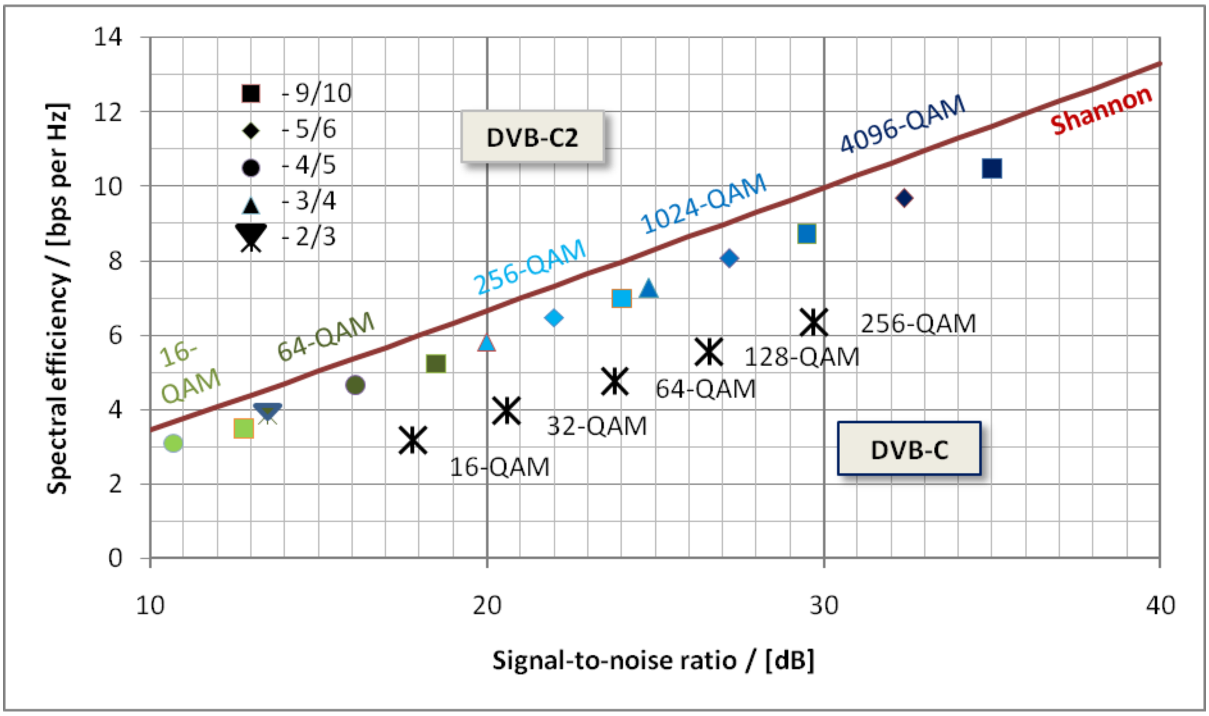

Figure 6: Overall spectral efficiency of DVB-C and DVB-C2 for different code rates and modulation schemes (DVB-C2 parameters: 32 MHz signal bandwidth, Guard Interval 1/128, pilot density 1/96)

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), "-" means not applicable

	16-QAM	64-QAM	256-QAM	1024-QAM	4096-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	-	-	-
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

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

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.

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 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 C2-System is a certain multiplexing configuration of PLPs and Data Slices which form together with the preamble DVB-C2 frames which are transmitted as a C2-System 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

Data Slices contain one or more PLPs and are transmitted with identical time and frequency interleaving parameters. More details about Data Slices 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

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 Slices 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 [i.1].

5.6.1.6 BB-Frames

BB-Frames are the basic unit in the logical framing structure of DVB-C2: allocation and scheduling are performed in whole numbers of BB-Frames. Where packetized streams are being carried, the packets may be mapped to BB-frames either synchronously or asynchronously, i.e. each BB-Frame may contain a whole number of packets, or packets may be fragmented across two BB-Frames. BB-Frames contain a header, including the packet length and position of the first packet, which allows the original packets to be reconstructed at a receiver. BB-Frames may contain padding, in case insufficient data is available for a whole BB-frame, or in case it is desired not to avoid packet fragmentation. The total size of each BB-Frame, 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 BB-Frame header (see clause 5.1.6 in [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 a 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 x 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 part 2 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

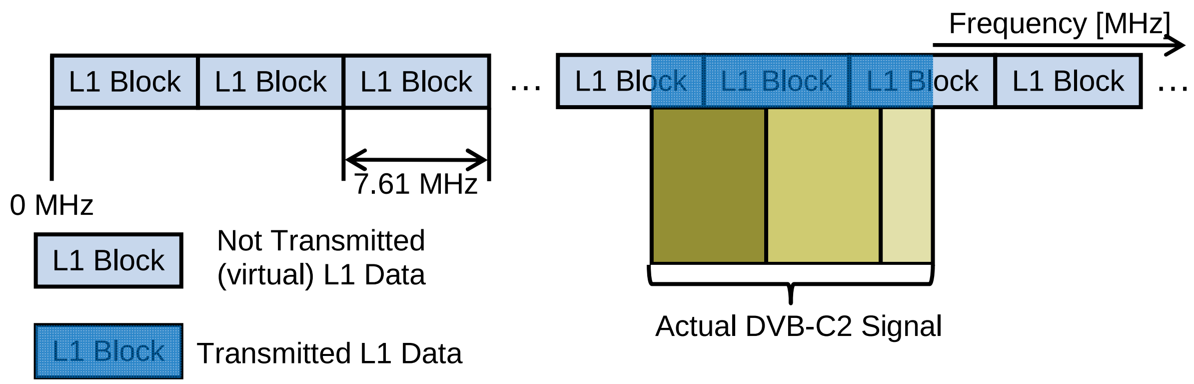

Figure 7: Absolute OFDM concept in DVB-C2

The concept of 'absolute OFDM' is unique to DVB-C2. The first L1 part 2 signalling block begins at the absolute frequency of 0 MHz and the further blocks are partitioned in steps of 7,61MHz 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 part 2 block partitioning and the related pilot sequences are defined for the whole cable spectrum - of course L1 blocks are only transmitted in these frequencies where Data Slices are present.

6 Choice of Basic Parameters

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, 1024-QAM and 4096-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 standardisation process non-rectangular QAM constellations were also considered but the performance benefits were not considered to be sufficient to justify inclusion in the standard.

Use of the LDPC codec first used in DVB-S2 and more recently DVB-T2 maximises 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.

AT least 1024-QAM modulation scheme is 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. 4096-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 4096-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 2dB granularity in SNR between the limits of 10 dB (16-QAM) and 35 dB (4096-QAM) would provide sufficient options (table 1) for configuration and optimisation 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 1). 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.

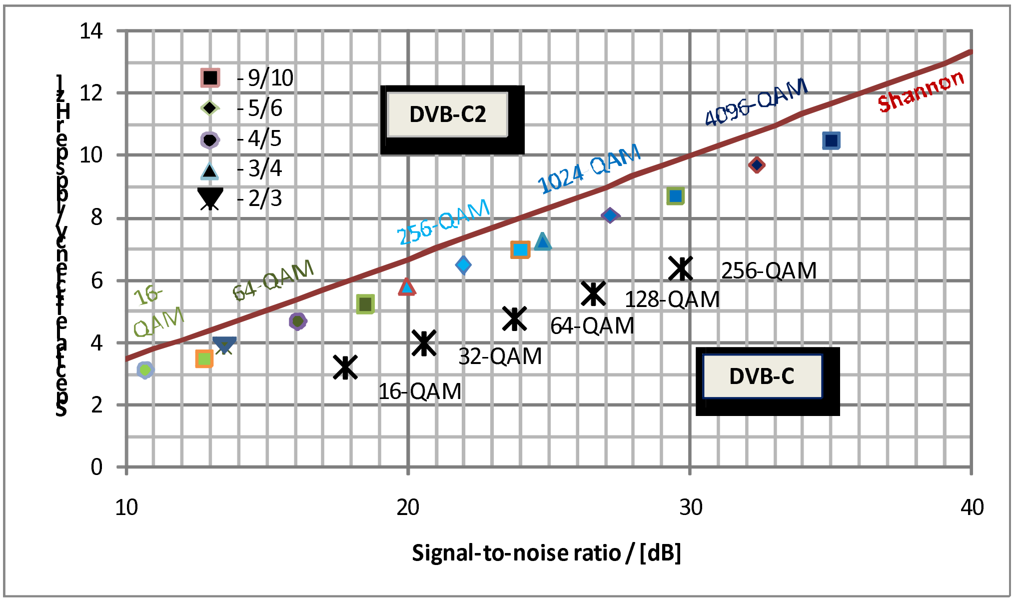

Figure 8: DVB-C2 Modulation Levels, Code Rates and Spectral Efficiency compared with Theoretical Shannon Limit and DVB-C

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)
4096-QAM	9/10	10,8	34,97	-
4096-QAM	5/6	10,0	32,36	2,6
1024-QAM	9/10	9,0	29,50	2,8
1024-QAM	5/6	8,33	27,15	2,3
1024-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 (CPE) 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 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 448µs. 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*3 584 + 3 409 = 14 161 subcarriers. In this case a 16 k IFFT could be used for signal generation on transmitter side.

There are no alternative carrier spacings used in DVB-C2. Fundamental features of C2 are the provision of Data Slices with their flexible number of carriers providing sufficient flexibility for configuring networks in practical applications of the standard. 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). On a 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 part 2 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 Δ 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 Δ, 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 µs 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 minimises impact on spectral efficiency and thereby maximises 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 (µs)	Carrier Spacing (kHz)	Echo Tolerance (µs)	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 synchronisation 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 must be sufficiently dense that they can follow channel variations as a function of both frequency (carrier index) and time.

The scattered pilot pattern density must fulfil the sampling theorem in time and frequency direction. This can be summarised 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 equalisation in a DVB-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 (Dy) should be as low as possible to avoid increasing memory requirements in the receiver. Dy 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 summarised 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
$1/D_XD_Y$	2,08 %	1,04 %	SP overhead
$1/D_X$	1/12	1/24	$T_{Nyquist}/T_U$, for f-&-t interpolation
$1/D_XD_Y$	1/48	1/96	$T_{Nyquist}/T_U$, for f-only interpolation
$1/(2D_Y)$	0,125	0,125	$f_{NYQUIST}/f_S$, for f-&-t interpolation (±)
	0,5	0,5	$f_{NYQUIST}/f_S$, for f-only interpolation (±)

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 Dx is the separation of pilot bearing carriers and Dy is the number of symbols forming one scattered pilot sequence. In DVB-T2 many of the SP patterns are optimised 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 part 2 signalling bits permitted in a C2 frame is 32 766 [i.1], clause 8.3;
- the maximum number of L1 signalling bits in an OFDM symbol is 4 759 [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 Slices;
- 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 (min.)	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 (min.)	0,22	449	204,3
448	1/64	203,8	00, 01	7 (max.)	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 Slices 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 minimised 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 part 2 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 part 2 (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 Slices.

Time interleaving can be applied differently to Data Slices and to L1 part 2 signalling in the preamble. If time interleaving is applied to Data Slices it should also be applied to the L1 part 2 signalling to ensure the robustness of the signalling exceeds that of the Data Slices. The time interleaving depths available for the Data Slices are summarized in table 11.

Table 11: Data Slice/L1 part 2 Time Interleaving Parameters

DSLICE_TI_DEPTH	DSLICE TI Depth (OFDM symbols)	L1_TI_MODE	L1 part 2 TI Depth (OFDM symbols)	Overhead (OFDM symbols)
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

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

- Normal Mode (NM) or High Efficiency Mode (HEM).
- Use of Input Stream Synchronisation (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 synchronise packets from different Data Slices 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 optimise system performance. This can be by increasing the total throughput or minimising 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 DV-C2 specification [i.1] states in annex 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 must 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:

- 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 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 Slices 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 Slices, 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, 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 maximise robustness. This defines upper limits on the number of bits available for signalling.

6.9 Number of Data Slices vs. PLPs

The cable operator needs to consider how to segment the required downstream services into and across Data Slices and PLPs to maximise 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 Slices.
- A PLP may carry one or multiple services.
- PLP bundling enables a single PLP to be carried over several Data Slices - 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 Slices for a C2 system have an upper limit determined by maximum number of bits (32 766) available in L1-part 2 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 Slices may influence the performance of the C2 system, especially if narrow bandwidth Data Slices 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 C2 System allows spreading data of a single PLP connection over different Data Slices (PLP Bundling, see annex F in [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

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 the standard 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 summarised in table 12.

Table 12: Notching Characteristics in DVB-C2

	Narrowband notches	**Broadband notches**
Bandwidth	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
Notch position within a Data Slice	Yes	No
Signaled in L1	Yes	Yes
Receiver handling	Notch corrected by L1 FEC	Notch not inside tuning bandwidth

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 must 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. Not more than one narrowband notch must be located in a Data Slice although more than one narrowband notch could exist in a standard 8 MHz receiver tuning window if multiple Data Slices are defined to exist within the bandwidth of the L1 signalling (7,61 MHz).

Broadband notches (notches with bandwidth above 53 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 part 2 signalling field (6 143 subcarriers).

In general the notching features available in C2 should only be used to mitigate the effects of possible interference scenarios and maximise 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 minimise 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.

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.

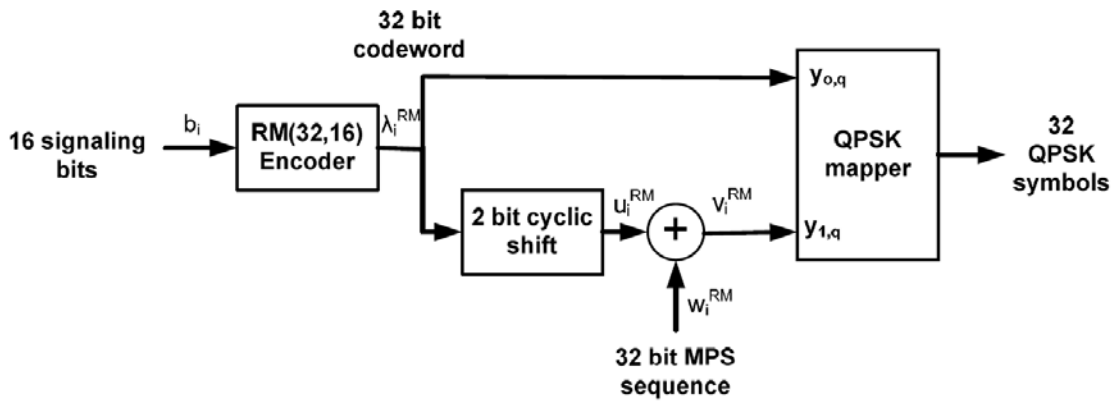

Figure 9(a): Robust FECFrame header

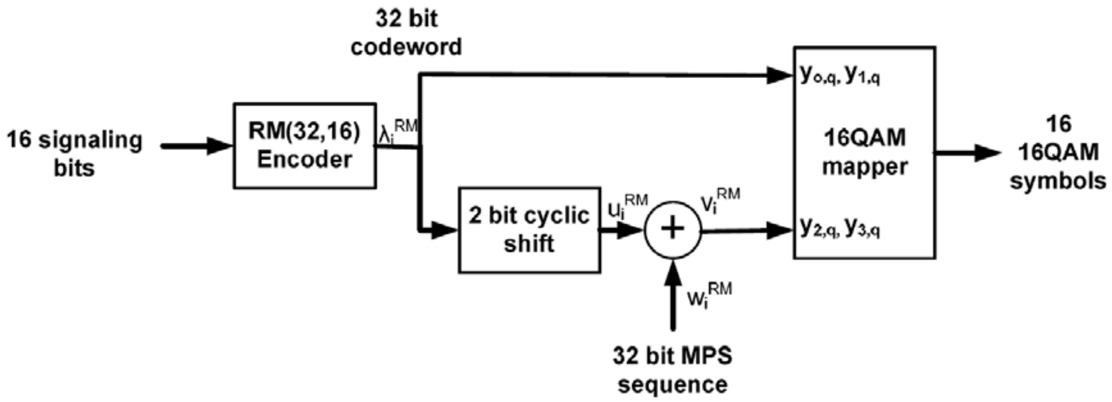

Figure 9(b): High efficiency FECFrame header

The 32 Reed-Muller [i.14] encoded data bits λ_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

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

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.

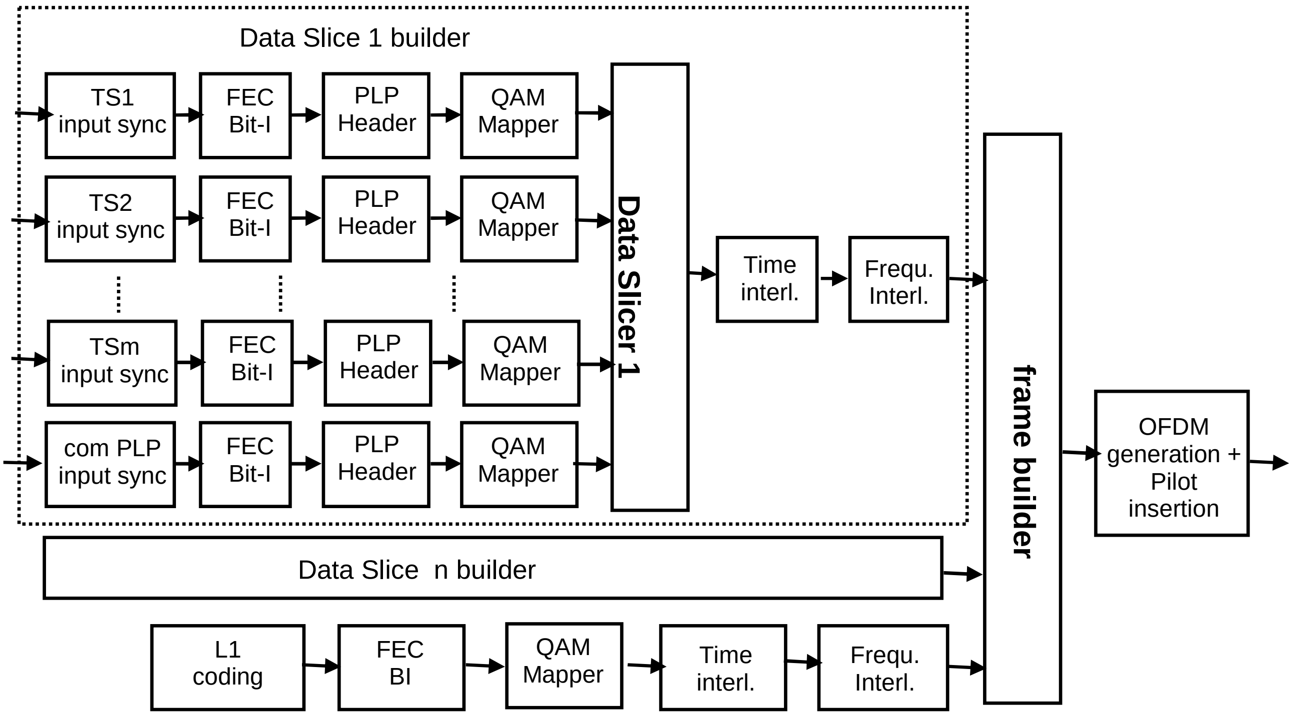

Figure 10: DVB-C2 modulation including common PLP concept

A cable playout center generating DVB TSs uses statistical multiplexing in order to optimize the overall spectrum efficiency. The capacity for CA-messages data and SI data have 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 headend (e.g. several Astra-type DTH-signals are locally multiplexed into one cable channel), or to centralize headend where premium tier HDTV services and basis tier SDTV service which are required to be transmitted in the same cable channel so to ensure different levels of robustness required 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 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 [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 [i.1].

7.3 PLP bundling

In case that there is a 200 Mbit/s input signal and the DVB-C2 signal should be receivable by 8 MHz (6 MHz for the rescaled carrier spacing) receivers, the input signal has to distributed into three 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 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. A Common PLP has to be transmitted twice in order to allow every relevant receiving window to access the common PLP data.

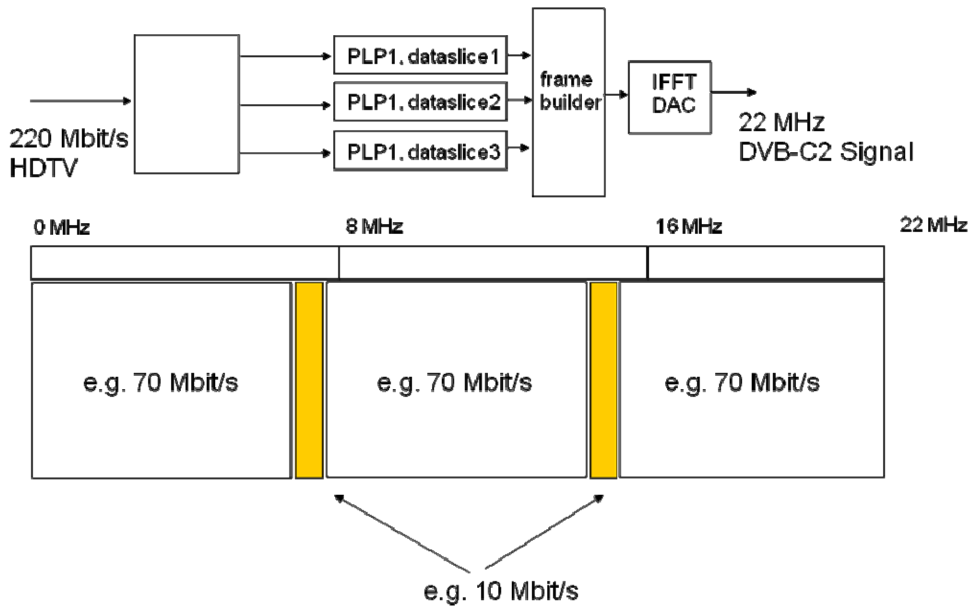

Figure 11: PLP bundling use case example

PLP bundling is signaled 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 reorder packets received from multiple PLPs bundled.

See annex F in [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 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 must not be interfered due to the presence of bundled PLPs in a C2-system.

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.

7.4 Stuffing Mechanism

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.

7.4.1 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 [i.10].

7.4.2 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.

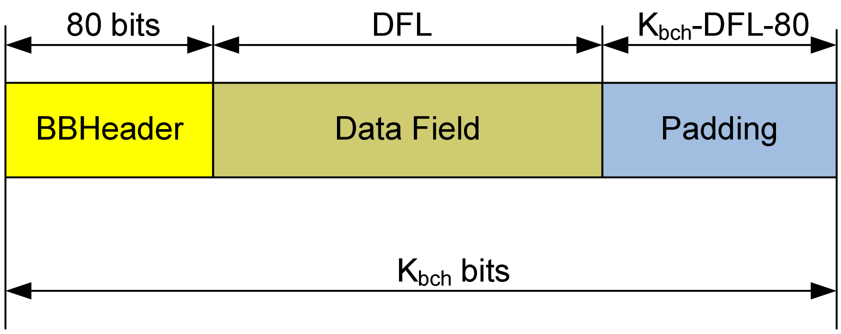

Figure 12: Structure of BBFrame with padding field

Figure 12 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.

7.4.3 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 [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).

7.4.4 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.

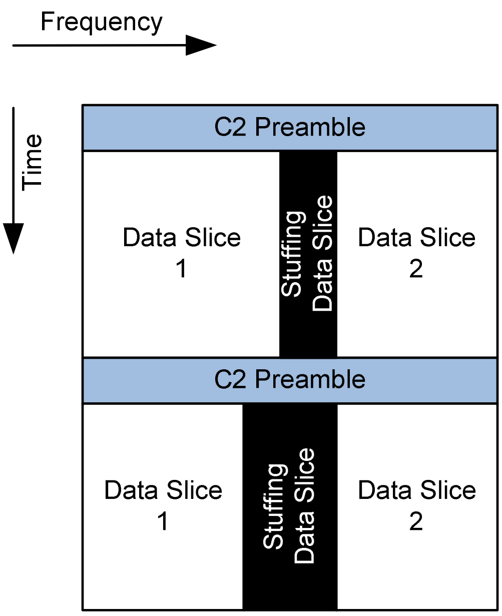

Figure 13: Principle of Data Slice Stuffing

The principle of Data Slice Stuffing is depicted in figure 13. The two (payload) Data Slices do not fill the complete bandwidth of the C2 system completely. Hence, a Stuffing Data Slices 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 avoid high peaks of the OFDM time domain signal.

7.5 Multiplexing, Dimensioning of PLPs and Data Slices

The following clauses list possible scheduling algorithms. These 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.

7.5.1 Single PLP per Data Slice

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.

7.5.1.1 Data Slice Type 1

Figure 14 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.

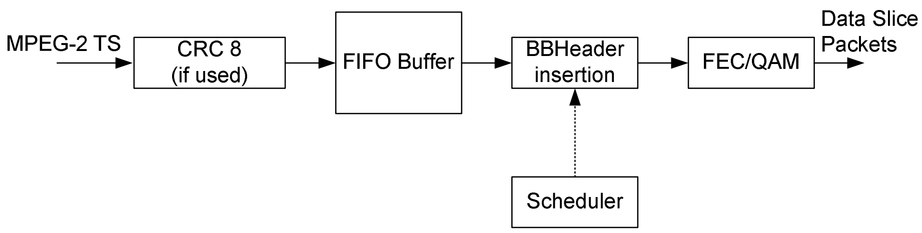

Figure 14: Block Diagram for Data Slice Type 1 Scheduler

7.5.1.2 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.

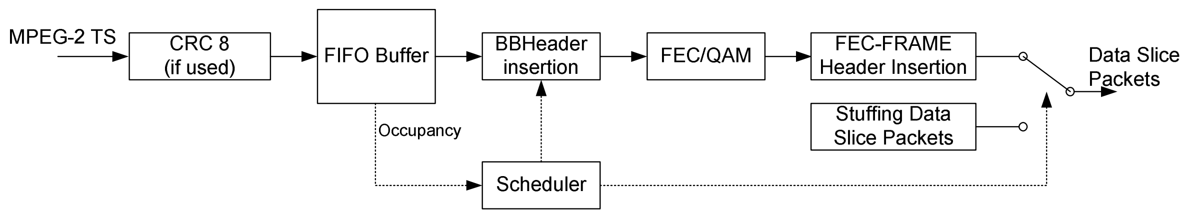

Figure 15: Block Diagram for Data Slice Type 2 Scheduler

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.

7.5.2 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 16 shows the block diagram of a possible scheduling algorithm. The associated pseudo code for the scheduling algorithm is listed in figure 17.

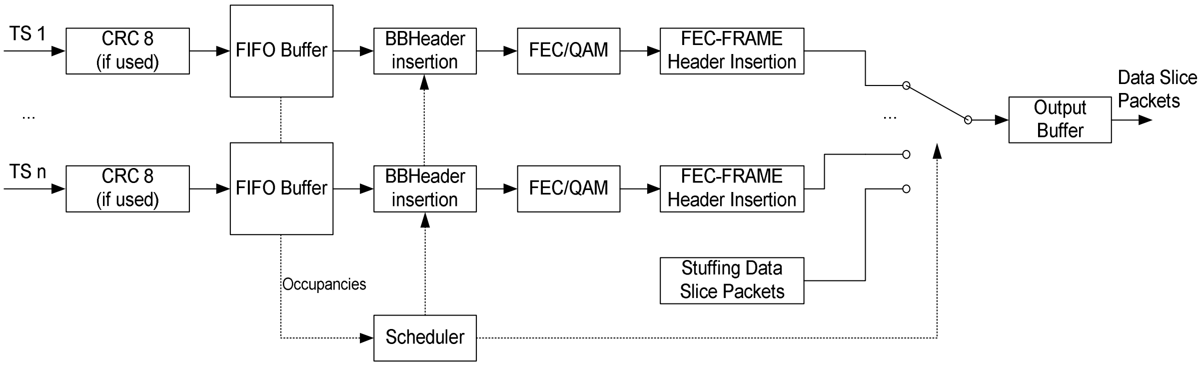

Figure 16: Block Diagram for Multiple PLP per Data Slice Scheduling

```
while (true) {                         // Loop forever

   for (inputFIFO = 1; inputFIFO <= numberOfPLPs; inputFIFO++) {
      if (hasBBFrame(inputFIFO)) {     // Transmit one BBFRAME of
                                       // this PLP if avialble in FIFO
         transmitBBFrame(inputFIFO);
      }
   }
   if (outputBufferEmpty())            // Transmit Stuffing if buffer empty
      transmitStuffing();

   while (!outputBufferEmpty());       // Wait until data transmitted
}
```

Figure 17: 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.

7.6 Layer-2 signalling

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 [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.

7.6.1 Transport Streams

For the Transport-Stream case a single new PSI/SI signalling element is used: the C2 delivery-system descriptor (C2dsd): see figure 18. 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, see figure 19.

The C2dsd maps a Transport Stream being signalled with the NIT, and heading the TS descriptor loop, to the corresponding C2 system (C2_system_id) and the PLP (PLP_id) that carries this Transport Stream within the C2 system. This mapping is handled by the upper part (blue part of figure 18) of the C2dsd. Each TS carried requires a separate instance of the C2 delivery-system descriptor, i.e. another loop cycle of the NIT's TS loop.

The optional lower part of C2dsd (which appears only in the C2dsd long version) describes the physical parameters of the C2 system (orange part). Since these parameters don't change within the same C2 system, the lower part, i.e. the C2 system parameters, (orange) occurs only once per C2 system. This means that a receiver needs to check the instances of the C2dsd until it finds the long version (if present at all) in order to get the information about the C2 system parameters, see figure 19. The presence of the C2 system parameters (orange) is optional.

```
C2_delivery_system_descriptor(){
   descriptor_tag                          8  uimsbf
   descriptor_length                       8  uimsbf
   descriptor_tag_extension                8  uimsbf
   plp_id                                  8  uimsbf
   C2_system_id                            16  uimsbf
   -------------------------------------------------------------
   if (descriptor_length > 4){
      C2_system_tuning_frequency           32  bslbf
      active_OFDM_symbol_duration          3  bslbf
      guard_interval                       3  bslbf
      reserved                             2  bslbf
   }
}
```

Mapping of TS to PLP and C2 system

C2 system parameters (apart from C2_system_id)

Figure 18: C2 delivery system descriptor

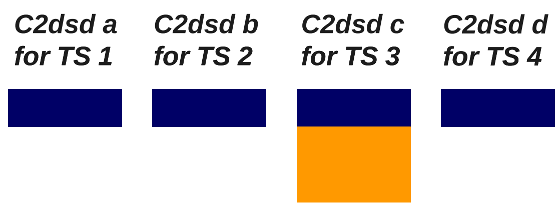

Figure 19: Instances of C2 delivery system descriptor (within same NIT), example

The physical parameters provided for the C2 system (orange) reflect a full description of the signal on air, including its centre frequency. One 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 [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 [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 an other medium. In this case, layer-2 signalling is not valid any more 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.

7.6.2 Generic Streams.

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

a) Generic Fixed Packet size Stream (GFPS).

b) Generic Continuous Stream (GCS), also covering the case of a Generic Variable Packet size Stream.

c) Generic Stream Encapsulated (GSE) according to [i.9].

For cases a) and b) there is no layer 2 signalling definition available, for c) refer to clause 8 in [i.7].

8 Modulator

8.1 Preamble Generation

Figure 20 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.

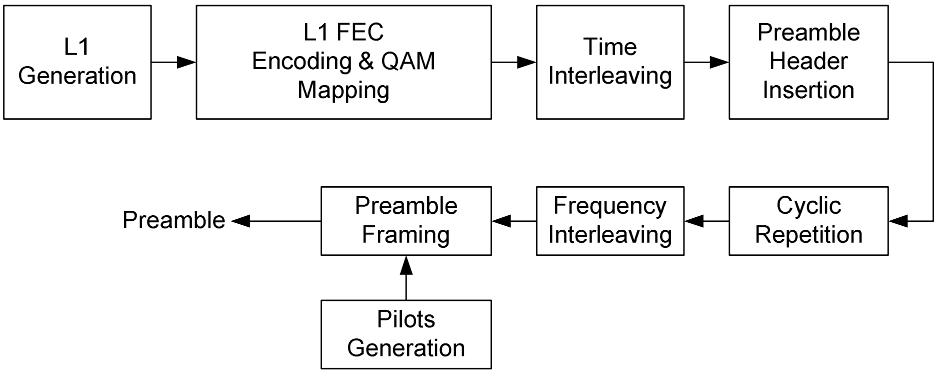

Figure 20: DVB-C2 preamble generation block diagram

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 [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 part 2 data bits and constellation size. Each L1 TI block contains integer number of shortened/punctured FEC blocks of L1 part 2 data. Each L1 TI block should contain all cells for encoded and modulated L1 part 2 data; the cells for L1 part 2 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 [i.1]).

The address generation equation in clause 8.5 of [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 time interleaving gain and preamble overhead can be traded-off through four L1 TI modes (see table 17 of [i.1]). The overheads of "no time interleaving" and "best-fit" mode are 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 part 2 data is relatively small and the number of OFDM symbols for preamble is just one or two for example, the robustness of L1 part 2 data may be greatly damaged by 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 part 2 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 [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 [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 shall be used (see clause 9.4.5), which works on the N_{L1}=2 840 data cells of each L1 Block.

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

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

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

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 from this symbol 0 is described in clause 8.5.2.

8.1.2 Preamble Pilot Generation

The DVB-C2 preamble repeats itself cyclically in the frequency domain. As repetitions wihtin the frequency domain lead to high peak-to-average power ratios it is desireable to avoid this direct repetion. For the DVB-C2 preamble this aim is reached by means of scrambling the preamble with two scrambling sequences. The data scrambling sequence hereby 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 21 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:

$$X^{11} + X^2 + 1 \qquad (2)$$

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

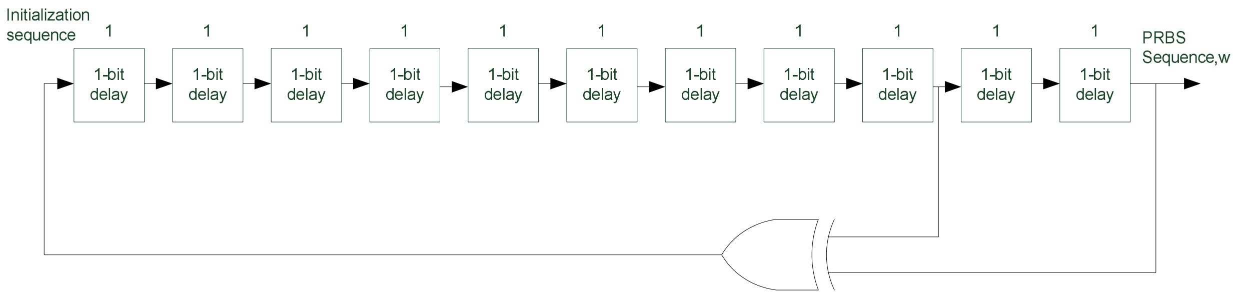

Figure 21: Data Scrambling Sequence

The first 100 output values w_0, w_1, …,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, 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, 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:

$$w_k = w_{(k \bmod 2047)} \tag{3}$$

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 22 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:

$$X^{11}+X^{2}+1 \tag{4}$$

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

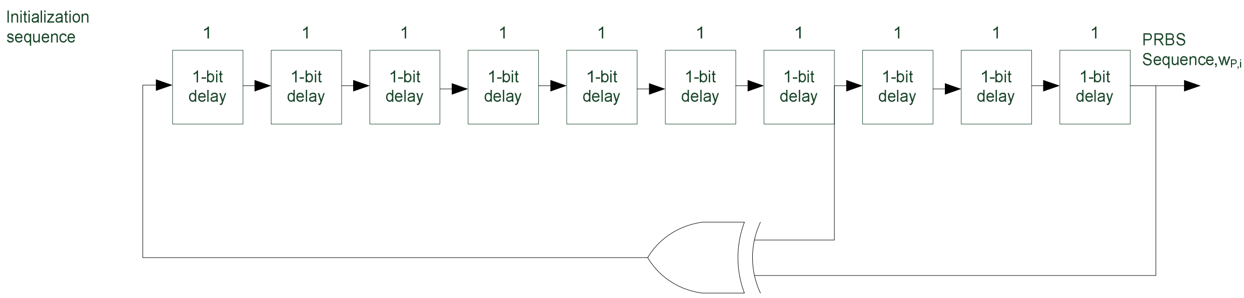

Figure 22: Pilot Scrambling Sequence

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

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

Due to the chosen polynomial, the sequence repeats itself every $2^{10}-1=1\,023$ outputs:

$$w'_i = w'_{(i \bmod 1\,023)} \tag{5}$$

8.1.2.3 Preamble Pilot Modulation Sequence

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

$$w_k^P = w_k \oplus w_i' \text{ with } i = (k \bmod K_{L1}) / D_P \tag{6}$$

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 reminder 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 we can calculate the values of w_k^P for L1 Block 100 (starts at 760,714 MHz in 8 MHz operation). Firstly, we have to calculate the starting OFDM sub-carrier of this L1 Block k:

$$k = K_{L1} \cdot 100 = 3\,408 \cdot 100 = 340\,800 \tag{7}$$

Hence, we obtain the following sequence:

$$\begin{aligned} w_{340\,800}^P &= w_{340\,800} \oplus w'_{(340\,800 \bmod 3\,408)/6} \\ &= w_{340\,800} \oplus w'_0 \end{aligned} \tag{8}$$

If we consider the cyclic repetition of the sequences, we obtain:

$$\begin{aligned} w_{340\,800}^P &= w_{(340\,800 \bmod 2\,047)} \oplus w'_0 \\ &= w_{998} \oplus w'_0 \\ &= 0 \oplus 1 = 1 \end{aligned} \tag{9}$$

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

$$\begin{aligned} &w_{340\,806}^P = w_{(340\,806 \bmod 2\,047)} \oplus w'_1 = w_{1\,004} \oplus w'_1 = 0 \oplus 1 = 1 \\ &w_{340\,812}^P = w_{1\,010} \oplus w'_2 = 1 \oplus 1 = 0, \quad w_{340\,818}^P = 1 \oplus 1 = 0, \quad w_{340\,824}^P = 1 \oplus 1 = 0 \end{aligned} \tag{10}$$

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:

$$r_k = \begin{cases} w_k^P & \text{if } k \bmod K_{L1} = 0 \\ r_{k-6} \oplus w_k^P & \text{otherwise} \end{cases} \tag{11}$$

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:

$$r_{340\,800} = w_{340\,800}^P = 1 \text{ (absolute modulation)} \tag{12}$$

$$r_{340\,806} = r_{340\,800} \oplus w_{340\,806}^P = 1 \oplus 1 = 0 \tag{13}$$

$$r_{340\,812} = r_{340\,806} \oplus w_{340\,812}^P = 0 \oplus 0 = 0 \tag{14}$$

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 wihtin 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:

$$\begin{aligned} \mathrm{Re}\left\{c^P_{m,l_P,k}\right\} &= A_{PP} \cdot 2\left(1/2 - r_k\right) \\ \mathrm{Im}\left\{c^P_{m,l_P,k}\right\} &= 0 \end{aligned} \tag{15}$$

while $c^P_{c,l_P,k}$ 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 not 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 $r_k = 1 \Rightarrow c^P_{m,l_P,k} = -1$ and $r_k = 0 \Rightarrow c^P_{m,l_P,k} = 1$.

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:

$$\begin{aligned} \mathrm{Re}\left\{c^P_{m,l_P,k}\right\} &= \mathrm{Re}\left\{a^P_{l_P,q}\right\} \cdot \left(-1\right)^{w_k} \\ \mathrm{Im}\left\{c^P_{m,l_P,k}\right\} &= \mathrm{Im}\left\{a^P_{l_P,q}\right\} \cdot \left(-1\right)^{w_k} \end{aligned} \quad \text{with } q = \left(k \bmod K_{L1}\right) - \left\lceil \left(k \bmod K_{L1}\right)/6 \right\rceil \tag{16}$$

while w_k is the k-th value of the data scrambling sequence, $a^P_{l_P,q}$ is the q-th output of the preamble frequency interleaver for the preamble OFDM symbol l_P, and $c^P_{c,l_P,k}$ 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 we can calculate the number of q for a given k:

$$k \rightarrow q : 0 \rightarrow Pilot, 1 \rightarrow 0, 2 \rightarrow 1, 3 \rightarrow 2, 4 \rightarrow 3, 5 \rightarrow 4, 6 \rightarrow Pilot, 7 \rightarrow 5, 8 \rightarrow 6, ... \tag{17}$$

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 equalisation. 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 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 we can treat them 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 & 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 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

When reserved carriers are activated by a relevant L1 signalling part 2, 'RSERVED_TONES', these reserved carriers defined in clause 9.7 of [i.1] can be used for PAPR reduction.

Figure 23 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 don't 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 disjointed 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.

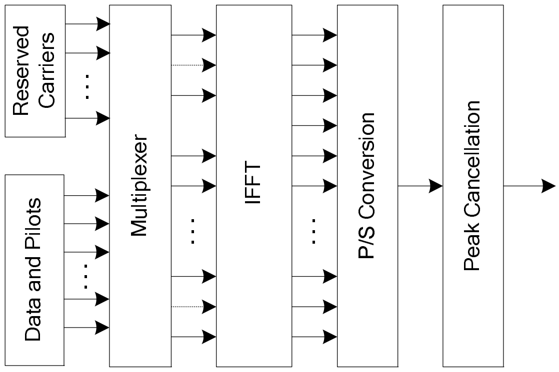

Figure 23: The structure of the OFDM transmitter with PAPR reduction using reserved carriers

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 [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 [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:

$$\mathbf{p} = \frac{\sqrt{N_{FFT}}}{N_{TR}} IFFT(1_{TR}) = \left[p_0, p_1, ..., p_{N_{FFT}-1} \right] \quad (18)$$

where N_{FFT} and N_{TR} indicate the FFT size and the number of reserved carriers, respectively. The (N_{FFT} , 1) vector $\boldsymbol{1}_{TR}$ 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_{NFFT-1}) of the kernel have a value considerably smaller than p_0.

8.3.3 Algorithm of PAPR reduction using reserved carriers

Figure 24 shows the detailed block diagram of the peak-cancellation algorithm. The IFFT output (**x**) is fed into the peak-cancellation block, and the peak position and value of **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 **x** and the new PAPR is calculated. The principle is shown in figure 25. 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.

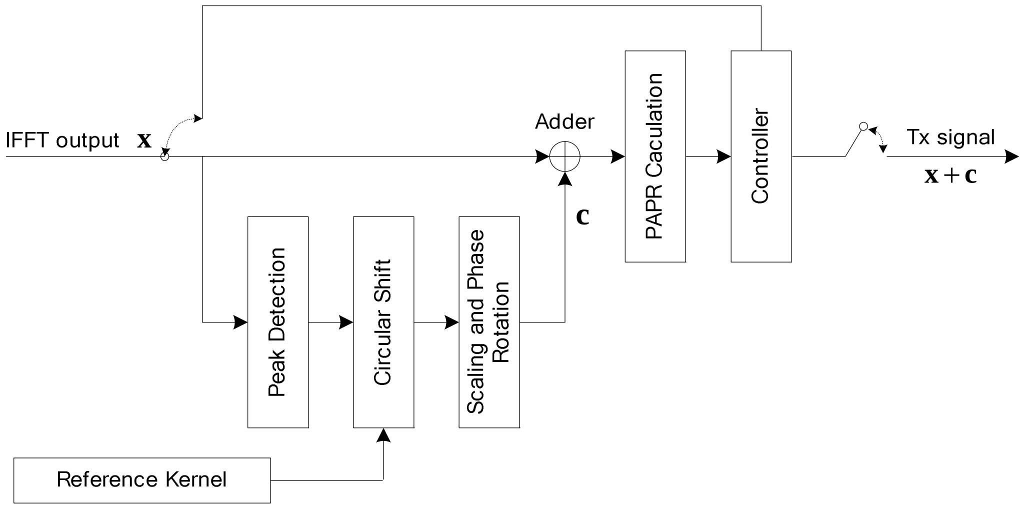

Figure 24: Block diagram of the peak-cancellation algorithm

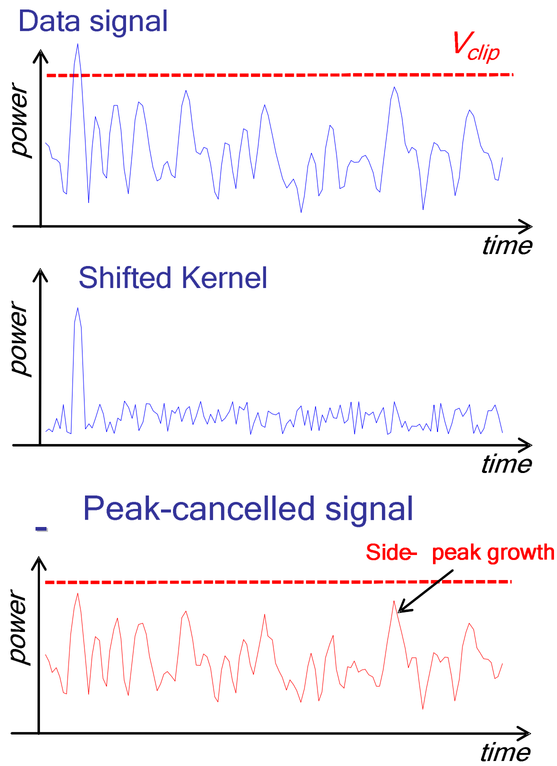

Figure 25: Principle of peak-cancellation algorithm

8.3.3.1 PAPR cancellation algorithm

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

Initialization:

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

$$\mathbf{c}^{(0)} = [0 \cdots 0]^T \qquad (19)$$

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

Iteration:

1) i starts from 1.

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

$$\begin{cases} y_i = \max\limits_n \left| x_n + c_n^{(i-1)} \right| \\ m_i = \arg\max\limits_n \left| x_n + c_n^{(i-1)} \right| \end{cases} , \quad \mathit{for}\ n = 0,1,\cdots N_{FFT}-1 \tag{20}$$

where x_n and $c_n^{(i-1)}$ represent the nth element of vector $\boldsymbol{x}$ and $\boldsymbol{c}^{(i-1)}$, 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 $\boldsymbol{c}^{(i)}$ as

$$\mathbf{c}^{(i)} = \mathbf{c}^{(i-1)} - \alpha_i \mathbf{p}(m_i) \quad , \mathit{where} \quad \alpha_i = \frac{x_{m_i} + c_{m_i}^{(i-1)}}{y_i}\left(y_i - V_{clip}\right) \tag{21}$$

where $\boldsymbol{p}(m_i)$ 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 1and 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:

$$\mathbf{x}' = \mathbf{x} + \mathbf{c}^{(i)} \tag{22}$$

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 jth iteration of the simple method is given as:

$$x' = x + \alpha_1 p(m_1) + \alpha_2 p(m_2) + \cdots + \alpha_{j-1} p(m_{j-1}) + \alpha_j p(m_j) = x + \sum_{n=1}^{j} \alpha_n p(m_n) \tag{23}$$

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:

$$x' = x + \underbrace{\alpha_1 p(m_1) + \alpha_2 p(m_2)}_{1} + \cdots + \underbrace{\alpha_{j-1} p(m_{j-1}) + \alpha_j p(m_j)}_{j/2} = x + \sum_{n=1}^{j/2} \left(\alpha_{2n-1} p(m_{2n-1}) + \alpha_{2n} p(m_{2n}) \right) \tag{24}$$

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

8.4 Signalling (L1 part 2, incl 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 part 2 signalling structure is illustrated in figure 26. All information is conveyed in L1 signalling part 2 data part and it can be changed frame by frame. L1 part 2 data has the Data slice loop and the Notch band loop and each Data slice loop has the PLP loop.

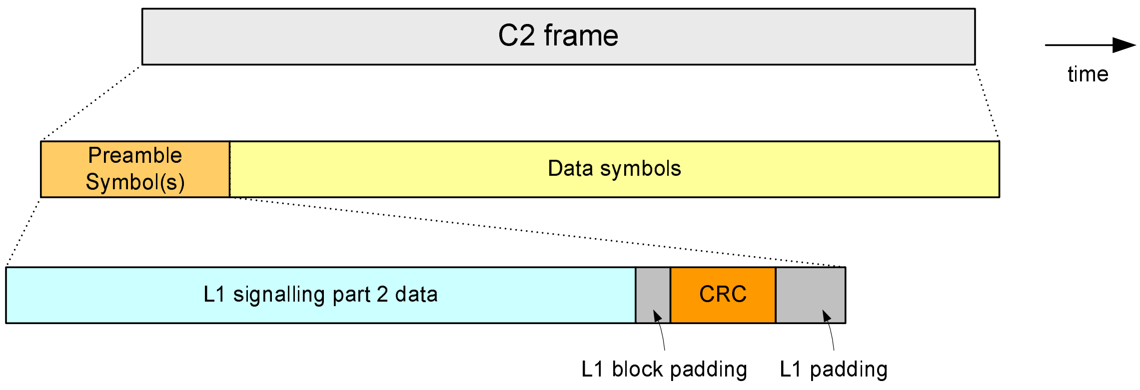

Figure 26: The L1 part 2 signalling structure

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 signalling part 2. 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 doesn't 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 part 2 signalling are specified in [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 13 gives a worked example of the signalling fields of the L1 part 2 signalling with Data Slice type 1.

Table 13: An example of the contents of the L1 part 2 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 C2 system ID is 0x0000
START_FREQUENCY	24	0352E0h	The start frequency of this C2 System is 486,2 MHz (in case of 7,61 MHz BW)
C2_BANDWIDTH	16	0000000010001110	Total number of carriers of this C2 system 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 Data Symbols in one C2 Frame
L1_PART 2_CHANGE_COUNTER	8	00000000	There are no changes foreseen in L1 configurable parameters
NUM_DSLICE	8	00000001	There is 1 Data Slice
NUM_NOTCH	4	0001	There is 1 Notch Band
DSLICE_ID	8	00000000	The Data Slice ID is 0x00
DSLICE_TUNE_POS	13	0000001000111	The tuning position of this Data Slice is 1 704th carrier frequency of this C2 System
DSLICE_OFFSET_LEFT	8	10111001	The left edge of this Data Slice is start frequency (apart from tuning position as much as 1 704 carrier spacing)
DSLICE_OFFSET_RIGHT	8	01000111	The right edge of this Data Slice 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	0	This Data Slice is Type 1
DSLICE_CONST_CONF	1	0	The configuration of this Data Slice can be changed every C2 Frame
DSLICE_LEFT_NOTCH	1	0	There is no left neighboured Notch band of this Data Slice
DSLICE_NUM_PLP	8	00000001	There is 1 PLP in this Data Slice
PLP_ID	8	00000000	The PLP ID is 0x00
PLP_BUNDLED	1	0	This PLP is not bundled with other PLP
PLP_TYPE	2	10	This PLP is normal Data PLP
PLP_PAYLOAD_TYPE	5	00011	This PLP carries a TS
PLP_START	14	00000000000010	First complete XFECframe starts from 2nd data cell of this Data Slice
PLP_FEC_TYPE	1	1	This PLP uses 64 K LDPC
PLP_MOD	3	011	This PLP uses 256-QAM modulation
PLP_COD	3	100	This PLP uses code rate 5/6
PSI/SI_REPROCESSING	1	1	PSI/SI of this PLPs TS is reprocessed
RESERVED_1	8	00000000	Reserved for future use
RESERVED_2	8	00000000	Reserved for future use
NOTCH_START	13	0000000000010	The Notch band starts from 49th carrier of this C2 System
NOTCH_WIDTH	8	00000001	The Notch band ends at 71th carrier of this C2 System
RESERVED_3	8	00000000	Reserved for future use
RESERVED_TONE	1	0	There is no reserved tone in this C2 Frame
RESERVED_4	16	0000000000000000	Reserved for future use

Since all of the bits of the L1 signalling part 2 data are 254-bits, there is a 1-bit L1 block padding following the L1 signalling part 2 data.

Table 14 gives an example of the signalling fields of the L1 part 2 signalling with Data Slice type 2.

Table 14: An example of the contents of the L1 part 2 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 C2 system ID is 0x0000
START_FREQUENCY	24	0352E0h	The start frequency of this C2 System is 486,2 MHz (in case of 7,61 MHz BW)
C2_BANDWIDTH	16	0000000010001110	Total number of carriers of this C2 system 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 Data Symbols in one C2 Frame
L1_PART 2_CHANGE_COUNTER	8	00000000	There are no changes foreseen in L1 configurable parameters
NUM_DSLICE	8	00000001	There is 1 Data Slice
NUM_NOTCH	4	0001	There is 1 Notch Band
DSLICE_ID	8	00000000	The Data Slice ID is 0x00
DSLICE_TUNE_POS	13	0000001000111	The tuning position of this Data Slice is 1704th carrier frequency of this C2 System
DSLICE_OFFSET_LEFT	8	10111001	The left edge of this Data Slice is start frequency (apart from tuning position as much as 1 704 carrier spacing)
DSLICE_OFFSET_RIGHT	8	01000111	The right edge of this Data Slice 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 Data Slice is Type 2
FEC_HEADER_TYPE	1	0	The type of the FECFrame header of this Data Slice is normal mode
DSLICE_CONST_CONF	1	0	The configuration of this Data Slice can be changed every C2 Frame
DSLICE_LEFT_NOTCH	1	0	There is no left neighboured Notch band of this Data Slice
DSLICE_NUM_PLP	8	00000001	There is 1 PLP in this Data Slice
PLP_ID	8	00000000	The PLP ID is 0x00
PLP_BUNDLED	1	0	This PLP is not bundled with other PLP
PLP_TYPE	2	10	This PLP is normal Data PLP
PLP_PAYLOAD_TYPE	5	00011	This PLP carries a TS
PSI/SI_REPROCESSING	1	1	PSI/SI of this PLPs TS is reprocessed
RESERVED_1	8	00000000	Reserved for future use
RESERVED_2	8	00000000	Reserved for future use
NOTCH_START	13	0000000000010	The Notch band starts from 49th carrier of this C2 System
NOTCH_WIDTH	8	00000001	The Notch band ends at 71th carrier of this C2 System
RESERVED_3	8	00000000	Reserved for future use
RESERVED_TONE	1	0	There is no reserved tone in this C2 Frame
RESERVED_4	16	0000000000000000	Reserved for future use

Since all of the bits of the L1 signalling part 2 data are 234-bits, there is a 1-bit L1 block padding following the L1 signalling part 2 data.

8.4.5 FEC for the L1 signalling part 2

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 part 2 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_{bch} = 7 032 and a 16 K LDPC code with K_{ldpc} = 7 200 defined in clause 6.1 in [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_{ldpc} is the same as (K_{bch} + 168) = 7 200 since the number of BCH parity bits in BCH encoding, $N_{\mathrm{bch\_parity}}$ 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 [i.1]. A segmented L1 signalling part 2 has a length less than or equal to $N_{L1part\,2\_\max\_per\_Symbol}$ = 4 759. $N_{L1part\,2\_\max\_per\_Symbol}$ means the maximum number of L1 part 2 information bits for transmitting the coded L1 part 2 through one OFDM symbol.

For the shortening operation, an input parameter K_{sig} is first obtained from the length of the segmented L1 signalling part 2 as described in clause 8.4.2 in [i.1]. Since K_{sig} is always less than K_{bch}, the BCH information part is shortened for encoding. More precisely, K_{bch} BCH information bits are filled with K_{sig} signalling bits and (K_{bch} - K_{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_{ldpc}/360 (= N_{group}) bit-groups which are (N_{group} - 1) groups of 360 bits and 1 group of 192 bits as illustrated in figure 27. That is, K_{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_{bch} - K_{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.

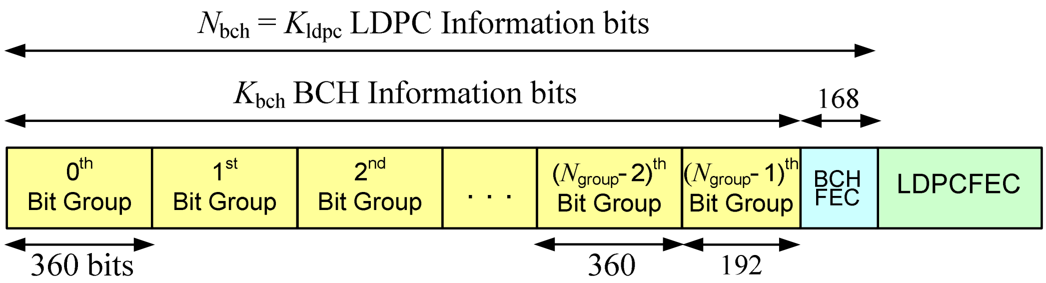

Figure 27: Format of data after LDPC encoding of L1 signalling part 2

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. $K_{L1part\,2\_ex\_pad}$ = 11 956. $K_{L1part\,2\_ex\_pad}$ 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 [i.1]. Then, the number of LDPC blocks needed for the protection of L1 signalling part 2, $N_{L1part\,2\_FEC\_Block}$ is determined by:

$$N_{L1part2\_FEC\_Block} = \left\lceil \frac{K_{L1part2\_ex\_pad}}{N_{L1part2\_\max\_per\_Symbol}} \right\rceil = \left\lceil \frac{11956}{4759} \right\rceil = 3 \qquad (25)$$

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:

$$\begin{aligned} K_{L1\_PADDING} &= \left\lceil \frac{K_{L1part2\_ex\_pad}}{N_{L1part2\_FEC\_Block}} \right\rceil \times N_{L1part2\_FEC\_Block} - K_{L1part2\_ex\_pad} \\ &= \left\lceil \frac{11956}{3} \right\rceil \times 3 - 11956 = 2. \end{aligned} \qquad (26)$$

Then the final length of the whole L1 signalling part 2 including the padding field, $K_{L1part\,2}$ is set as follows:

$$K_{L1part2} = K_{L1part2\_ex\_pad} + K_{L1\_PADDING} = 11956 + 2 = 11958 \quad (27)$$

These $K_{L1part\,2}$ signalling bits are divided into 3 (= $N_{L1part\,2\_FEC\_Block}$) blocks consisting of K_{sig} bits which is calculated as

$$K_{sig} = \frac{K_{L1part2}}{N_{L1part2\_FEC\_Block}} = \frac{11958}{3} = 3986 \quad (28)$$

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 [i.1].

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

$$X_0 = \{ m_0, m_1, m_2, \ldots, m_{359} \} \quad \text{(360 bits)} \quad (29)$$

$$X_1 = \{ m_{360}, m_{361}, m_{362}, \ldots, m_{719} \} \quad \text{(360 bits)} \quad (30)$$

$$X_{18} = \{ m_{6480}, m_{6481}, m_{6482}, \ldots, m_{6839} \} \quad \text{(360 bits)} \quad (31)$$

$$X_{19} = \{ m_{6840}, m_{6841}, m_{6842}, \ldots, m_{7031} \} \quad \text{(192 bits)} \quad (32)$$

The bits in each bit-group correspond to the data field in ascending order as illustrated in figure 28.

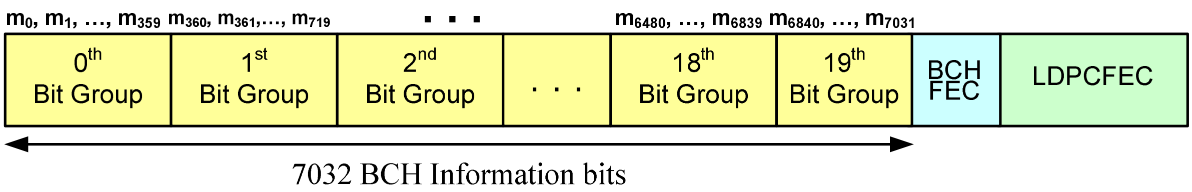

Figure 28: Allocation of BCH information bits in Data field

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:

$$N_{pad} = \left\lfloor \frac{7032 - 3986}{360} \right\rfloor = \left\lfloor \frac{3046}{360} \right\rfloor = 8 \quad (33)$$

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 [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, $\{m_0, m_1, \ldots, m_{7031}\}$, by the above procedure.

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

$$m_i = s_i, \quad 0 \le i < 1634 \tag{34}$$

$$m_i = 0, \quad 1634 \le i < 1800 \tag{35}$$

$$m_i = s_{i-166}, \quad 1800 \le i < 3960 \tag{36}$$

$$m_i = 0, \quad 3960 \le i < 6840 \tag{37}$$

$$m_i = s_{i-3046}, \quad 6840 \le i < 7032 \tag{38}$$

Figure 29 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.

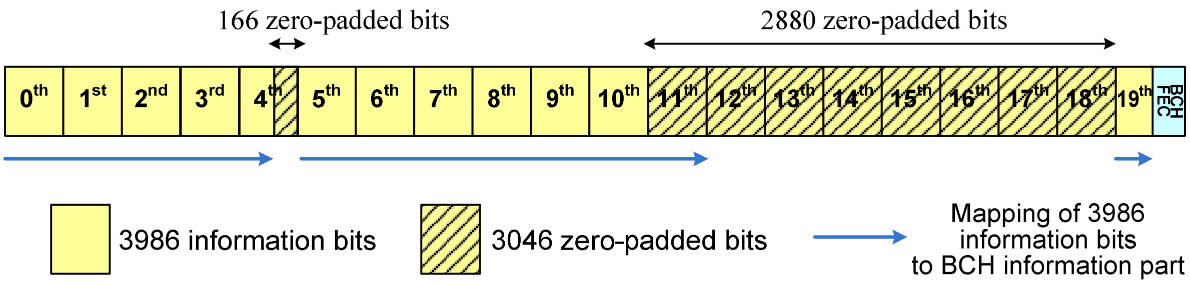

Figure 29: Example of Shortening of BCH information part

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 [i.1] to generate $N_{\text{bch}} = K_{\text{ldpc}}$ output bits $(i_0 \ldots i_{N_{\text{bch}}-1})$.

8.4.5.4 LDPC encoding

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

$$\Lambda = (i_0, i_1, \ldots, i_{K_{\text{ldpc}}-1}, p_0, p_1, \ldots, p_{N_{\text{ldpc}}-K_{\text{ldpc}}-1}) = (i_0, \ldots, i_{7199}, p_0, p_1, \ldots, p_{8999}) \tag{39}$$

according to clause 6.1.2 in [i.1].

8.4.5.5 Puncturing of LDPC parity bits

The 16 K LDPC code with K_{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_{ldpc} is 16 200, its effective LDPC code rates (R_{eff}) is 4/9. All LDPC parity bits, denoted by $\{p_0, p_1, \ldots, p_{8999}\}$, are divided into 25 (= Q_{ldpc}) parity bit groups where each parity group is formed from a sub-set of the LDPC parity bits as follows:

$$P_0 = \left\{ p_0, p_{Q_{ldpc}}, p_{2Q_{ldpc}}, \ldots, p_{359Q_{ldpc}} \right\} = \left\{ p_0, p_{25}, p_{50}, \ldots, p_{8975} \right\} \tag{40}$$

$$P_1 = \left\{ p_1, p_{Q_{ldpc}+1}, p_{2Q_{ldpc}+1}, \ldots, p_{359Q_{ldpc}+1} \right\} = \left\{ p_1, p_{26}, p_{51}, \ldots, p_{8976} \right\} \tag{41}$$

$$P_2 = \{ p_2, p_{Q_{ldpc}+2}, p_{2Q_{ldpc}+2}, \ldots, p_{359Q_{ldpc}+2} \} = \{ p_2, p_{27}, p_{52}, \ldots, p_{8977} \} \quad (42)$$

$$P_{24} = \{ p_{Q_{ldpc}-1}, p_{2Q_{ldpc}-1}, p_{3Q_{ldpc}-1}, \ldots, p_{360Q_{ldpc}-1} \} = \{ p_{24}, p_{49}, p_{74}, \ldots, p_{8999} \} \quad (43)$$

where P_j represents the jth parity group and $Q_{\text{ldpc}} = 25$ is given in table 5b in [i.1]. Each group has 9 000/Q_{ldpc} = 360 bits, as illustrated in figure 30. Note that the number of parity bits of a DVB-C2 LDPC code is always a multiple of 360 and Q_{ldpc}.

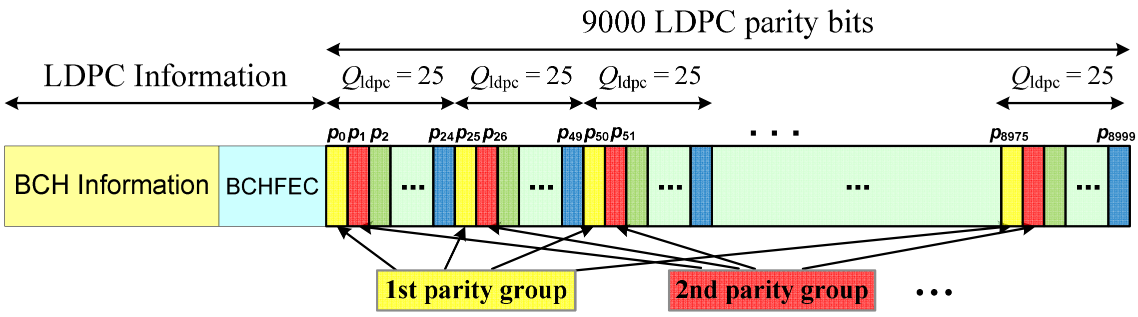

Figure 30: Parity-bit groups in an FEC block

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_{punc}, the positions of the parity bits are allocated by unit of parity-bit groups. The procedure for calculating N_{punc} is a little complicated as described in clause 8.4.2 in [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 part 2 must be a multiple of the number of OFDM symbols for transmitting L1 part 2.

For easy understanding, a concrete example of calculating N_{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_{sig} is 3 986 as calculated in clause 8.4.6.2. Note that the coded L1 part 2 is always transmitted using 16-QAM. For a given K_{sig} = 3 986 and modulation order 16-QAM, N_{punc} is determined by the following steps:

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

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

Step 3) For the sake of convenience, assume that the coded L1 part 2 is transmitted without an extended time interleaving, that is, the value of L1_TI_MODE is 00 or 01.

$$N_{L1part2} = \left\lceil \frac{N_{L1part2\_temp}}{2\eta_{MOD} \times N_{L1part2\_FEC\_Block}} \right\rceil \times 2\eta_{MOD} \times N_{L1part2\_FEC\_Block} = \left\lceil \frac{9499}{8 \times 3} \right\rceil \times 8 \times 3 = 9504. \ (\eta_{\text{MOD}} = 4) \quad (44)$$

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, $R_{eff\_L1part\,2}$ is always less than $R_{eff\_16K\_LDPC\_1\_2}$ (= 4/9). Furthermore, $R_{eff\_L1part\,2}$ 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 in Step 1).

Step 3 guarantees that $N_{L1part\,2}$ is a multiple of the number of columns of the bit interleaver (described in clause 8.4.3.6 in [i.1]) and that $N_{L1part\,2}/\eta_{\mathrm{MOD}}$ 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. $N_{L1part\,2}$ 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

$$N_{MOD\_per\_Block} = \frac{N_{L1part2}}{\eta_{MOD}} = \frac{9504}{4} = 2376 \tag{45}$$

modulated cells. The total number of cells in all $N_{L1part\,2\_FEC\_Block}$ blocks, N_{MOD_Total} is $N_{MOD\_per\_Block} \times N_{L1part\,2\_FEC\_Block}$ = 2 376 × 3 = 7 128.

Finally, for N_{punc} = 3 650 given in Step 4), the positions of bits to be punctured are determined as follows:

Step 1) Compute N_{punc_groups} 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 (= 3 650 - 3 600) parity bits in the first part of the group will be additionally punctured. That is, the parity bits p_2, p_{27}, p_{52}, ..., 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 [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 9 000 - N_{punc} (= N_{ldpc} - K_{ldpc} - N_{punc}) LDPC parity bits without N_{punc} punctured bits, as illustrated infigure 31.

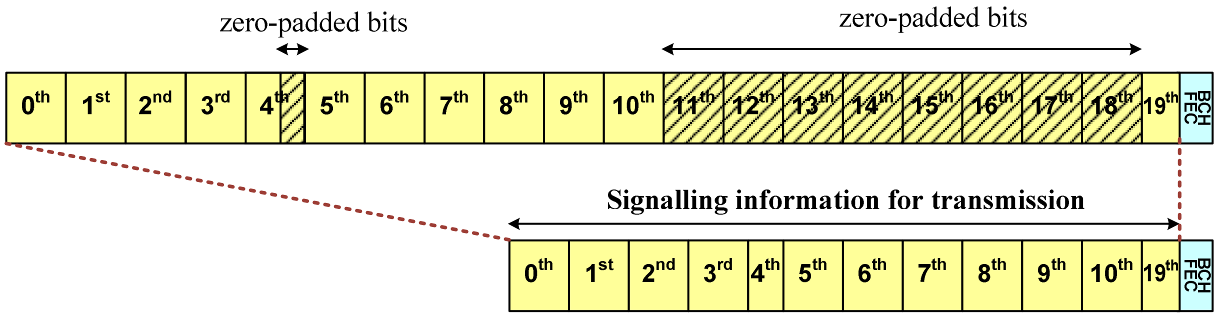

Figure 31: Example of removal of zero-padding bits

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 [i.1].

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

8.5 Interleaving

8.5.1 Bit interleaving

Bit Interleaving is described in clause 6.1.3 of [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 [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 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 C2 Frame.

Each TI block of a Data Slice within a C2 Frame contains exactly same number of OFDM cells (see clause 9.4.4 of [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 same PLP. The cells from different PLPs can be interleaved into one TI block. However, each TI block should contains all data cells carried by TI-depth OFDM symbols within the Data Slice.

The address generation equation in clause 9.4.4 of [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.

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 symbol to the next. The number of payload cells per Data Slice per OFDM symbol (N_{DS}) can vary from symbol to 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, symbol number, L1 info*) exists in the modulator to provide N_{DS} for a given Data Slice number, symbol number and from the Layer 1 information. The frequency interleaver must 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 must be capable of dealing with N_{max} payload cells where:

$$N_{max} = (K_{DS,max} - K_{DS,min}) - N_{SP,Dx=24} - N_{CP} \quad (46)$$

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 symbols and Bank B for odd OFDM symbols. Each memory bank comprises of $N_{\max}$ locations. DVB-C2 uses odd-only pseudo-random interleaving. In this the payload cells from even symbols (symbol number of form 2*n*) 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 symbols (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 [i.1]. In order to produce a continuous stream of cells at its output, when Bank A is being written (incoming even symbol), Bank B is also being read (outgoing previous odd 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 symbols in the Data Slice contained $N_{\max} = C_{\text{data}}$ payload cells, then the number of write addresses for symbol number $2n + 1$ would match the number of read addresses for symbol number 2*n* otherwise some data cells of symbol 2*n* will be skipped. Unfortunately N_{DS} can be different from symbol to 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, symbol number, L1 info*) returns the number of payload cells in the current slice for the given 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 symbol of number 2*n*:

1) $q = 0$;
2) C_{max} = **max**(**DataCells**(*slice number* , 2*n-1, L1 info*), **DataCells**(*slice number* , 2*n*, *L1 info*));
3) Generate address $H_1(q)$;
4) rdEnable = ($H_1(q)$ < **DataCells**(*slice number* , 2*n-1*, *L1 info*));
5) wrEnable = (q < **DataCells**(*slice number* , 2*n*, *L1 info*));
6) if (rdEnable) Read cell *q* of output interleaved symbol 2*n* - 1 from location $H_1(q)$ of memory Bank B;
7) Store cell *q* of incoming un-interleaved symbol 2*n* 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 ($q < C_{\max}$) goto 3.

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

1) $q = 0$;
2) C_{max} = **max**(**DataCells**(*slice number* , 2*n*, *L1 info*), **DataCells**(*slice number* , 2*n + 1*, *L1 info*));
3) Generate address $H_0(q)$;
4) rdEnable = ($H_0(q)$ < **DataCells**(*slice number* , 2*n*, *L1 info*));
5) wrEnable = (q < **DataCells**(*slice number* , 2*n + 1*, *L1 info*));
6) if (rdEnable) Read cell *q* of output interleaved symbol 2*n* from location $H_0(q)$ of memory Bank A;
7) Store cell *q* of incoming un-interleaved symbol 2*n + 1* into location *wptr* of HoldBuffer and increment *wptr;*

8) if (wrEnable):

   a) Write cell *rptr* 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 ($q < C_{\max}$) goto 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 symbol, he 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 x. The C2 Frame structure comprises L_P Preamble Symbols (L_P >=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:

$$T_F = (L_P+L_{data})* T_s \qquad (47)$$

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 neighboured 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).

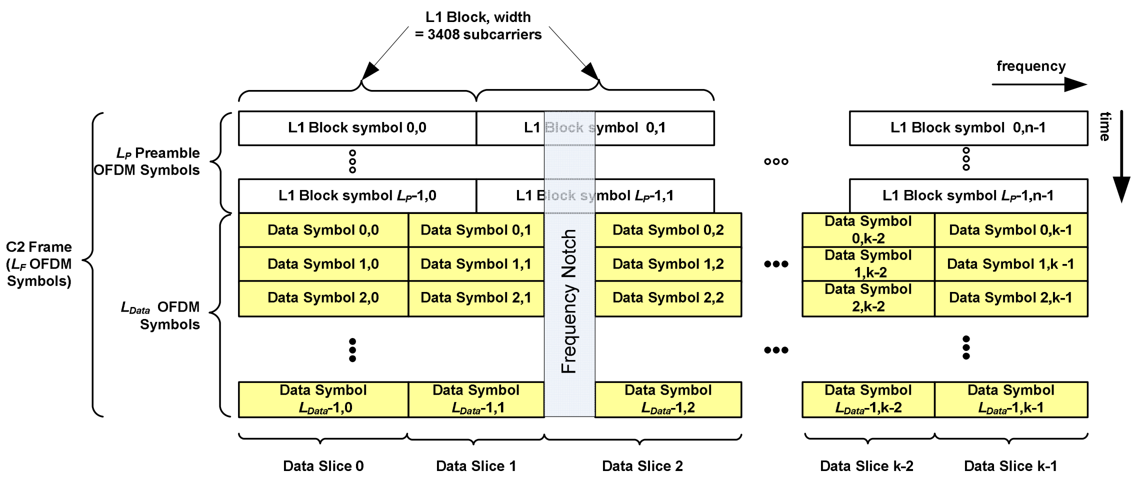

Figure 32: The C2 Frame structure.
The C2 Frame starts with at least one Preamble Symbol (L_P) followed by L Data Symbols

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 [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, …). The C2 standard differs between narrowband and broadband notches and is described in more detail in clause 9.3.5 of [i.1].

8.7 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 [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. Else, unwanted phase jumps may be generated between adjacent OFDM symbols that could disturb the synchronisation 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 [i.1] defines the emitted passband signal by the following expressions:

$$s(t) = \operatorname{Re}\left\{ \sum_{m=0}^{\infty} \left[\frac{1}{\sqrt{K_{total}}} \sum_{l=0}^{L_F-1} \sum_{k=K_{\min}}^{K_{\max}} c_{m,l,k} \cdot \psi_{m,l,k}(t) \right] \right\} \quad (48)$$

where:

$$\psi_{m,l,k}(t)=\begin{cases} e^{j2\pi\frac{k}{T_U}(t-\Delta-lT_s-mT_F)} & mT_F+lT_S \le t < mT_F+(l+1)T_S \\ 0 & \text{otherwise} \end{cases} \tag{49}$$

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;
- Δ 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 Ψ to obtain the same output signal:

$$s(t)=\frac{1}{\sqrt{K_{total}}}\cdot\text{Re}\left\{e^{j2\pi f_c t}\cdot\sum_{m=0}^{\infty}\sum_{l=0}^{L_F-1}\sum_{k=K_{\min}}^{K_{\max}}c_{m,l,k}\cdot\Psi'_{m,l,k}(t)\right\} \tag{50}$$

with

$$\psi'_{m,l,k}(t)=\begin{cases} e^{j2\pi\frac{k}{T_U}(t-\Delta-lT_s-mT_F)}\cdot e^{-j2\pi f_c t} & mT_F+lT_S \le t < mT_F+(l+1)T_S \\ 0 & \text{otherwise} \end{cases} \tag{51}$$

Equation (51) cannot be directly transformed into the equation known from clause 9.5 of the DVB-T2 specification [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:

$$f_c=\frac{k_c}{T_U} \tag{52}$$

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:

$$\psi'_{m,l,k}(t) = \begin{cases} e^{j2\pi \frac{k'+k_c}{T_U}(t-\Delta-lT_s-mT_F)} \cdot e^{-j2\pi \frac{k_c}{T_U} t} & mT_F + lT_S \le t < mT_F + (l+1)T_S \\ 0 & \text{otherwise} \end{cases} \quad (53)$$

which can be reformulated as:

$$\psi'_{m,l,k}(t) = \begin{cases} e^{j2\pi \frac{k'}{T_U}(t-\Delta-lT_s-mT_F)} \cdot e^{-j2\pi \frac{k_c}{T_U}\Delta(1+l+mL_F)} & mT_F + lT_S \le t < mT_F + (l+1)T_S \\ 0 & \text{otherwise} \end{cases} \quad (54).$$

Equation (54) looks similar to the signal definition of the DVB-T2 signal as described in clause 9.5 of [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 (54) can be written as:

$$\psi'_{m,l,k}(t) = \begin{cases} e^{j2\pi \frac{k'}{T_U}(t-\Delta-lT_s-mT_F)} \cdot e^{-j2\pi \frac{k_c}{T_U} \cdot T_U \left(\frac{\Delta}{T_U}\right)(1+l+mL_F)} & mT_F + lT_S \le t < mT_F + (l+1)T_S \\ 0 & \text{otherwise} \end{cases} \quad (55),$$

where $\left(\frac{\Delta}{T_U}\right)$ is the relative Guard Interval duration, i.e. 1/128 of 1/64. Additional simplification of (55) leads to:

$$\psi'_{m,l,k}(t) = \begin{cases} e^{j2\pi \frac{k'}{T_U}(t-\Delta-lT_s-mT_F)} \cdot e^{-j2\pi k_c \cdot \left(\frac{\Delta}{T_U}\right)(1+l+mL_F)} & mT_F + lT_S \le t < mT_F + (l+1)T_S \\ 0 & \text{otherwise} \end{cases} \quad (56).$$

Hence, this leads to a common phase rotation of:

$$\varphi_{k_c} = -2\pi \cdot k_c \left(\frac{\Delta}{T_U}\right) \quad (57)$$

for all OFDM subcarriers between two consecutive OFDM symbols, which depends on the choice of the relative Guard Interval duration (Δ / T_U) (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 2π. Hence, if k_c is multiple of 128 for Guard Interval 1/128, or multiple of 64 for Guard Interval 1/64, equation (56) can be written as:

$$\psi'_{m,l,k}(t) = \begin{cases} e^{j2\pi \frac{k'}{T_U}(t-\Delta-lT_s-mT_F)} & mT_F + lT_S \le t < mT_F + (l+1)T_S \\ 0 & \text{otherwise} \end{cases} \tag{58}$$

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

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 optimised 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 (CPE) 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:

$$f_c = \frac{k_c}{T_U} \tag{59}$$

with

$$k_c = \frac{K_{\max} + K_{\min}}{2} \tag{60}$$

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:

$$s(t) = \frac{1}{\sqrt{K_{total}}} \cdot \mathrm{Re}\left\{ e^{j2\pi f_c t} \cdot \sum_{m=0}^{\infty} \sum_{l=0}^{L_F-1} \sum_{k=K_{\min}}^{K_{\max}} \left(c_{m,l,k} \cdot e^{j\varphi_{m,l}} \right) \cdot \Psi''_{m,l,k}(t) \right\} \tag{61}$$

with

$$\Psi''_{m,l,k}(t) = \begin{cases} e^{j2\pi \frac{k'}{T_U}(t-\Delta-lT_s-mT_F)} & mT_F + lT_S \le t < mT_F + (l+1)T_S \\ 0 & \text{otherwise} \end{cases} \tag{62}$$

and

$$\varphi_{m,l} = -\varphi_{k_c} \cdot (1 + l + m \cdot L_F) \tag{63}$$

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;
- Δ 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;
- φ_{k_c} Phase jump between two consecutive OFDM symbols as calculated equation (57) of clause 8.7.1.

Practically, this generation is equivalent to the generation of a DVB-T2 signal [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'_k that is used for calculating the inverse FFT is the inner bracket of equation (61), i.e. $\left(c_{m,l,k} \cdot e^{j\varphi_{m,l}}\right)$.

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 [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:

$$k_c = \left\lfloor \frac{K_{\max} + K_{\min}}{2} \cdot \frac{\Delta}{T_U} + \frac{1}{2} \right\rfloor \cdot \frac{1}{\left(\frac{\Delta}{T_U}\right)} \quad (64),$$

where Δ / T_U is the relative Guard Interval duration (i.e. 1/64 or 1/128). Practically, equation (64) obtains the carrier k_c that it is closest to the centre OFDM subcarrier $\left(K_{\max} + K_{\min}\right)/2$, and additionally, generates multiples of 2π in equation (57) of the previous clause.

Consequently, the optimum carrier frequency f_c is:

$$f_c = \frac{k_c}{T_U} \quad (65),$$

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:

$$s(t) = \frac{1}{\sqrt{K_{total}}} \cdot \mathrm{Re}\left\{ e^{j2\pi f_c t} \cdot \sum_{m=0}^{\infty} \sum_{l=0}^{L_F-1} \sum_{k=K_{\min}}^{K_{\max}} c_{m,l,k} \cdot \Psi'_{m,l,k}(t) \right\} \qquad (66)$$

with

$$\Psi'_{m,l,k}(t) = \begin{cases} e^{j2\pi \frac{k'}{T_U}(t-\Delta-lT_s-mT_F)} & mT_F + lT_S \le t < mT_F + (l+1)T_S \\ 0 & \text{otherwise} \end{cases} \qquad (67)$$

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;
- Δ 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 (66).

8.7.3 OFDM Generation Block Diagram

Figure 33 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.

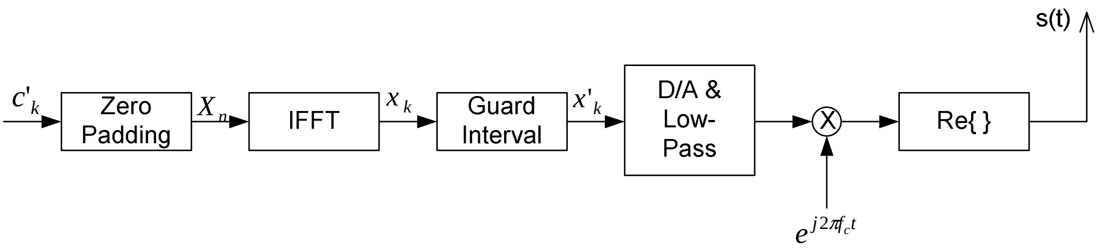

Figure 33: Possible implementation of OFDM generation

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

$$K_{total} = K_{\max} - K_{\min} + 1 \le N = K_{total} + x \qquad (68)$$

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

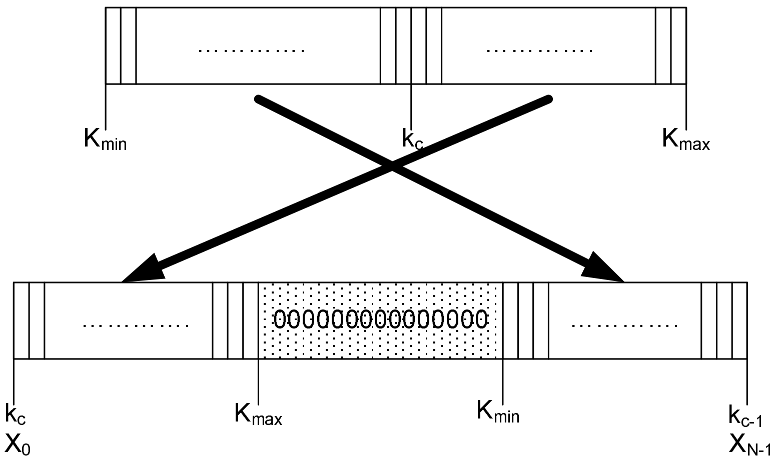

Figure 34: Principle of the Zero Padding

Figure 34 depicts the principle of the Zero Padding. In principle, it realises a cyclic shift operation on the actually used OFDM subcarriers and inserts zeros to the *x* (see equation (68)) remaining positions. Mathematically this operation can be described as:

$$X(n)_{m,l} = \begin{cases} c'_{m,l,k_c+n} & 0 \le n \le K_{\max} - k_c \\ 0 & \text{otherwise} \\ c'_{m,l,k_c+(n-N)} & N - (k_c - K_{\min}) \le n < N \end{cases} \quad \textit{for } 0 \le n < N \qquad (69)$$

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:

$$x(k)_{m,l} = \frac{1}{\sqrt{K_{total}}} \sum_{k=0}^{N-1} X(n)_{m,l} \cdot e^{j2\pi \frac{k \cdot n}{N}} \qquad (70$$

for $0 \le k < N$, 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 35 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 (71).

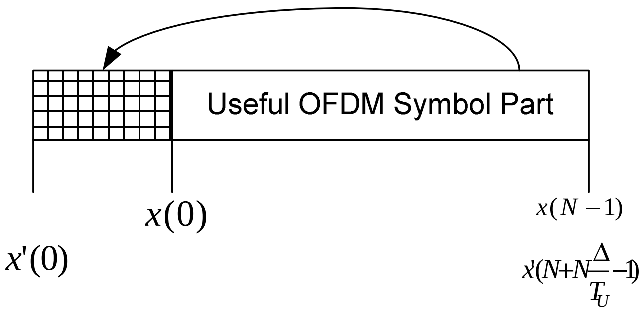

Figure 35: Generation of the Guard Interval

$$x'(k)_{m,l} = \begin{cases} x\left(k + N - N \cdot \frac{\Delta}{T_U}\right) & 0 \le k < N \cdot \frac{\Delta}{T_U} \\ x\left(k - N \cdot \frac{\Delta}{T_U}\right) & N \cdot \frac{\Delta}{T_U} \le k < N + N \cdot \frac{\Delta}{T_U} \end{cases} \qquad (71)$$

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 36 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 (66) of clause 8.7.3.1) are impractical.

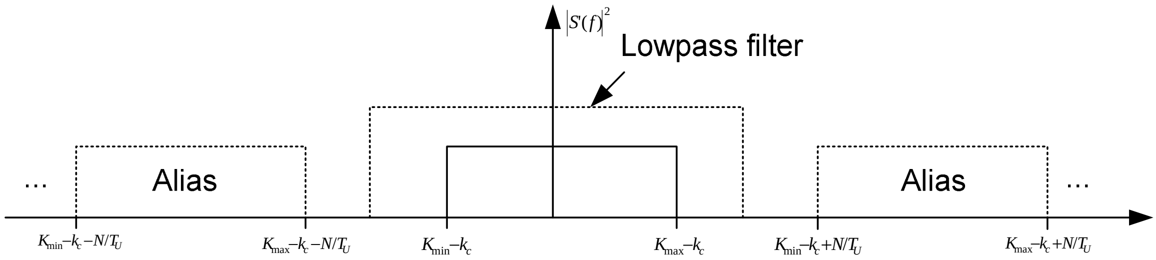

Figure 36: Spectrum of the digital signal and its aliases

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

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 who 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 [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 [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 37 shows a DVB-C2 signal transmitted in an 8 MHz channel. The corresponding mathematical expression for calculating the PDF is given in the DCB-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 $K_{min}/T_U = f_{sc,min} \leq f \leq f_{sc,msx} = K_{max}/T_U$ 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 $f < f_{sc,min} = K_{min}/T_U$ and above the highest sub-carrier frequency $f > f_{sc,max} = K_{max}/T_U$. 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 $f_{sc,min}$ and $f_{sc,max}$, 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.

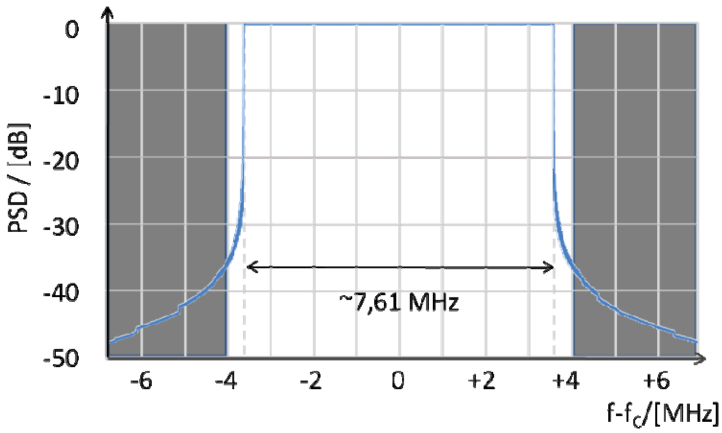

Figure 37: Power Spectral Density of a DVB-C2 RF signal transmitted in an 8 MHz channel

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 38 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 standardised 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 optimisation 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 standardised by IEC and CENELEC in their standards series IEC/EN 60728, and are defined in part 1 [i.18].

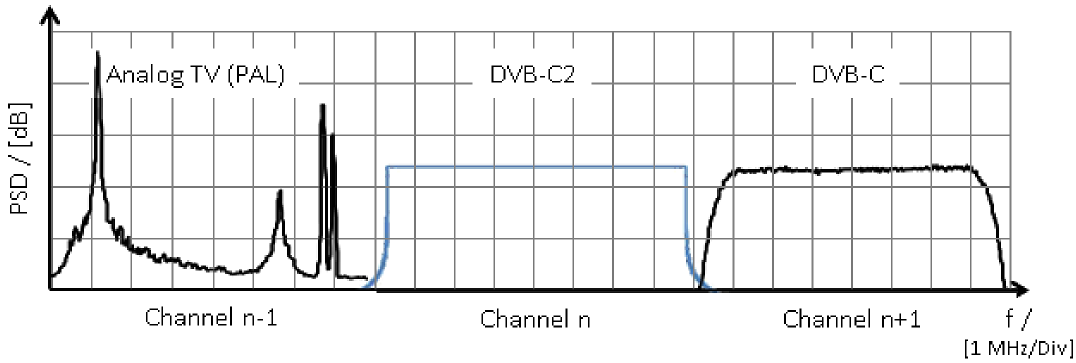

Figure 38: Qualitative depiction of an adjacent channel scenario comprising an analogue TV signal (measured PSD), a DVB-C2 signal (simulated PSD), and a DVB-C signal (measured PSD) in a cable system supporting 8 MHz channel spacing

9 Network

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-distortion signal ratio (SNR and CINR).

9.1 Components of a cable network

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 Ω. 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

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, the worst-case echo as specified by the IEEE 802.14 [i.16] model has been assumed as a reference. This IEEE 802.14 [i.16] worst-case echo specification is shown figure 39. To validate this echo requirement, the echo in two life 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].

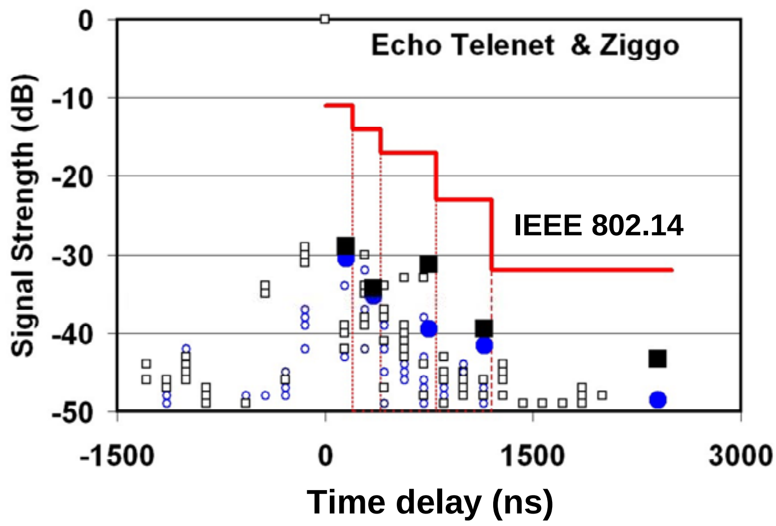

Figure 39: Worst case echo

NOTE 1: Figure 39 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 15. 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 15 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 15: 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/100m (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

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 practise only 2nd and 3rd 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 dBμV for which respectively a 60 dB worst-case CIR_{CSO} and CIR_{CTB} value is found. If it is assumed that only 2nd and 3rd 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 40 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 2nd and 3rd 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 40 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.

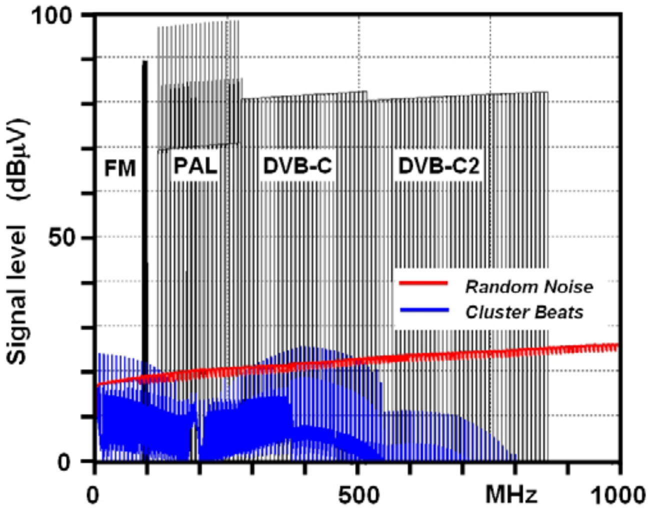

Figure 40: Spectrum of cable signals and intermodulation products

NOTE: Figure 40 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 2nd and 3rd 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 41. The figure 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 100ns associated with the 5 MHz bandwidth resolution of the RF capturing system.

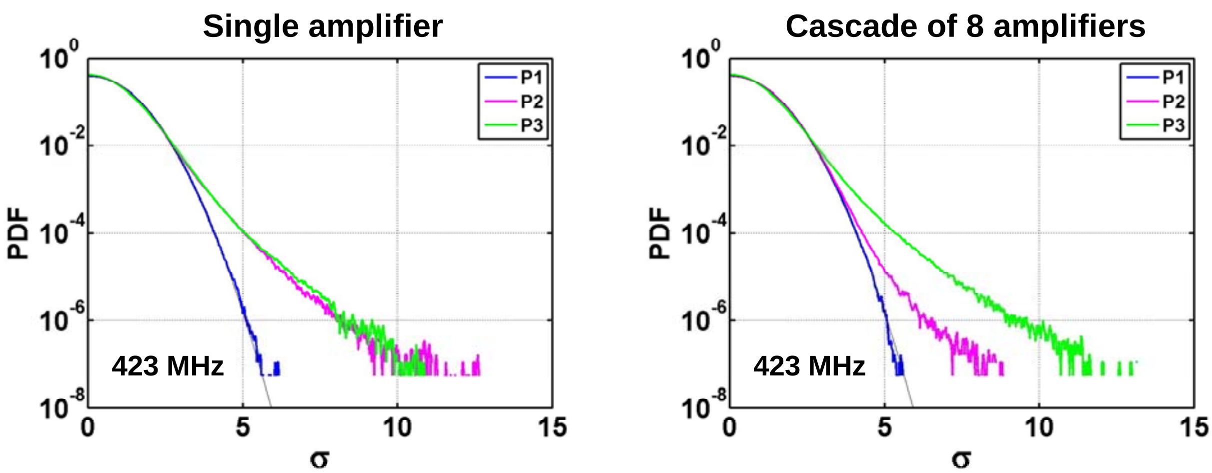

Figure 41: Probability density function (PDF) of the distortion signal caused by amplifiers

NOTE: Figure 41 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 P2 are associated with impulse events.

9.3 Signal Requirements

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 must 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 42.

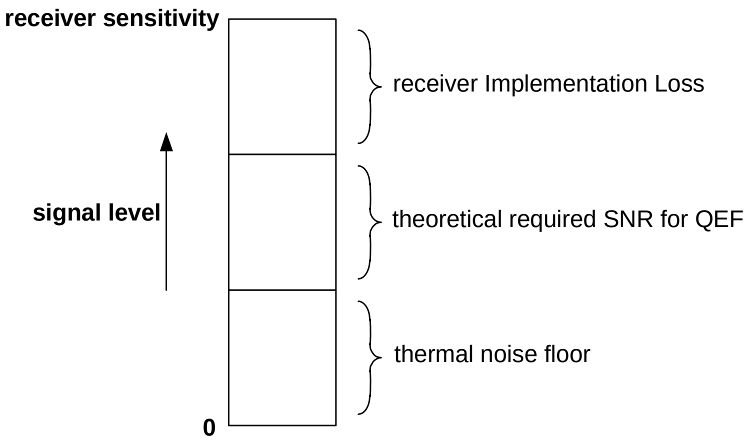

Figure 42: Budget calculation for receiver sensitivity

The thermal noise floor of broadband cable technologies with a 8 MHz channel width is about 4 dBµV. 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 18 and figure 62 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 1024 and lower QAM modulation modes and 12 dB for the 4096-QAM modulation mode. Table 16 lists the breakdown and the minimum DVB-C2 signal level.

Table 16: Estimated minimum DVB-C2 signal level at the system outlet

Modulation	**CR**	**Noise Floor (dBµV)**	**SNR (dB)**	**Implementation margin (dB)**	**In-home margin (dB)**	**Minimum signal level (dBµV)**
16-QAM	4/5	4	**10,7**	10	7	**31,7**
	9/10	4	**12,8**	10	7	**33,8**
64	2/3	4	**13,5**	10	7	**34,5**
	4/5	4	**16,1**	10	7	**37,1**
	9/10	4	**18,5**	10	7	**39,5**
256	3/4	4	**20,0**	11	7	**42,0**
	5/6	4	**22,0**	11	7	**44,0**
	9/10	4	**24,0**	11	7	**46,0**
1024	3/4	4	**24,8**	11	7	**46,8**
	5/6	4	**27,2**	11	7	**49,2**
	9/10	4	**29,5**	11	7	**51,5**
4096	5/6	4	**32,4**	12	7	**55,4**
	9/10	4	**35,0**	12	7	**58**

9.3.2 Signal quality requirements

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]. For all services, the signal quality must be compliant to the appropriate specifications.

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], paragraph 5.8.

The Carrier to ntermodulation Ratio (CIR) 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], paragraph 4.5.3. The requirement is specified in IEC 60728-1 [i.18], paragraph 5.9.3.

9.3.2.2 FM radio

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

9.3.2.3 DVB-C

The minimum composite intermodulation noise ration (CINR) is specified in IEC 60728-1 [i.18], paragraph 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 43. 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 43a depicts a simulation where 44 subcarriers ($44 \cdot 2{,}232 kHz \approx 100 kHz$ for 8 MHz channel raster) have been notched (see figure 43b). 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 43a. Here, approximately 0,1 dB would have to be added additionally to the minimum required SNR value for error free reception.

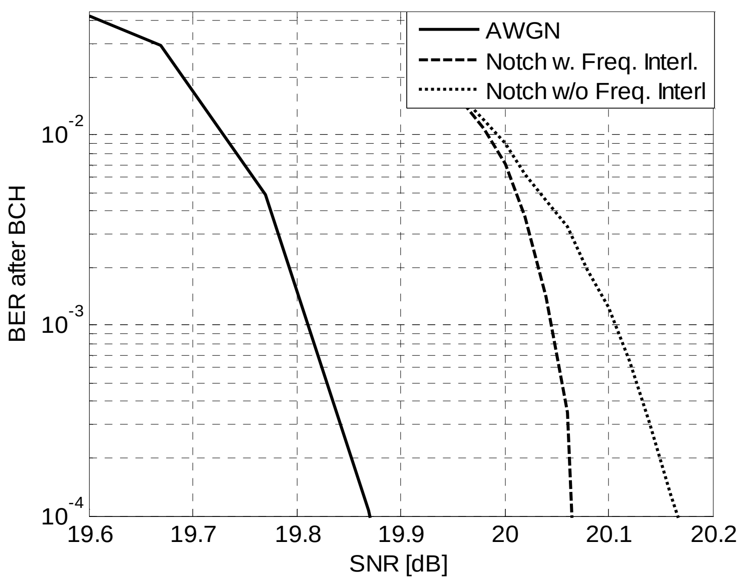

Figure 43a: Impact of a narrowband interferer (100 kHz bandwidth) on the SNR requirements of a DVB-C2 signal (8 MHz bandwidth), with and without frequency interleaving

- **SNR**

  First minimum requirements for the SNR for DVB-C2 are obtained from system calculations as shown in table 18 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.

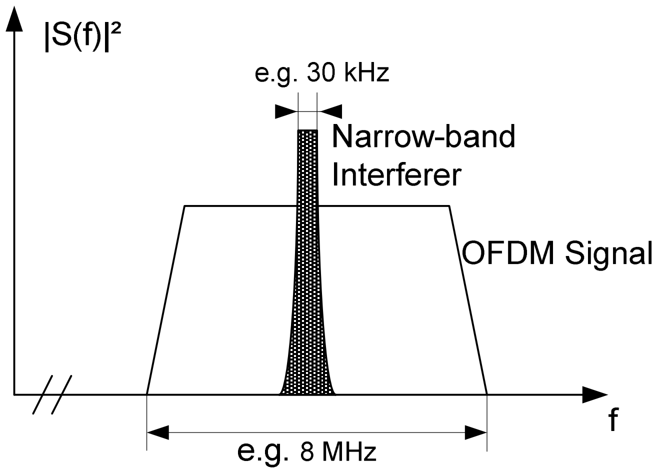

Figure 43b: DVB-C2 signal with narrow-band interferer (cluster beat), only few OFDM subcarriers will be affected

9.4 Network optimization

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.

Here we will summarize and explain the issues relevant in network optimization for DVB-C2 deployment. In particular we will touch upon the optimization of the DVB-C2 signal level.

9.4.1 The effect of the DVB-C2 carrier level

Figure 44 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 44 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).

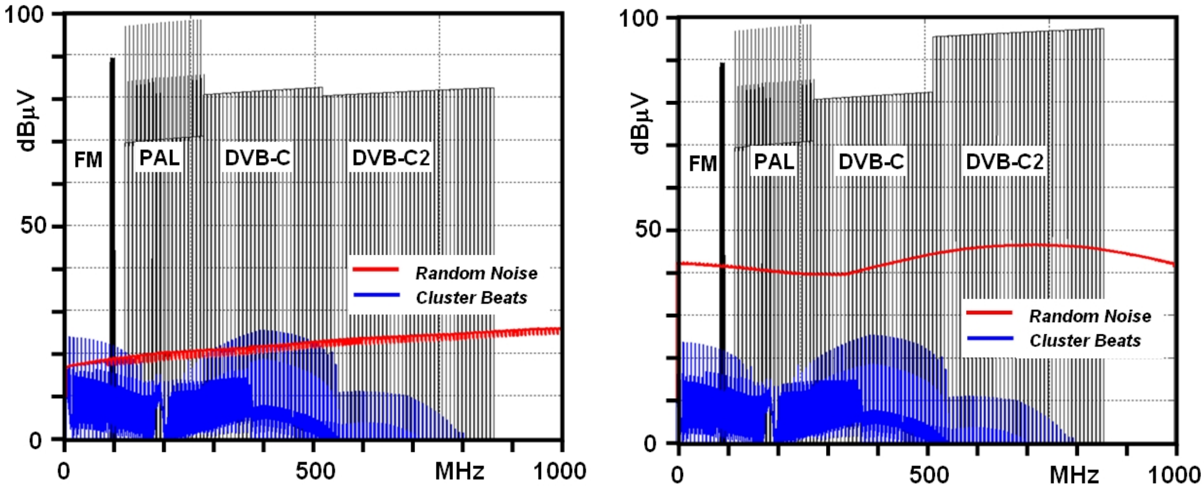

Figure 44: Full spectra at the output port of the distribution amplifier of a cascade

NOTE: Figure 44 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 44 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 45 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 45:

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 5th 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.

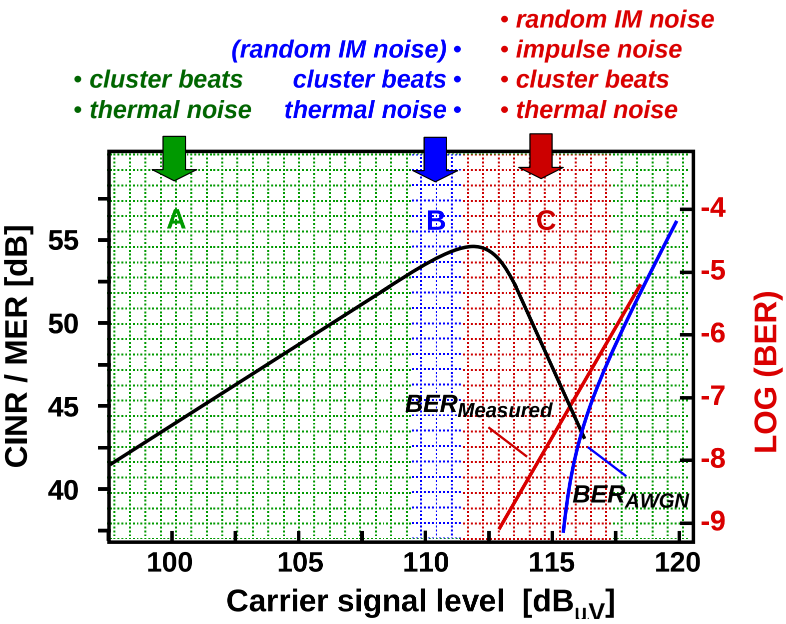

Figure 45: Schematic diagram explaining the nature of the quality of the signal of a DVB-C signal

NOTE: Figure 45 shows the schematical 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 must 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], paragraph 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 2nd and 3rd order effects are considered. However, analysis of degradation data strongly suggests that not the 2nd and 3rd order nonlinear behaviour degrades the digital signal, but 4th and 5th 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 5th 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 2nd or 3rd order dominance. Figure 46 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 2nd and 3rd 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 2nd and 3rd order intermodulation. An example of $CINR_{CSO}$ and $CINR_{CTB}$ curves and the occurrence of dominant 2nd and 3rd order degradation can be found in figure 47.

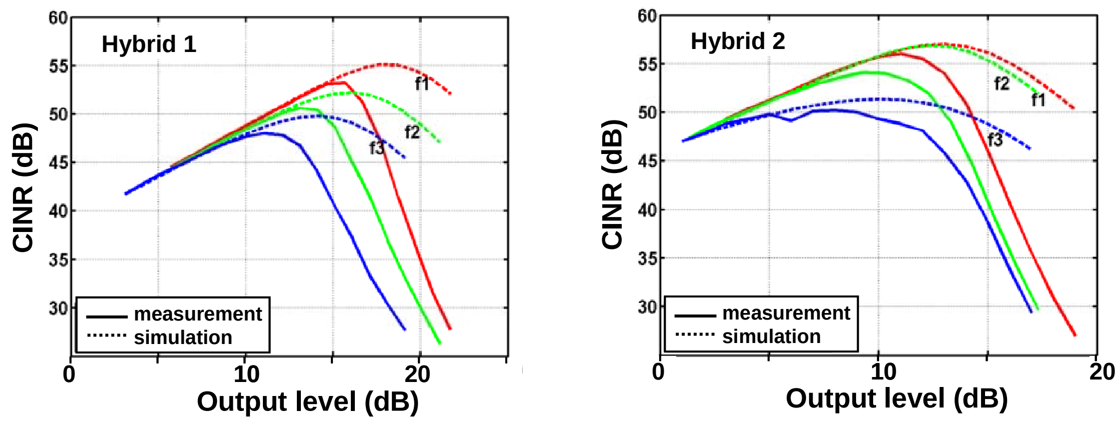

Figure 46: Measured and simulated CINR curves for 119 MHz (f1), 420 MHz (f2) and 855 MHz (f3)

NOTE 1: Figure 46 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 2nd and 3rd order component model is used. The simulated and measured curves clearly are not congruent, showing that 2nd and 3rd order intermodulation does not dominate the CINR curves.

3) The absence of visible or measureable 2nd and 3rd order degradation in case of digital carriers can be explained logically and straightforward, namely, as long as the 2nd and 3rd order intermodulation dominates the intermodulation distortion signal, these 2nd and 3rd 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 48.

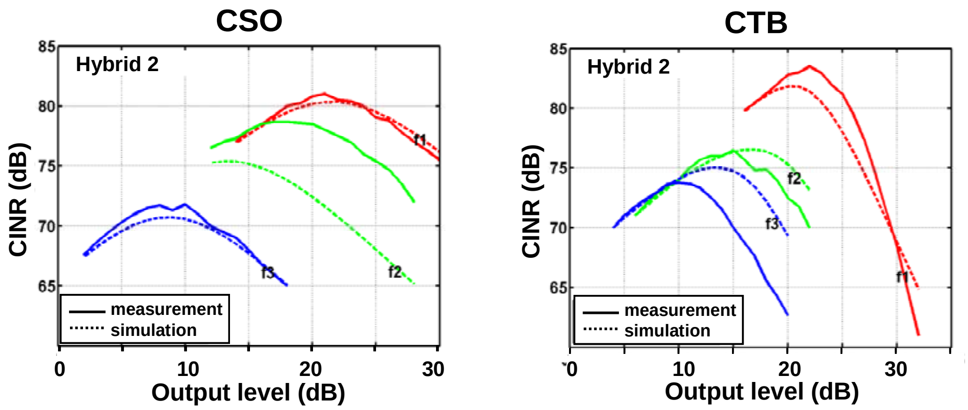

Figure 47: Measured and simulated CINR curves

NOTE 2: Figure 47 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 2nd and 3rd 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 2nd and 3rd order intermodulation dominates the CINR curves.

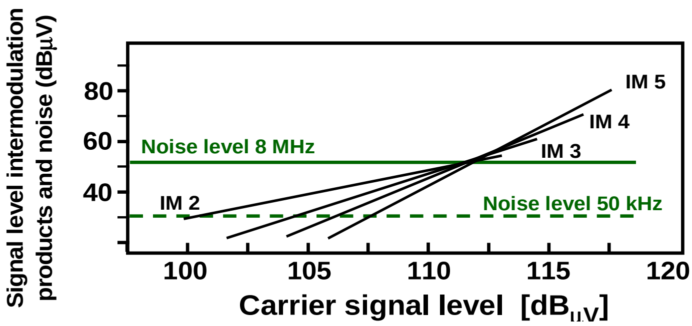

Figure 48: Schematical diagram of the signal level of the intermodulation products

NOTE 3: Figure 48 shows the s chematical 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 Synchronisation Procedure

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

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 49. Details on the different blocks are given in the following clauses.

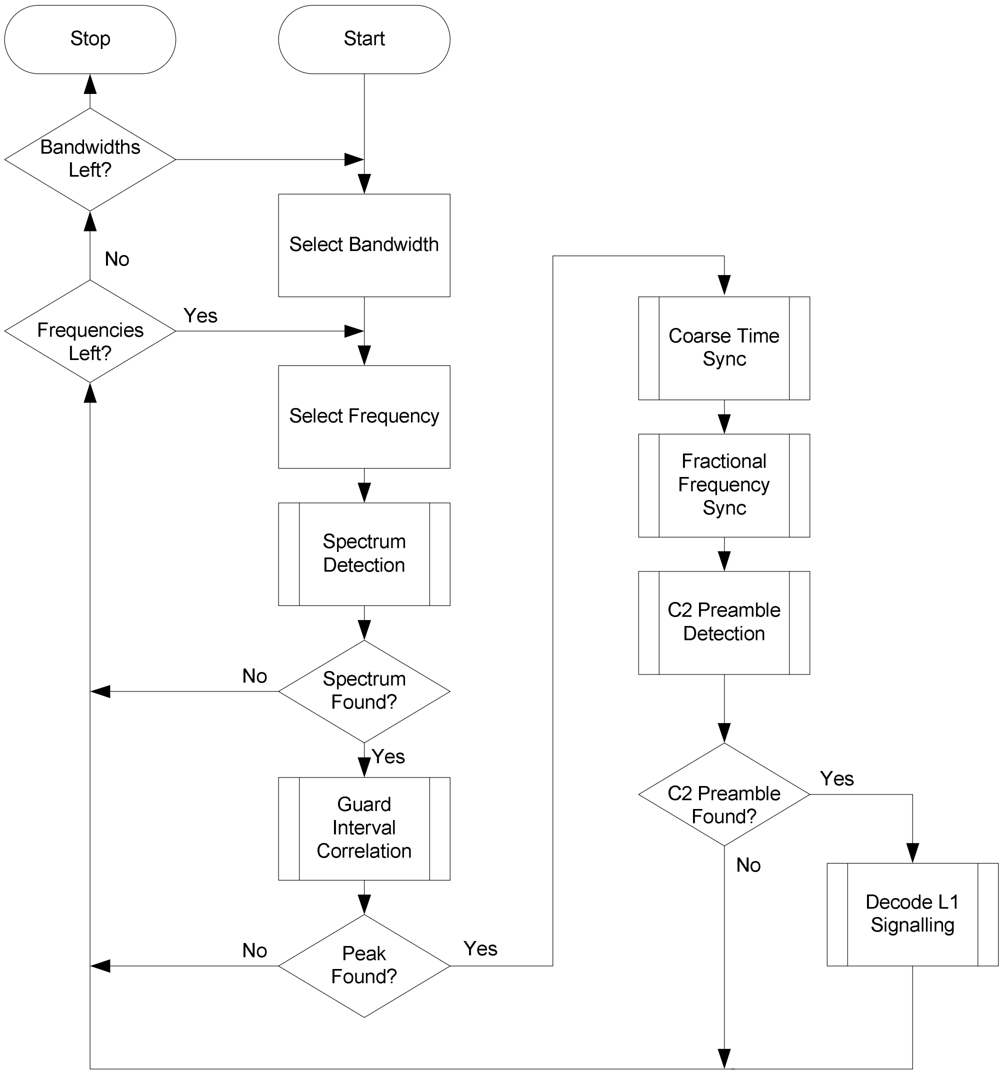

Figure 49: Initial acquisition flow chart

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 synchronise 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 part 2 signalling, the receiver must 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-part 2 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.

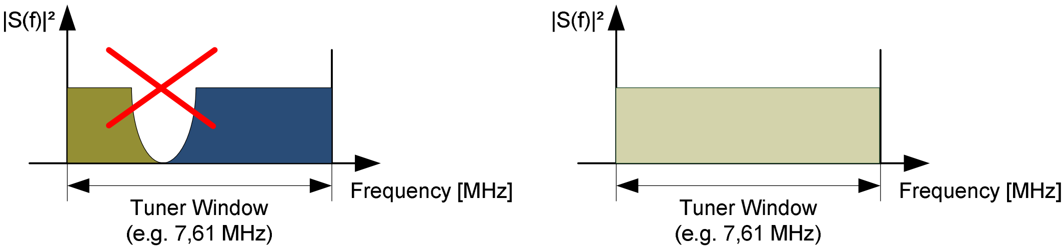

Figure 50: Principle of the spectrum detection, false spectrum (left hand side) and correct spectrum (right hand side)

One means to overcome this problem is the application of spectrum detection. Figure 50 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 synchronisation 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.

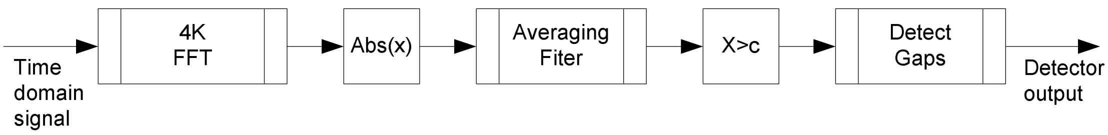

Figure 51: Possible implementation of the spectrum detector

A possible implementation of the spectrum detector is depicted in figure 51. 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 synchronisation algorithm are given in [i.12].

10.1.1.3 Coarse Time and Fractional Frequency Synchronisation

The coarse time and the fractional frequency synchronisation can also be achieved by means of the Guard Interval. For details see [i.12].

10.1.1.4 Preamble Detection and Synchronisation

Within this clause we assume that the temporal synchronization to the OFDM symbols as well as the fractional frequency offset compensation has already been achieved. Thus, we are able to demodulate the data modulated on each OFDM sub-carrier correctly, but we cannot perform channel estimation and we do not know the actual carrier index *k*. If we are able to detect the index *k* of the received OFDM sub-carriers, we are completely synchronized to the data stream. Also the channel estimation is possible in this case, as we know the transmitted values of the pilots.

The synchronization to the preamble is shown in figure 52. 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 D-BPSK demodulation between two OFDM sub-carriers that have a distance of $D_P=6$, i.e. the distance of the preamble pilots. If we assume that the channel conditions are nearly static between to pilots in the frequency direction, which is true due to the very short echoes in cable networks, the D-BSPK 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 D-BPSK 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.

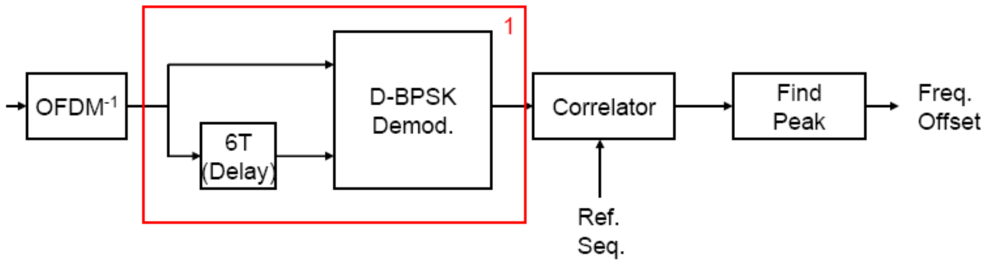

Figure 52: Synchronization to the preamble by means of the preamble pilots

10.1.1.5 Preamble Data Decoding Procedure

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 on 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 53.

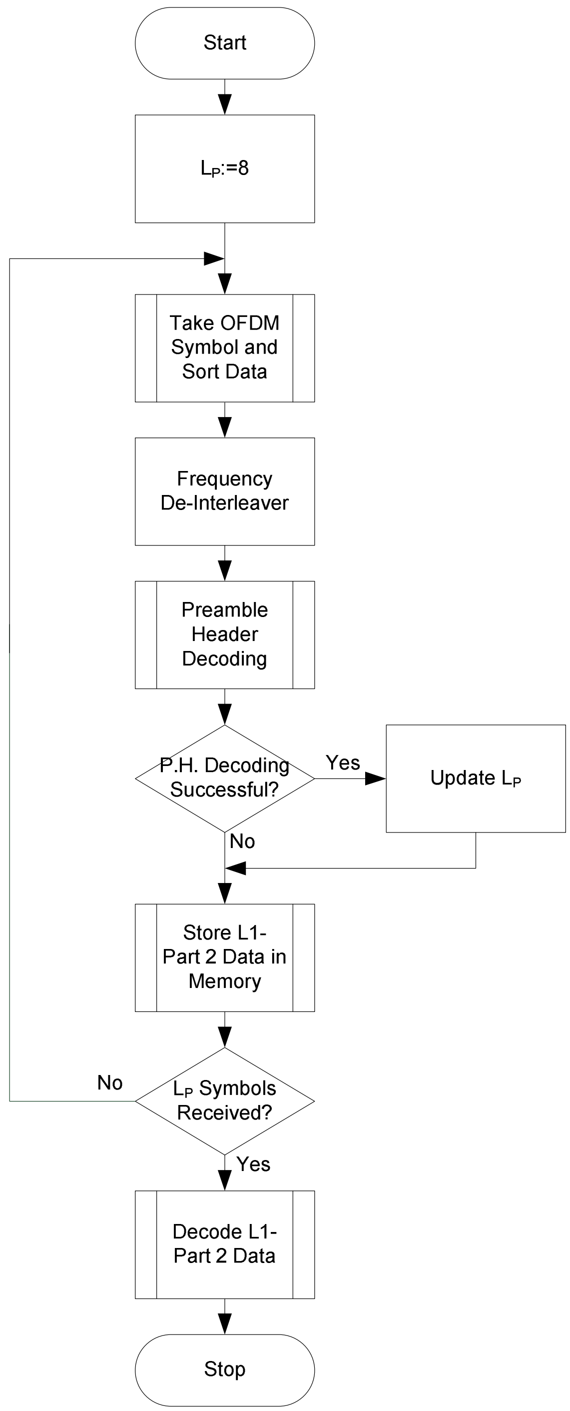

Figure 53: Preamble Decoding Procedure

Firstly, we assume that the receiver knows the position of the first preamble symbol. This is trivial if we were already synchronized, 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 we have tuned to the preamble recently, the receiver 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 we tune into the preamble and miss the first preamble symbol, the decoding procedure as depicted in figure 43 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, as especially 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 not 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 44 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 54.

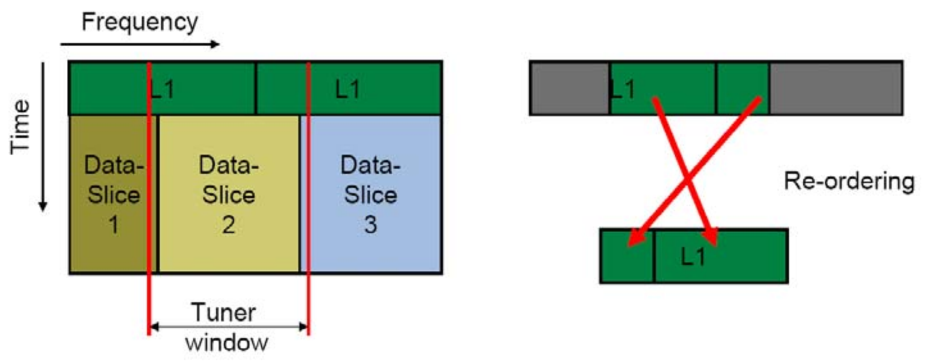

Figure 54: Recovery of a L1 Signalling Block

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 part 2 Signalling Blocks.

One additional aspect is the presence of notches within the preamble, which may be required if no power must 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 part 2 cells to fill the entire band of the L1 block, accumulation of those L1 part 2 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. This descriptor includes the OFDM parameters, the tuning frequency to obtain the Layer 1 - part 2 signalling information and the DVB-C2 System Id.

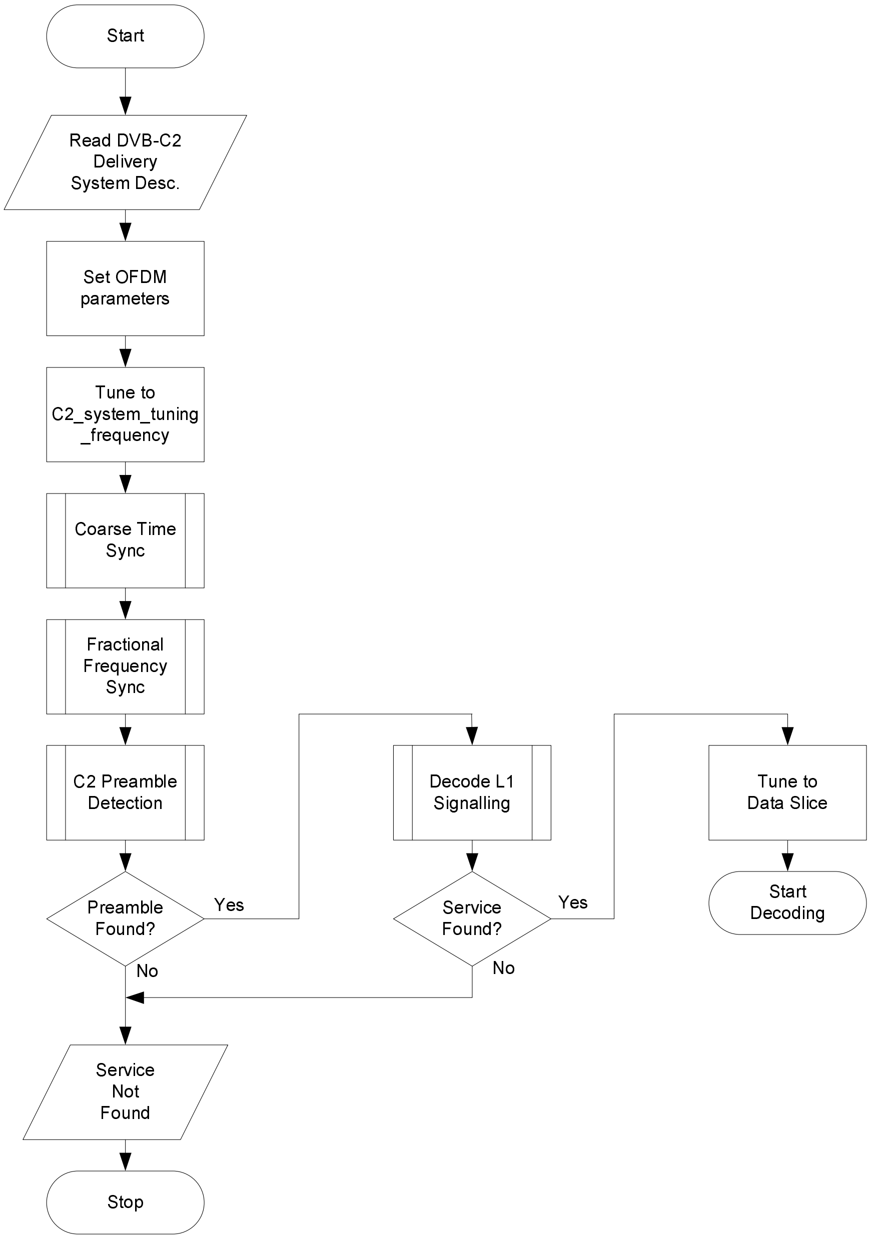

Figure 55: Channel tuning block diagram

Figure 55 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. 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 is able to calculate the tuning position out of the Layer 1 - part 2 signalling information, 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 56, which is a de-interleaving process as a counterpart to the time interleaving shown in figure 28 of [i.1].

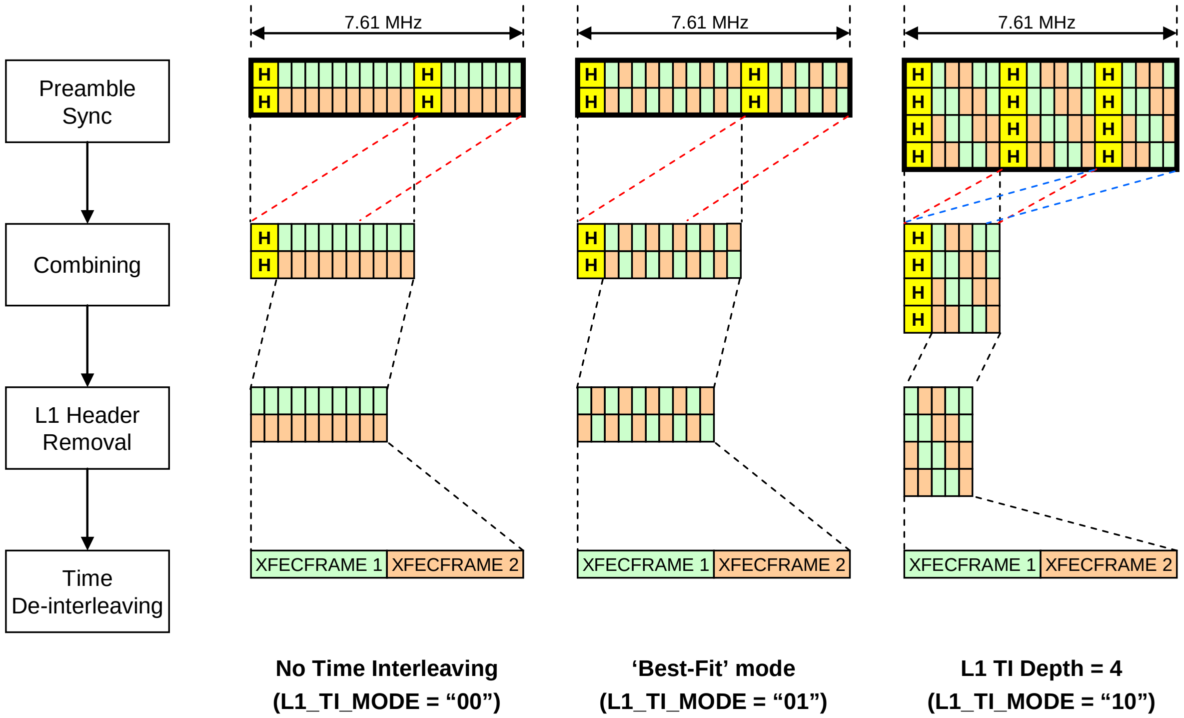

Figure 56 Time de-interleaving of L1 part 2 data

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 56. 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 56. Note that the interleaving depth may be different for the same amount of L1 part 2 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 (77) 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 part 2 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 time direction, the receiver can begin time de-interleaving after C2 Frame detection. The receiver first detects the preamble and decodes L1 part 2 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 57.

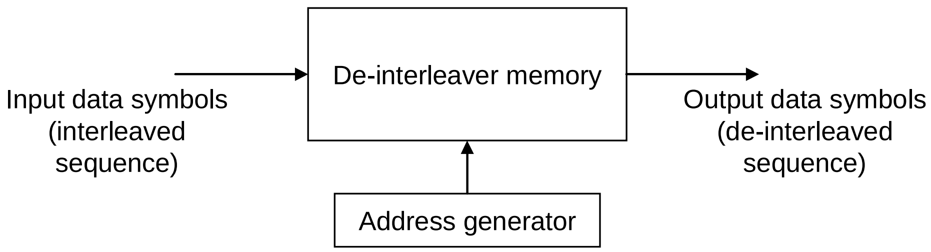

Figure 57: De-interleaver block diagram

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.

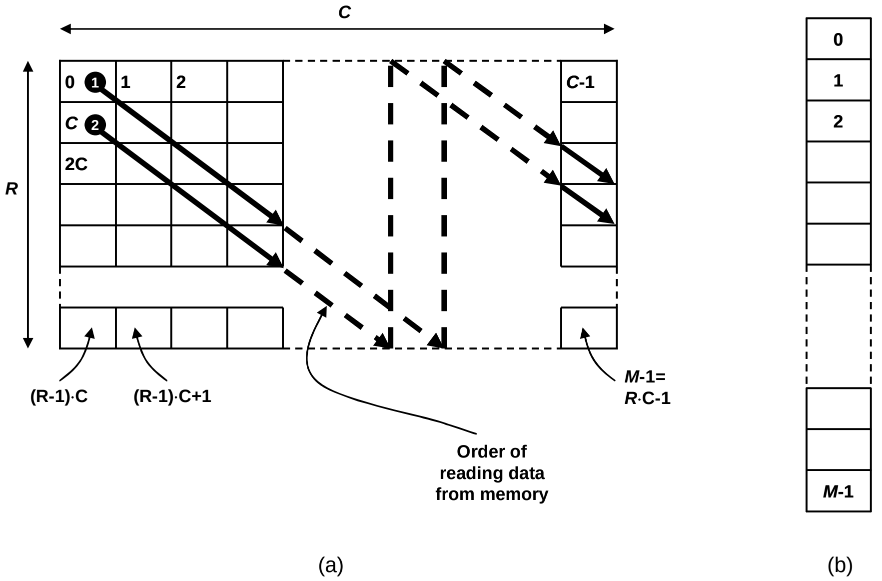

Figure 58: Implementation of de-interleaver memory

NOTE: Figure 58 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 58 (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:

$$M = R \times C \tag{72}$$

The actual block of memory to be used will be implemented as a sequential block of memory, as shown in figure 58 (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:

$$0 \le i \le M - 1 \tag{73}$$

The address generation can be better understood and explained by using 2-dimensional (2-D) memory shown in figure 58 (a). The addresses for sequential memory shown in figure 58 (b) can be easily calculated from the addresses of 2-D memory by simple conversion. 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 58 (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), …, (R-1, R-1), (R, 0), (R+1, 1), (R+2, 2), …, (C-1, (C *mod* R)-1)

This completes one cycle of reading the first line in diagonal direction in figure 3(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), …, (R-2, R-1), (R-1 0), (R, 1), (R+1, 2), …, (C-1, $C \, mod \, R$)

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:

$$\begin{aligned} c_{i,0} &= i \ mod \ C \\ s_{i,0} &= c_{i,0} \ mod \ R \\ r_{i,0} &= [s_{i,0} + (i \ div \ C)] \ mod \ R \end{aligned} \tag{74}$$

The term ($i \ div \ C$) can be interpreted as the WRITE address for data input into the ($i \ div \ C$)-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 (74) will be:

$$\begin{aligned} c_{i,j} &= i \ mod \ C \\ s_{i,j} &= (j \times c_{i,j}) \ mod \ R \\ r_{i,j} &= [s_{i,j} + (i \ div \ C)] \ mod \ R \end{aligned} \tag{75}$$

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 58 (b), we can convert the coordinate address ($c_{i,j}$, $r_{i,j}$) of 2-D memory into the linear address for the i-th data of j-th TI block by:

$$L_{i,j} = C \cdot r_{i,j} + c_{i,j} \tag{76}$$

The corresponding linear address for the 2-D memory is shown in figure 3 (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 outputs 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:

$$\begin{aligned} & for \ (i = 0; i < M; i = i + 1) \ \{ \\ & \quad Generate \ Address \ L_{i,j}; \\ & \quad if \ (c_{i,j}, r_{i,j}) = Data \ Cell \ Address \\ & \qquad Read / Write \ Cell \ at \ L_{i,j}; \\ & \} \end{aligned} \tag{77}$$

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 equalised payload cells of a given Data Slice from one OFDM symbol to the next. The number of payload cells per Data Slice per OFDM symbol (N_{DS}) can vary from symbol to 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*, *symbol number*, *L1 info*) exists in the demodulator to provide N_{DS} for a given Data Slice number, symbol number and from the Layer 1 information. The frequency de-interleaver must 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 must be capable of dealing with N_{max} payload cells where:

$$N_{max} = (K_{DS,max} - K_{DS,min}) - N_{SP,Dx=24} - N_{CP} \quad (78)$$

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 de-interleaver memory is split into two banks: Bank A for even OFDM symbols and Bank B for odd OFDM 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 equalised payload cells from even OFDM symbols (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, equalised payload cells from odd OFDM symbols (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 $H_{[0,1]}(q)$ are provided by the pseudo-random address generator from clause 9.4.5 of [i.1]. In order to produce a continuous stream of cells at the de-interleaver output, when Bank A is being written (incoming even symbol), Bank B is also being read (outgoing previous odd 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 $H_{[0,1]}(q)$ which provide the write addresses. If all symbols in the Data Slice contained $N_{max} = C_{data}$ payload cells, then the number of write addresses for symbol number 2n+1 must match the number of read addresses for symbol number 2n otherwise some data cells of symbol 2n will be skipped. Unfortunately N_{DS} can be different from symbol to 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*, *symbol number*, *L1 info*) returns the number of payload cells in the current slice for the given 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 symbol number 2n:

1) $q = 0$;

2) C_{max} = **max**(**DataCells**(*slice number* , 2*n-1*, *L1 info*), **DataCells**(*slice number* , 2*n*, *L1 info*));

3) Generate address $H_0(q)$;

4) rdEnable = (q < **DataCells**(*slice number* , 2*n-1*, *L1 info*));

5) wrEnable = ($H_0(q)$ < **DataCells**(*slice number* , 2*n*, *L1 info*));

6) if (rdEnable) Read cell q of output de-interleaved symbol $2n$ - 1 from location q of memory Bank B;

7) Store cell q of incoming interleaved 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 ($q < C_{\max}$) goto 3.

Then with symbol $2n$+1 at the input of the de-interleaver:

1) $q = 0$;

2) C_{max} = **max**(**DataCells**(*slice number* , $2n$, *L1 info*), **DataCells**(*slice number* , $2n+1$, *L1 info*));

3) Generate address $H_1(q)$;

4) rdEnable = (q < **DataCells**(*slice number* , $2n$, *L1 info*));

5) wrEnable = ($H_1(q)$< **DataCells**(*slice number* , $2n+1$, *L1 info*));

6) if (rdEnable) Read cell q of output de-interleaved symbol $2n$ from location q of memory Bank A;

7) Store cell q of incoming interleaved 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 ($q < C_{\max}$) goto 3.

The required width for each memory location depends on the resolution with which each cell is represented after channel equalisation. 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 synchronisation.
- Integer frequency offset estimation.
- Channel estimation.
- Common Phase Error (CPE) estimation.

The C2-frame synchronisation 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

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 centered 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: CPE and ICI.

Low-frequency noise gives rise to common phase error (CPE). CPE 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 standard includes continual pilots which, amongst other uses, can be used for the purpose of cancelling CPE, according to the method described in clause 10.5.1. Taking this and other possible cancellation measures into account, the impact of CPE 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 (CPE). Since the random rotation of CPE 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 CPE 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 CPE 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 CPE.

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 CPE which is being corrected, otherwise the correction process will an 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 CPE-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 CPE correction for preamble and data symbols if the CPE 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 synchronisation is maintained, the received (complex) carrier amplitudes can be given by:

$$Y_{k,l} = H_{k,l} X_{k,l} + N_{k,l} \tag{79}$$

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 equalisation: 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 we have noted: weakly-received carriers have a poorer SNR than others.

Dividing by the estimated channel response is somewhat similar to a zero-forcing equaliser, and would normally be deprecated on the grounds of aggravating the effects of noise. However, under the assumption that coded OFDM is being used, we 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 we use 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. We calculate the measured channel estimate for this cell as:

$$H'_{k,l} = \frac{Y_{k,l}}{X_{k,l}} = H_{k,l} + \frac{N_{k,l}}{X_{k,l}} \tag{80}$$

In principle this can be done for any cell where we know what information $X_{k,l}$ has been transmitted. 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 [i.1] defines the SP patterns of DVB-C2, and introduces terms D_X and D_Y to characterise 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, then we will have either a measurement or an interpolated estimate of the channel response for every cell on the pilot-bearing carriers. 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:

$$\frac{1}{D_Y(T_U + T_G)} \cdot \frac{T_U}{D_X} = \frac{1}{D_X D_Y (1 + GIF)} \tag{81}$$

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 equalisation 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 realise 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 minimising delay, will use at most simple linear temporal interpolation. To begin with we will consider simplest case of interpolation between scattered pilots.

In DVB-C2 both of two scattered pilot patterns have Dy=4. It follows that simply to perform simple linear interpolation requires three (Dy-1) symbols' worth of storage for the main data stream.

Although it is unlikely to have any Doppler in cable channel environment, still we could see some channel estimation variation 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 4096-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 isn't 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 must 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 is inside the C2 signal). As soon as the receiver is able to recognize the preamble it can start to compensate offsets and to extract the L1 part 2 signalling within the preamble symbols. The L1 part 2 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 part 2 information block repetition rate (3 408 subcarriers). Therefore the L1 part 2 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 part 2 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.

Beside the DSLICE_TUNE_POS the start and stop carriers of a Data Slice are given in the L1 part 2 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:

GUARD_INTERVAL:	00	Guard Interval is 1/128
NUM_BUNDLED_CH:	00001	The C2 signal width is 7,61 MHz
START_FREQUENCY:	0000000000001100000010000	Start Frequency is 330 MHz = subcarrier 330E6*448 usec = 147 840 24 carriers granularity -> 6 160 = 0x1 810 = 1100000010000
DSLICE_TUNE_POS:	0000001000111	The tuning position of this Data Slice is 1 704th carrier frequency of this C2 System 24 carriers granularity -> 71 = 0x47 = 0000001000111
DSLICE_OFFSET_LEFT:	10111001	The left edge of this Data Slice is start frequency (apart from tuning position as much as 1 704 carrier spacing)

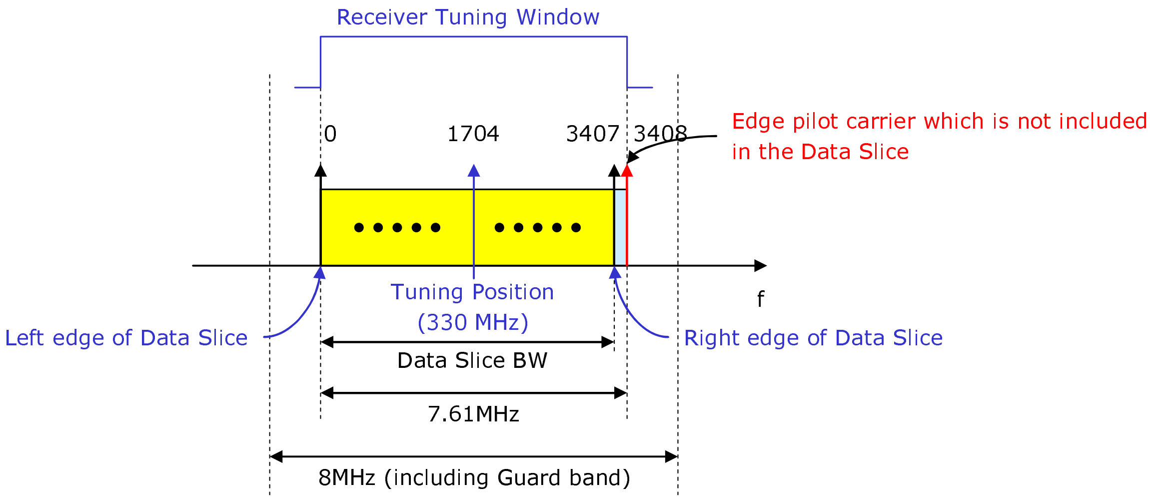

Figure 59: Relation between Data Slice and tuning window positioning

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*8 MHz - 0,39 MHz)

START_FREQUENCY: 000000000001100000010000
Start Frequency is 330 MHz = subcarrier 330E6*448 usec = 147 840
24 carriers granularity -> 6 160 = 0x1 810 = 1100000010000

DSLICE_TUNE_POS: 0000001000111 The tuning position of this Data Slice is 1 704th 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

Discussion of the implementation of receiver buffers and its appropriate management are subject to the next version 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. Here, we take robust FECFrame header as an example. The FECFrame header detection can be performed by the following steps:

1) Assume that the 32-symbol complex sequence ($s_0, s_1,\ldots, s_{31}$) = ($r_i, r_{i+1},\ldots, r_{i+31}$) is the robust FECFrame header and demodulate them into a 64-bit sequence ($a_0, a_1,\ldots, a_{63}$) by a QPSK demapper. The complex symbol, r_i is a received data symbol. Note that the notation of ($s_0, s_1,\ldots, s_{31}$) = ($r_i, r_{i+1},\ldots, r_{i+31}$) 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 $\tilde{\mathbf{w}}^{RM} = (\tilde{w}_0^{RM}, \tilde{w}_1^{RM}, \ldots, \tilde{w}_{31}^{RM})$ by $\tilde{w}_{(k+2)_{32}}^{RM} = a_{2k} \oplus a_{(2k+5)_{64}}$ where $(x)_y$ is the result of x modulo y.

3) Compute the binary correlation of $\tilde{\mathbf{w}}^{RM}$ and $\mathbf{w}^{\mathrm{RM}}$ by $C_p = \sum_{k=0}^{31}(2\cdot\tilde{w}_k^{RM} - 1)(2\cdot w_k^{RM} - 1)$.

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 $\tilde{\boldsymbol{\lambda}} = (\tilde{\lambda}_0, \tilde{\lambda}_1, \ldots, \tilde{\lambda}_{31})$ is decoded by

$$\tilde{\lambda}_k = \begin{cases} 0, \ \mathrm{Re}(s_k) + \mathrm{Im}(s_{(k+2)_{32}})\cdot(1-2w_{(k+2)_{32}}^{RM}) \geq 0 \\ 1, \ \mathrm{Re}(s_k) + \mathrm{Im}(s_{(k+2)_{32}})\cdot(1-2w_{(k+2)_{32}}^{RM}) < 0 \end{cases} \quad (82)$$

5) The estimated 32-bit RM codeword $\tilde{\boldsymbol{\lambda}} = (\tilde{\lambda}_0, \tilde{\lambda}_1, \ldots, \tilde{\lambda}_{31})$ can be decoded by a 3-stage majority-logic decoding. The last 10 bits, $(b_6, b_7, \ldots, b_{15})$, are decoded from the received code vector $\tilde{\boldsymbol{\lambda}} = (\tilde{\lambda}_0, \tilde{\lambda}_1, \ldots, \tilde{\lambda}_{31})$ in the first stage. These 10 bits are removed from $\tilde{\boldsymbol{\lambda}}$ to form a modified code vector $\tilde{\boldsymbol{\lambda}}^{(1)} = \tilde{\boldsymbol{\lambda}} - (0,0,\ldots,0,\tilde{b}_6,\tilde{b}_7,\ldots,\tilde{b}_{15})\cdot\mathbf{G}$.

6) The modified code vector $\tilde{\boldsymbol{\lambda}}^{(1)}$ has a symmetric structure and it can be used to double- check if the 32-symbol complex sequence ($s_0, s_1,\ldots, s_{31}$) is the FECFrame header. The RM autocorrelation of the received modified code vector is computed by

$$R_{\mathrm{RM}}(k) = \sum_{m=0}^{2^k-1}\sum_{n=0}^{2^{4-k}-1}(2\cdot\tilde{\lambda}_{m\cdot 2^{5-k}+n}^{(1)} - 1)\cdot(2\cdot\tilde{\lambda}_{m\cdot 2^{5-k}+2^{4-k}+n}^{(1)} - 1) \quad (83)$$

The RM symmetry measure is then computed by $C_{RM} = \sum_{k=0}^{4}\left|R_{\mathrm{RM}}(k)\right|^2$.

6.1) If $C_{\mathrm{RM}} < T_2$, go to step 1 and advance symbol index by 1, *e.g.*, $i=i+1$.

6.2) If $C_{\mathrm{RM}} \geq T_2$, 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 $\tilde{\boldsymbol{\lambda}} = (\tilde{\lambda}_0, \tilde{\lambda}_1, ..., \tilde{\lambda}_{31})$ in step 4. The estimated 32-bit RM codeword $\tilde{\boldsymbol{\lambda}} = (\tilde{\lambda}_0, \tilde{\lambda}_1, ..., \tilde{\lambda}_{31})$ is decoded by computing $k = 0,1,\ldots,15$:

$$x_k^{+1} = \exp\left[\frac{-\left(\mathrm{Re}(s_k) - 1/\sqrt{10}\right)^2}{\sigma^2}\right], \quad x_k^{+3} = \exp\left[\frac{-\left(\mathrm{Re}(s_k) - 3/\sqrt{10}\right)^2}{\sigma^2}\right]$$

$$x_k^{-1} = \exp\left[\frac{-\left(\mathrm{Re}(s_k) + 1/\sqrt{10}\right)^2}{\sigma^2}\right] \quad x_k^{-3} = \exp\left[\frac{-\left(\mathrm{Re}(s_k) + 3/\sqrt{10}\right)^2}{\sigma^2}\right] \tag{84}$$

$$y_k^{+1} = \exp\left[\frac{-\left(\mathrm{Im}(s_k) - 1/\sqrt{10}\right)^2}{\sigma^2}\right] \quad y_k^{-1} = \exp\left[\frac{-\left(\mathrm{Im}(s_k) + 1/\sqrt{10}\right)^2}{\sigma^2}\right]$$

$$y_k^{-3} = \exp\left[\frac{-\left(\mathrm{Im}(s_k) + 3/\sqrt{10}\right)^2}{\sigma^2}\right] \quad y_k^{+3} = \exp\left[\frac{-\left(\mathrm{Im}(s_k) - 3/\sqrt{10}\right)^2}{\sigma^2}\right] \tag{85}$$

$$\tilde{\lambda}_{2k} = \begin{cases} 0, \log\dfrac{x_k^{-1} + x_k^{-3}}{x_k^{+1} + x_k^{+3}} + (1 - 2w_{(2k+2)_{32}}^{RM}) \cdot \log\dfrac{x_{(k+1)_{16}}^{-1} + x_{(k+1)_{16}}^{+1}}{x_{(k+1)_{16}}^{-3} + x_{(k+1)_{16}}^{+3}} < 0 \\ 1, \log\dfrac{x_k^{-1} + x_k^{-3}}{x_k^{+1} + x_k^{+3}} + (1 - 2w_{(2k+2)_{32}}^{RM}) \cdot \log\dfrac{x_{(k+1)_{16}}^{-1} + x_{(k+1)_{16}}^{+1}}{x_{(k+1)_{16}}^{-3} + x_{(k+1)_{16}}^{+3}} \geq 0 \end{cases} \tag{86}$$

$$\tilde{\lambda}_{2k+1} = \begin{cases} 0, \log\dfrac{y_k^{-1} + y_k^{-3}}{y_k^{+1} + y_k^{+3}} + (1 - 2w_{(2k+3)_{32}}^{RM}) \cdot \log\dfrac{y_{(k+1)_{16}}^{-1} + y_{(k+1)_{16}}^{+1}}{y_{(k+1)_{16}}^{-3} + y_{(k+1)_{16}}^{+3}} < 0 \\ 1, \log\dfrac{y_k^{-1} + y_k^{-3}}{y_k^{+1} + y_k^{+3}} + (1 - 2w_{(2k+3)_{32}}^{RM}) \cdot \log\dfrac{y_{(k+1)_{16}}^{-1} + y_{(k+1)_{16}}^{+1}}{y_{(k+1)_{16}}^{-3} + y_{(k+1)_{16}}^{+3}} \geq 0 \end{cases} \tag{87}$$

where σ^2 is the estimated variance of noise.

2) The modified code vector $\tilde{\boldsymbol{\lambda}}^{(1)} = \tilde{\boldsymbol{\lambda}} - (0,0,...,0,\tilde{b}_6,\tilde{b}_7,..,\tilde{b}_{15}) \cdot \mathbf{G}$ has a symmetric structure because for the transmitted 32-bit RM code vector $\boldsymbol{\lambda}$, $\boldsymbol{\lambda}^{(1)}$ 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 synchronisation 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 60 shows the synchronisation to the headers for the robust mode. As the equalisation 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.

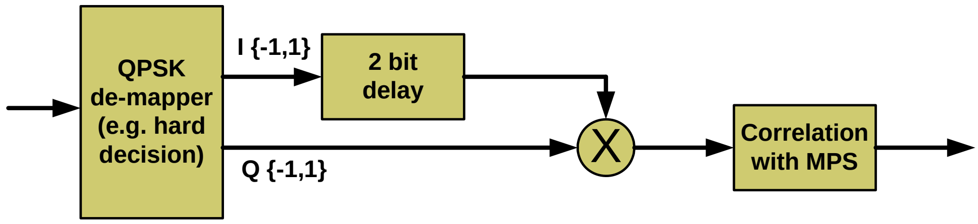

Figure 60: Synchronisation scheme for robust header

After the synchronisation to the header is reached, the signalling data has to be decoded. One approach is shown in figure 61. 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.

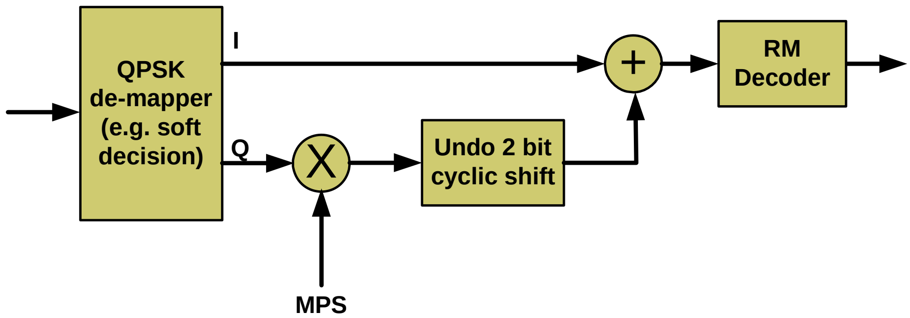

Figure 61: Decoding of the signalling data

Whilst the decoding of the decoding of the Data Slice Packet Headers is required for every Data Slice Packet, the synchronisation 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 [i.1], as this allows for the calculation of the actual XFECFRAME length.

Figure 62: 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 63 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 1024-QAM, respectively.

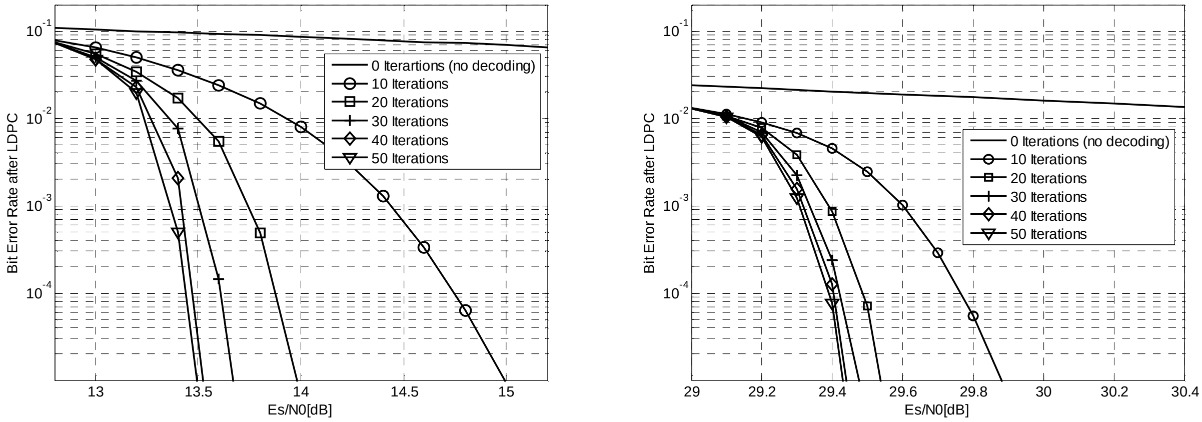

Figure 63: Performance sensitivity for rate 2/3 (N_{ldpc}=64 800) at 64-QAM (left) and for rate 9/10 (N_{ldpc}=64 800) at 1024-QAM (right) on AWGN channel

The number of decoding iterations per FECFRAME achievable at the receiver for a given hardware complexity can be estimated from the following equation:

$$I = (N_{ldpc} \cdot F_{clk} \cdot P_{dec}) / (C_{code} \cdot E_{mat} \cdot a_{dec}) \quad (88)$$

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: 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 PLPs (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 and 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 errored 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:

$$R_{TS} = \frac{N_{bits}}{(ISCR_2 - ISCR_1) \times T} \tag{89}$$

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.1].

If ISSY is not used, the receiver should use a self-balancing buffer as described in annex G of [i.1]. 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.1].

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.1]. 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.1] 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.1].

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.

10.11.1.7 Re-combining data of bundled PLPs

As specified in annex F in [i.1], DVB-C2 allows optional bundling of PLPs carried in different Data Slices. Annex F in [i.1] mentions that in this case all data packets of a bundled PLP connection shall 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 [i.1], shows the block diagram for the PLP Bundling operation mode.

A DVB-C2 demodulator, which supports the PLP bundling mode, shall therefore apply the inverse signal processing. All data packet marked to be bundled and the targeted PLP-ID shall be demodulated and FEC processed. Data packets of the targeted PLP shall be multiplexed in the order given by the ISSY timestamp. As the BUFSTAT field is not available in case of PLP bundling, the demodulator shall recover the output clock on basis of transmitted ISCR field and has to establish an appropriate output buffer management, which does not require the BUFSTAT information.

More details about the implementation of the bundled PLP mode will be discussed in the next version of the present document.

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.

Furthermore, reducing the Data Slice bandwidth can also reduce the power consumption as this reduces the data rate of all decoding stages within a chip.

11 Theoretical Performance

This clause presents simulation results indicating the performance of DVB-C2 in a range of cable channels. The results are based on the simulation and validation effort carried out during and following the development of the DVB-C2 standard [i.1].

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

During the development of the DVB-C2 specification [i.1], 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

$$\mathrm{y}(t) = k \sum_{i=1}^{N} \rho_i \, \mathrm{e}^{-j\theta_i} \, \mathrm{x}(t - \tau_i) \qquad (90)$$

where

- $\mathrm{x}(t)$ any $\mathrm{y}(t)$ are input and output signals, respectively,
- $k = \dfrac{1}{\sqrt{\sum_{i=1}^{N} \rho_i^2}}$ and
- the values of the relative power ρ_i, the delay τ_i and the phase θ_i are given in table 17.

Table 17: Relative power, delay and phase values of the two cases of the echo model

	Power [dB]	Delay [ns]	Phase [rad]
Case 1	-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
Case 2	-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

Table 18 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 1000 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 synchronisation 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.1].

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.3] 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 18: Required raw $(C/N)_0$ to achieve a given bit error rate of 10^{-4}
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]
LDPC 64 800 (long)	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
	1024	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
	4096	5	6			9,97	32,20	32,60	33,20
		9	10			10,78	34,90	35,40	36,30
LDPC 16 200 (short)	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
	1024	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
	4096	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 64 shows the results given in table 18.

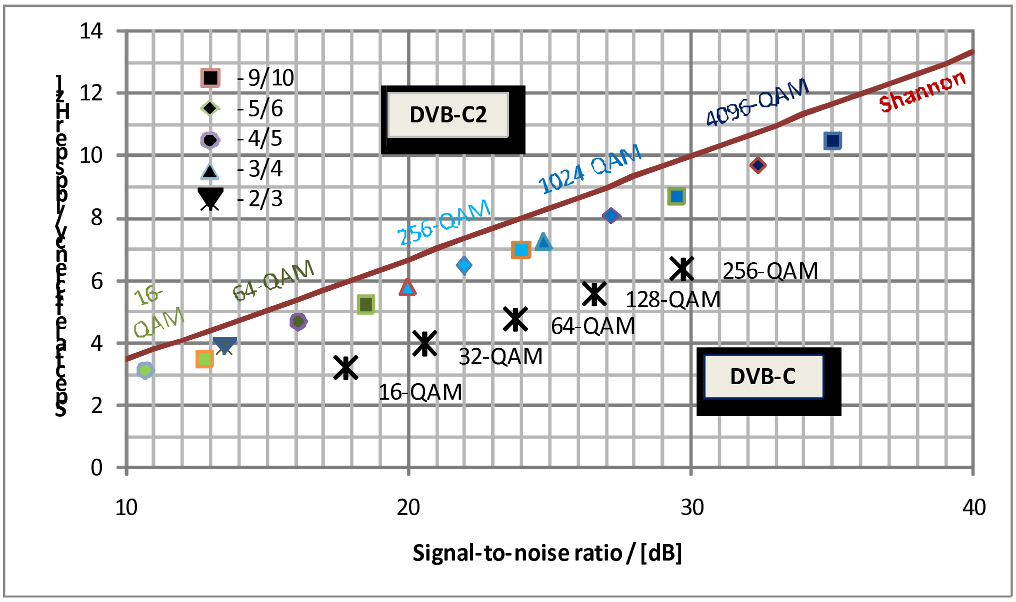

Figure 64: Graphical presentation of the simulation results according to table 18

11.2.1 Performance of L1 signalling part 2 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 65 shows the performance of the coded L1 signalling part 2 for a range of BCH information sizes. As shown in figure 65, 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 signalling part 2 vary as a function of the size of BCH information, as shown in figure 66.

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.

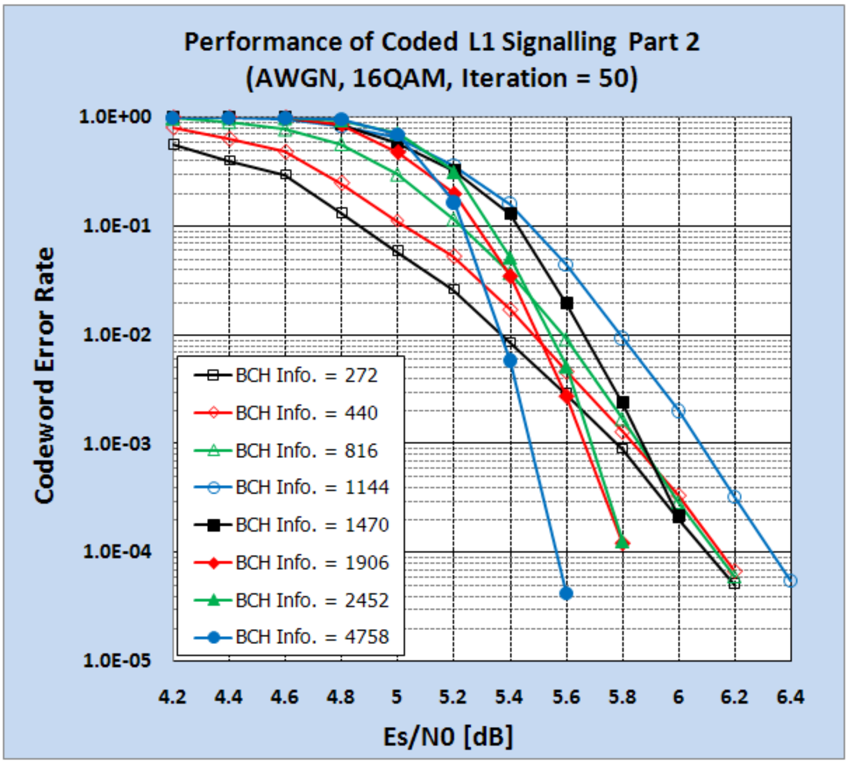

Figure 65: Performance of coded L1 signalling part 2

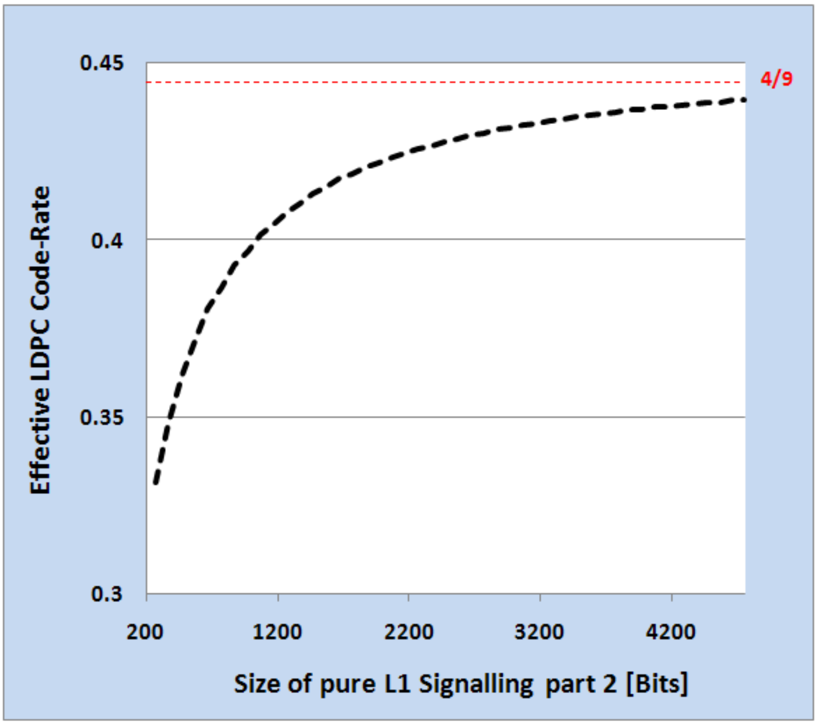

Figure 66: Effective LDPC code-rates for L1 signalling part 2

11.2.2 Correction values for pilot boosting

The values of required $(C/N)_0$ given in table 18 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 Δ_{BP} such that:

$$\frac{C}{N} = \left(\frac{C}{N}\right)_0 + \Delta_{\mathrm{BP}} \tag{91}$$

This correction factor is calculated by the following formula

$$\Delta_{\mathrm{BP}} = 10 \cdot \log_{10}\left(\frac{K_{\mathrm{data}} + K_{\mathrm{CP}} \cdot A_{\mathrm{CP}}^2 + K_{\mathrm{SP}} \cdot A_{\mathrm{SP}}^2}{K_{\mathrm{data}} + K_{\mathrm{CP}} + K_{\mathrm{SP}}}\right) \tag{92}$$

Where:

K_{data} Number of data cells per OFDM symbol;

K_{CP} Number of Continual Pilots per OFDM symbol;

K_{SP} Number of Continual Pilots per OFDM symbol;

A_{CP} Amplitude of the continual pilot cells;

A_{SP} Amplitude of the scattered pilot cells.

The exact value of Δ_{BP} depends on the number of pilots, but will not exceed 0,5 dB.

12 Examples of Possible Use of the System

The examples described below present practical implementation scenarios which should demonstrate the plurality of functionalities the DVB-C2 standard [i.1] 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

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 headends

This clause describes methods of signal processing, which are typically implemented in regional cable headends 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 headend 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 19 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. 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 44 Mbit/s. For the individual satellite signal 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 19: 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		Data Slices	6	
Transponder bit rate	4*·42,66	2*49,39	Slice bit rate (Mbit/s)	4*42,66	2*49,39
			Total C2 channel bit rate (Mbit/s)	5·42*66 + 1·49,39 = 269,44	
Transponder bandwidth (MHz)	5·27	33	C2 channel bandwidth (MHz)	32	
Modulation	8PSK 3 bit/symbol	QPSK 2 bit/symbol	Modulation (CCM)	1024-QAM 10 bit/symbol	1024-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:

$$N_c = \frac{R_{Sat}}{\Delta f * N * G * C1 * C2 * SP * CP} \tag{93}$$

- $\boldsymbol{R_{Sat}}$ = input transport stream data rate, in the example above 4 DVB-S2 transponders with 42,68 Mbit/s and two DVB-S2 transponders with 49,39 Mbit/s. According to [i.5] R_{Sat} is calculated in the formulas shown below;

  $\boldsymbol{R_{Sat}}$ = Symbolrate*M*LDPC-code rate*BCH-code rate*pilot loss (if used) from DVB-S2 transponder (94);

  $\boldsymbol{R_{Sat}}$ = 22MBaus*3*2/3*(43 040/43 200)*0,973 = 42,66 Mbit/s in case of the DVB-S2/8PSK transponders in table 1 in **[**i.5];

  $\boldsymbol{R_{Sat}}$ = 27,5MS/s*2*9/10*(58 192/58 320) = 49.39Mbit/s in case of the DVB-S2/QPSK transponders in table 1 in **[**i.5];

  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).

- $\triangle \boldsymbol{f}$: Carrier spacing (1/448 us for 8 MHz raster);
- ***N***: bits/s/Hz, in the example 10bit/s/Hz for 1024-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)*GI, where G=128/129 for GI=1/128;
- ***SP*** = (1-PP)*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 (93) and the values mentioned above the number of carriers N_c is:

$$N_c = \frac{42.7\text{Mbit/s} * 448\text{us}}{10 * (128/129) * (9/10) * (58128/58320) * (95/96) * 0.99 * (448/449)} = 2208 \qquad (94)$$

and N_c = 2 544 carriers for R_{Sat} = 49,39Mbit/s.

If the Guard Interval is equal to1/128 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 269,4 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 67.

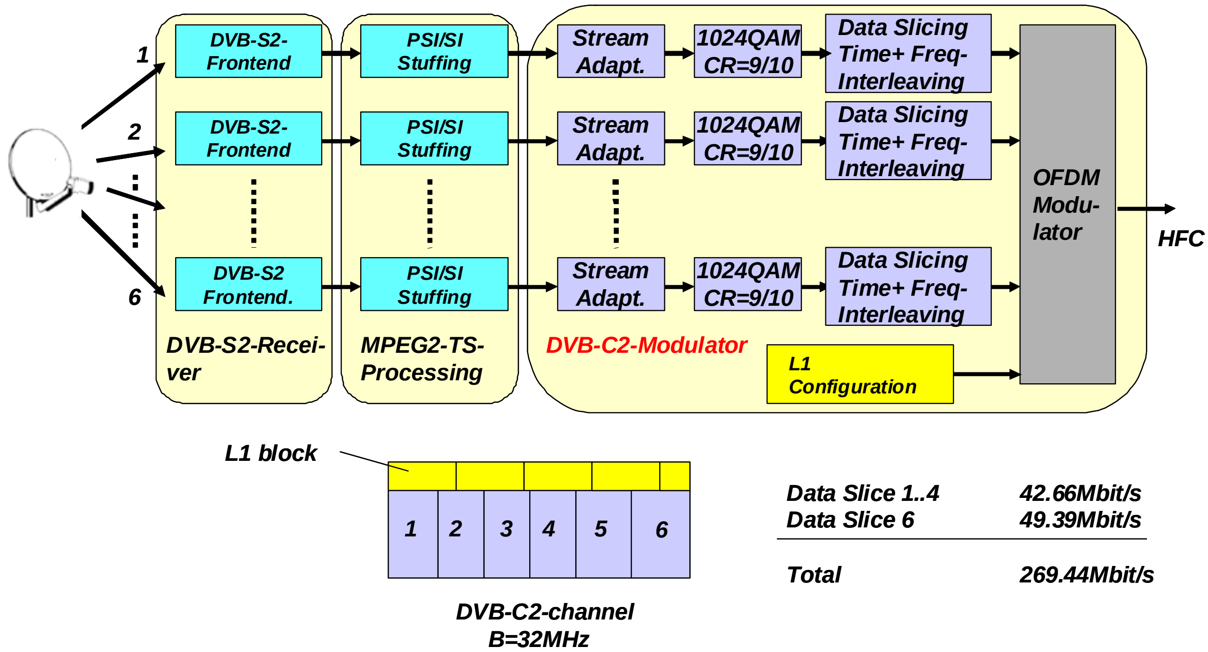

Figure 67: Block diagram of a headend processing unit applying an efficient signal conversion of 6 DVB-S2 signals to a 32 MHz wide DVB-C2 signal

In table 20 details of Data Slice configuration are shown. The 32 MHz C2 channel is inserted between 306 MHz and 338 MHz and centered to 322 MHz to achieve equal frequency distance to lower and higher adjacent channels.

Table 20: Configuration of start frequency and Data Slices of the 32 MHz C2 channel for retransmission of satellite transponders

Parameter	Value	Comment
Center frequency [MHz]	322	
OFDM carrier spacing	1/448 us	Due to 8 MHz channel raster
Dx	24 OFDM carriers	Due to GI=1/128
Center frequency (multiple of Dx) [MHz]	321,96	Aligned to Dx raster
Number of OFDM carriers	13 920 + 1	Within the 32 MHz C2 channel, including the upper edge pilot
Start frequency [MHz]	30,43 MHz	321,96 MHz-(13 920/2)*(1/448 us)
Number of carriers Data slice 1 to 4 Data slice 5 to 6	 2 208 2 544	 Width 1 to 4 = 2 208/448 us = 4,93 MHz Width 5 to 6 = 2 244/448 us = 5,68 MHz
Data Slice tuning positions [MHz] Data Slice 1 to 4 Data Slice 5 to 6	 308,9 313,8 318,7 323,.7 329 334,7	mid frequency of every Data Slice pos1 = start freq + width1/2 pos2 = start freq + width1 + width2/2 etc.

Table 21 shows all L1 part 2 signalling parameters.

Table 21: 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	C2 system identifier
START_FRQUENCY	000000100001100001000000	137 280: value of table 2 multiplied with 448 us (1/carrier spacing)
C2_ BANDWIDTH	0000001001000100	13 920/24 = 580 due to calculated number of carriers. Total 580*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 data symbols in one C2 frame
L1_PART 2_CHANGE_COUNTER	0000 0000	No changes between C2 frames, because one constant data rate multiplex (MPEG2-TS) in each Data Slice. Bit rate of those MPEG2-TS remain constant over time
NUM_DSLICE	0000 0110	6 Data Slices within the C2 channel
NUM_NOTCH	0000	No notch
DATA_SLICE_ID (Data Slice 1 to 6)	0000 0000…….0000 0101	identifier for Data Slice ("0"…."5")
DSLICE_TUNE_POS (Data Slice 1 to 6)	0000000101110 0000010001010 0000011100110 0000101000010 0000110100101 0001000001111	46 (2 208/2/24); tune pos = 1 104th OFDM carrier 138 (3*2 208/2/24); tune pos. = 3 312th carrier 230 (5*2 208/2/24); tune pos = 5 520th carrier 322 (7*2 208/2/24); tune pos = 7 728th carrier 421 (4*2 208+2 544/2)/24; tune pos = 10 104th carrier 527 (4*2 208+3*2 544/2)/24: Tuning position = 12 648th carrier.
DSLICE_OFFSET_LEFT/RIGHT Data slice 1 to 4 Data Slice 5 and 6	 11010010 / 00101110 11001011 / 00110101	 -46/+46 (±2 208/2/24); ± 1 104 carriers left and right from tune pos (Data Slice 1 to 4) -53/+53 (± 2 544/2/24); ±1 272 carriers left and right from the tuning position (Data Slice 5 and 6)

Parameter	Value (transmission format)	Comment
DSLICE_TI_DEPTH(Data Slice 1 to 6)	01	4 OFDM symbols
DSLICE_TYPE (Data Slice 1 to 6)	1	Type 2
DSLICE_CONST_CONF (Data Slice 1 to 6)	1	Data slice does not change
DSLICE_LEFT_NOTCH (Data Slice 1 to 6)	0	No notch
DSLICE_NUM_PLP	000000001	1 PLP per Data Slice
PLP_ID (Data Slice 1 to 6)	0	PLP identifier ("0")
PLP_BUNDLED (Data Slice 1 to 6)	0	Not bundled with PLPs of other Data Slices
PLP_TYPE (Data Slice 1 to 6)	10	Normal data PLP
PLP_PAYLOAD_TYPE (Data Slice 1 to 6)	00011	MPEG2-TS
PLP_GROUP_ID	00	No association with common PLP
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 straight-forward approach however does not exploit the full potentials 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 DVB-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 67. 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 must be foreseen to get a constant bit rate for the MPEG2 Transport Stream at the DVB-C2 modulator input as required by the standard. 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 the preamble symbols 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 preamble symbols is not available for the receiver.

In table 22 details of Data Slice configuration are shown. The 32 MHz C2 channel is inserted between 306 and 338 MHz and centered to 322 MHz to achieve equal frequency distance to lower and higher adjacent channels. One narrowband notches is inserted at the low frequency end of data Slice 6 at 331,85 MHz, with a notch width of 23 subcarriers.

Table 22: Configuration of start frequency and Data Slices of the 32 MHz C2 channel including one narrowband notches

Parameter	Value	Comment
Center frequency [MHz]	322	
OFDM carrier spacing	1/448 us	Due to 8 MHz channel raster
Dx	24 OFDM carriers	Due to GI=1/128
Center frequency (multiple of Dx) [MHz]	321,96	Aligned to Dx raster
Number of OFDM carriers	13 920	Within the 32 MHz C2 channel
Start frequency [MHz]	306,43 MHz	321,96 MHz-(1 3920/2)*(1/448 us)
Number of carriers Data slice 1 to 4 Data slice 5 Data Slice 6	 2 208 2 554 2 543	 Width 1 to 4 = 2 208/448 us = 4,93 MHz Width 5 = 2 554/448 us = 5,70 MHz Width 6 = 2 543/448 us = 5,68 MHz
Data Slice tuning positions [MHz] Data Slice 1 to 4 Data Slice 5 to 6	 308,9 313,8 318,7 323,7 329 334,7	 mid frequency of every Data Slice pos1 = start freq + width1/2 pos2 = start freq + width1 + width2/2 etc.
Narrowband notch	331,85 MHz	Low frequency end of Data Slice 6

Table 23 shows all L1 part 2 signalling parameters of the narrowband notch example.

Table 23: 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	C2 system identifier
START_FRQUENCY	000000100001100001000000	137 280: value of ***Tab 2*** multiplied with 448 us (1/carrier spacing)
C2_ BANDWIDTH	0000001001000100	13 920/24 = 580 due to calculated number of carriers. Total 580*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 data symbols in one C2 frame
L1_PART 2_CHANGE_COUNTER	0000 0000	No changes between C2 frames, because one constant data rate multiplex (MPEG2-TS) in each Data Slice. Bit rate of those MPEG2-TS remain constant over time
NUM_DSLICE	0000 0110	6 Data Slices within the C2 channel
NUM_NOTCH	0001	1 narrowband notch within the C2_system
DATA_SLICE_ID (Data Slice 1 to 6)	0000 0000…….0000 0101	identifier for Data Slice ("0"…."5")
DSLICE_TUNE_POS (Data Slice 1 to 6)	0000000101110 0000010001010 0000011100110 0000101000010 0000110100101 0001000001111	46 (2 208/2/24); tuning position = 1 104th OFDM carrier 138 (3*2 208/2/24); tuning pos.= 3 312th carrier 230 (5*2 208/2/24); tuning pos = 5 520th carrier 322 (7*2 208/2/24); tuning pos = 7 728th carrier 421 (4*2 208+2 544/2)/24; Tuning position = 10 104th carrier 527 (4*2 208+3*2 544/2)/24 + 1; Tuning position = 12 649th carrier.

Parameter	Value (transmission format)	Comment
DSLICE_OFFSET_LEFT/RIGHT Data slice 1 to 4 Data Slice 5 and 6	 11010010 / 00101110 11001011 / 00110101	 -46/+46 (±2 208/2/24); ± 1 104 carriers left and right from tune pos (Data Slice 1-4) -53/+53 (± 2 544/2/24); ±1 272 carriers left and right from tune pos (Data Slice 5 and 6)
DSLICE_TI_DEPTH(Data Slice 1 to 6)	01	4 OFDM symbols
DSLICE_TYPE (Data Slice 1 to 6)	1	Type 2, Data slice does not change
DSLICE_CONST_CONF (Data Slice 1to 6)	1	Data slice does not change
DSLICE_LEFT_NOTCH (Data Slice 1 to 6)	0	No notch
DSLICE_NUM_PLP	000000001	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_GROUP_ID	00	No association with common PLP
PSI/SI_ REPROCESSING	1	Cable NIT, inserted in every MPEG2-TS, therefore PSI/SI reprocessing
Notch_Start	0001000001111	At the low frequency end of Data Slice 6
Notch_Width	1	Lowest possible narrowband notch: 23 subcarriers

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 a broadband notch, where the preamble symbols located in the range of the narrowband notch are switched off. In such a configuration the notch does not weaken the error protection of the preamble as adjacent to the broadband notch in the regular case a complete preamble is located at both sides (lower and higher frequency part) of the broadband notch.

In table 24 details of Data Slice configuration are shown. The 32 MHz C2 channel is inserted between 306 MHz and 338 MHz and centered to 322 MHz to achieve equal frequency distance to lower and higher adjacent channels. One broadband notch is inserted at the high frequency end of data Slice 4 at 323,69 MHz, with a notch width of 1 104 sub-carriers.

Table 24: Configuration of start frequency and Data Slices of the 32 MHz C2 channel including a broadband notch

Parameter	Value	Comment
Center frequency [MHz]	322	
OFDM carrier spacing	1/448 us	Due to 8 MHz channel raster
Dx	24 OFDM carriers	Due to GI=1/128
Center frequency (multiple of Dx) [MHz]	321,96	Aligned to Dx raster
Number of OFDM carriers	13,920	Within the 32 MHz C2 channel
Start frequency [MHz]	306,43 MHz	321,96 MHz-(13 920/2)*(1/448 us)
Number of carriers Data slice 1 to 3 Data slice 4 Data slice 5 Data Slice 6	 2 208 1 104 2 544 2 544	Width 1 to 3 = 2 208/448 us = 4,93 MHz Width 4 = 1 104/448 us = 2,46 MHz Width 5 = 2 544/448 us = 5.m68 MHz Width 6 = 2 545/448 us = 5,68 MHz
Data Slice tuning positions [MHz] Data Slice 1 to 4 Data Slice 5 and 6	 308,9 313,8 318,7 322,45 329 334,7	 mid frequency of every Data Slice pos1 = start freq + width1/2 pos2 = start freq + width1 + width2/2 etc. Data Slice
Broadband notch	323,7 MHz	High frequency end of Data Slice 4 Data Slice

Table 25 shows all L1 part 2 signalling parameters of the broadband notch example.

Table 25: L1 part 2 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	C2 system identifier
START_FRQUENCY	000000100001100001000000	137 280: value of table 2 multiplied with 448 us (1/carrier spacing)
C2_ BANDWIDTH	0000001001000100	137 280: value of table 2 multiplied with 448 us (1/carrier spacing) Total 580*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 data symbols in one C2 frame
L1_PART 2_CHANGE_COUNTER	0000 0000	No changes between C2 frames, because one constant data rate multiplex (MPEG2-TS) in each Data Slice. Bit rate of those MPEG2-TS remain constant over time
NUM_DSLICE	0000 0110	6 Data Slices within the C2 channel
NUM_NOTCH	0001	1 broadband notch within the C2_system
DATA_SLICE_ID (Data Slice 1 to 6)	0000 0000…….0000 0101	identifier for Data Slice ("0"…."5")
DSLICE_TUNE_POS (Data Slice 1 to 6)	0000000101110 0000010001010 0000011100110 0000101000010 0000110100101 0001000001112	46 (2 208/2/24); tune pos = 1 104th OFDM carrier 138 (3*2 208/2/24); tune pos.= 3 312th carrier 230 (5*2 208/2/24); tune pos = 5 520th carrier 322 (7*2 208/2/24); tune pos = 7 728th carrier 421 (4*2 208+2 544/2)/24; Tuning position = 10 104th carrier 527 (4*2 208+3*2 544/2)/24 + 1; Tuning position = 12 649th carrier.
DSLICE_OFFSET_LEFT/RIGHT Data slice 1 to 4 Data Slice 5 and 6	 11010010 / 00101110 11001011 / 00110101	 -46/+46 (±2 208/2/24); ± 1 104 carriers left and right from tune pos (Data Slice 1-4) -53/ + 53 (± 2 544/2/24); ± 1 272 carriers left and right from tune pos (Data Slice 5 and 6)
DSLICE_TI_DEPTH(Data Slice 1 to 6)	01	4 OFDM symbols
DSLICE_TYPE (Data Slice 1 to 6)	1	Type 2, Data slice does not change
DSLICE_CONST_CONF (Data Slice 1 to 6)	1	Data slice does not change
DSLICE_LEFT_NOTCH (Data Slice 1 to 6)	0	No notch
DSLICE_NUM_PLP	000000001	1 PLP per Data Slice
PLP_ID (Data Slice 1 to 6)	0	PLP identifier ("0")
PLP_BUNDLED (Data Slice 1 to 6)	0	Not bundled with PLPs of other Data Slices
PLP_TYPE (Data Slice 1 to 6)	10	Normal data PLP
PLP_PAYLOAD_TYPE (Data Slice 1 to 6)	00011	MPEG2-TS
PLP_GROUP_ID	00	No association with common PLP
PSI/SI_ REPROCESSING	1	Cable NIT, inserted in every MPEG2-TS, therefore PSI/SI reprocessing
Notch_Start	0000101000010	At the high frequency end of Data Slice 4 (starting position 322,45 MHz)
Notch_Width	00101110	1 104 subcarriers ((1 104/24) = 46)

12.1.2 Coding and modulation on service level for "low power mode" in CPEs

Coding and modulation on service level reduce power consumption of CPEs, because demodulation and decoding of only one service instead of the complete transport stream significantly reduce demodulator and decoder speed and consequently the necessary clock frequencies in frontend chips.

If Single Program Transport Streams (SPTS) are transmitted to the headend via an IP backbone, the "low power mode" can be directly supported by the DVB-C2 physical layer without any further equipment in the headend.

For Multi Program Transport Streams (MPTS) as usually received from satellite or terrestrial links, coding and modulation on service level cannot be used. Therefore those MPTS streams must be demultiplexed into Single Program Transport Streams. Remultiplexing then is done by the C2 physical layer. Figure 68 shows the block diagram. SPTS streams are multiplexed in a Data Slice (width < 8 MHz) including a common PLP for PSI/SI information, which has to be generated additionally. Efficient transmission of common PLP is further explained in clause 12.1.3.

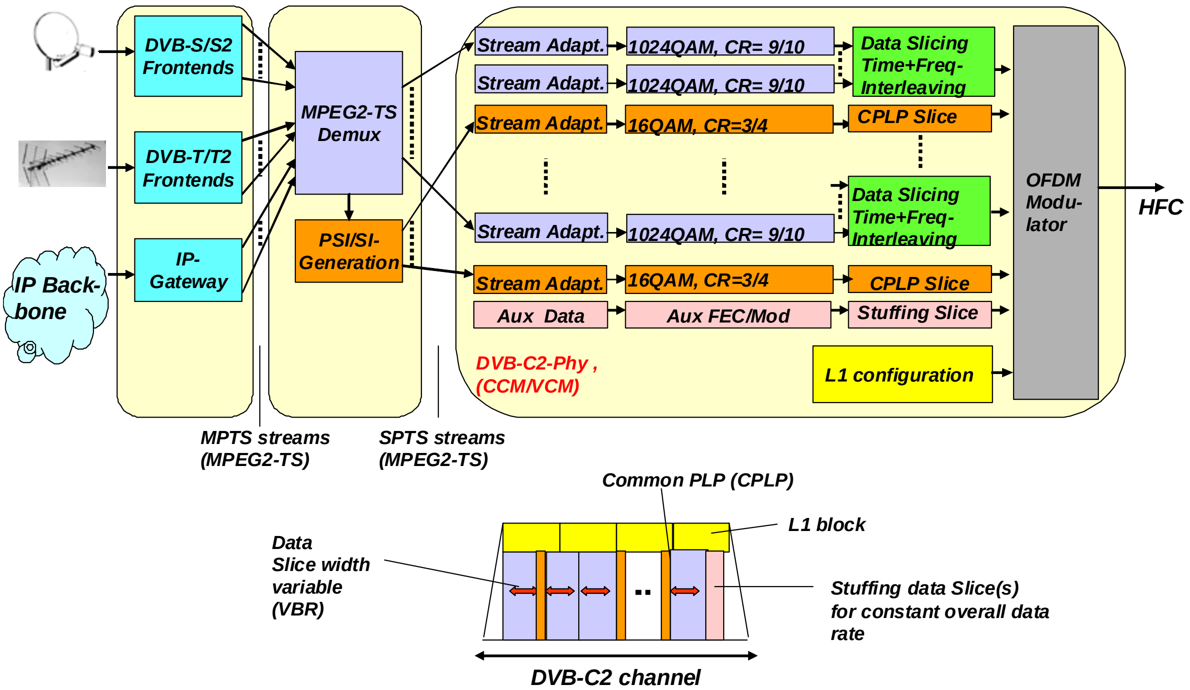

Figure 68: Coding and modulation on Service Level, support of "Low Power Mode" in CPEs

As shown in figure 68 Data Slices have varying bandwidth due to the varying data rate of the single input streams. Therefore on the one hand the maximum slice bandwidth must be lower than the tuner bandwidth (for example 8 MHz), on the other hand the slice bandwidth must be high enough to offer enough frequency interleaving performance, especially if CSO/CTB products from Pal signals occur. Stuffing Data Slices are needed to keep the overall data rate in the C2 channel constant.

12.1.3 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 synchronisation 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.

12.1.4 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.

12.1.5 Utilization for interactive services (IPTV, Internet)

The flowing system explanations are based on the simplified block diagram depicted in figure 69 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 system, whereas the second scenario shows how DVB-C2 is used as an integral part of DOCSIS.

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 cable modem.

Advantages of this scenario:

1. A higher possible downstream capacity for TV programs:
   - by avoiding overhead for IP;
   - by avoiding higher bandwidth buffer overhead by applying enhanced statistical MPEG2 transport stream multiplexers. The DVB-C2 physical layer only offers straight forward statistical multiplexing (PLP- and Data Slice multiplexing). However, "low power mode" for set-top-boxes is not possible in this case.
2. A capacity gain compared to what is explained in scenario 2) IP over HFC. 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).

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 system. Encapsulating the IP packets in Generic Stream Encapsulation (GSE), as supported by DVB-C2, reduces the IP overhead significantly compared to a method based on Multi Protocol Encapsulation (MPE). The capacity gain achieved by GSE goes up to 10 %. MPE is the method currently established for the delivery of IP via MPEG2 Transport Stream. 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.

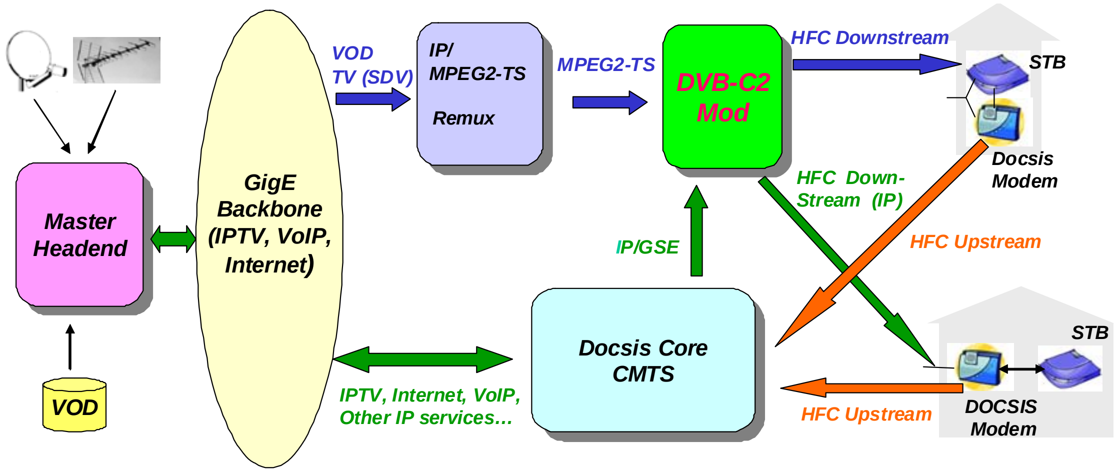

Figure 69: Scenarios for interactive service transmission with DVB-C2 in a stand-alone configuration

NOTE: Figure 69 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)).

12.1.5.1 Example for IP transmission

Below a more detailed IP transmission example is shown.

In a 32 MHz channel the maximum transmission capacity for 1024-QAM and code rate 8/9 is:

$$R = \frac{14\,160}{14\,336} * \frac{448\ us}{10*(128/129)*(8/9)*(14\,232/14\,400)*(95/96)*0.99*(448/449)} = 269{,}3 Mbit/s \qquad (95).$$

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. Figure 60 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.

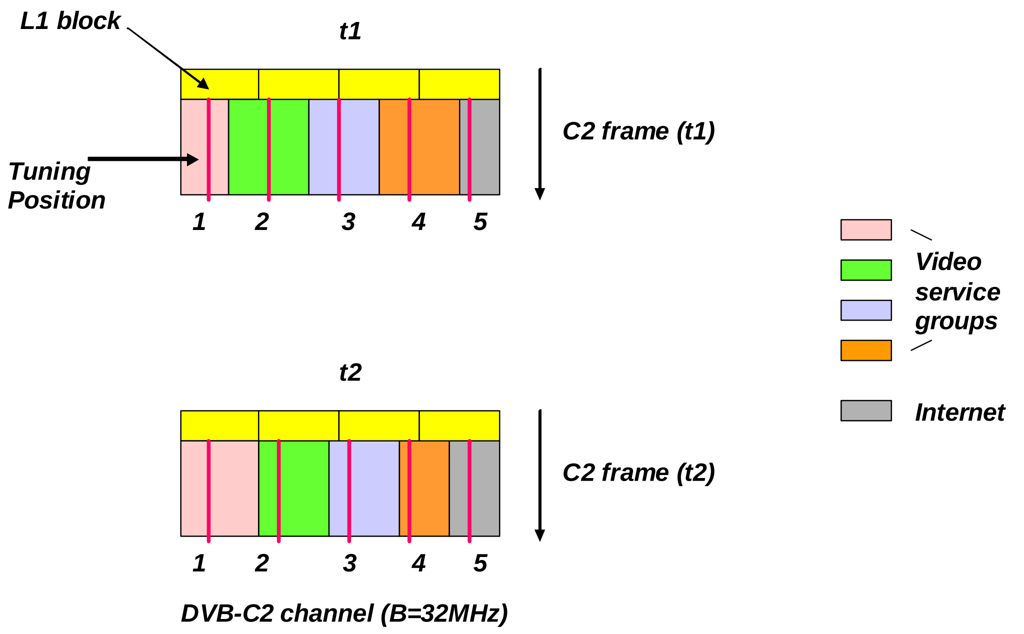

Figure 70: C2 channel configuration for variable bit rate input video and data streams during different C2 frame periods t1 and t2, block diagram of the headend

Figure 70 shows the configuration for variable bit rate input video and data streams and table 26 shows the characteristics (Data Slices, start frequency, GI etc.) of the C2 channel shown in figure 70.

Table 26: 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
Dx	24 OFDM carriers		Due to GI = 1/128
Number of OFDM carriers	14 160		Within the 32 MHz C2 channel
Start frequency [MHz]	306,16 MHz		Multiple of OFDM carrier spacing and Dx raster
Number of carriers, bandwidth	Carr. Number	Bandwidth [MHz]	Data rates [Mbit/s]
Data slice 1 (8 PLPs)	2 424(t1), 3 384(t2)	5,41(t1); 7,55(t2)	Video group 1: 46(t1); 64(t2)
Data slice 2 (8 PLPs)	3 384(t1), 2 904(t2)	7,55(t1); 6,48(t2)	Video group 2: 64(t1); 55(t2)
Data slice 3 (8 PLPs)	2 904(t1), 2 904(t2)	6,48(t1); 6,48(t2)	Video group 3: 55(t1); 55(t2)
Data slice 4 (8 PLPs)	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 10Mbit/s and 2 x SD 2 Mbit/s)
Data slice 5 (2 PLPs)	2 064(t1); 2 592(t2)	4,61(t1); 5,79(t2)	Data group: 40,3(t1); 49,3(t2) One high speed Internet Service
Data Slice tuning positions [MHz] Data Slice 1 to 4	309,32 ; 315,64 ; 321,96 ; 328,29; 334,61		Equidistant distribution within the C2 channel

The calculation of OFDM carrier numbers for the different Data Slices is done according to equation (93), 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 27 shows all L1 part 2 signalling parameters.

Table 27: 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	C2 system identifier
START_FRQUENCY	000000100001100001000000	137 280: value of ***Tab 3*** multiplied with 448 us (1/carrier spacing), which must also be a multiple integer value of Dx (24)
C2_BANDWIDTH	*Still to be added*0000001001001110	14 160/24 = 590 due to calculated number of carriers Total 590*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 data symbols in one C2 frame
L1_PART 2_CHANGE_COUNTER	0000 0001	Changes between C2 frames because of varying data rate of Data Slices
NUM_DSLICE	0000 0101	5 Data Slices within the C2 channel
NUM_NOTCH	0000	No notch within the C2_system
DATA_SLICE_ID (Data Slice 1 to 6)	0000 0000…….0000 0100	identifier for Data Slice ("0"…."5")
DSLICE_TUNE_POS (Data Slice 1 to 5)	0000000111011, 0000010110001, 0000100100111, 0000110011101, 0001000001111	59; 177; 295; 413; 531 see remark in ***Tab4*** Int (3 408/4/24/2) for the tuning pos 1, Tuning pos n = Tuning pos 1 + n*int(3 408/4/24), n = 1, 2, 3, 4
DSLICE_OFFSET_LEFT/RIGHT Data slice 1 Data slice 2 Data slice 3 Data Slice 4 Data Slice 5	 00111011/00101010(t1); 00111011/01010010(t2) 01001100/01000001(t1); 00100100/01010101(t2) 00110101/01000100(t1); 01000100/01011000(t2) 00110010/10110110(t1); 00100001/10001110(t2) 00011011/00111011(t1); 00101111/00111101(t2)	 59/42 (t1); 59/82(t2) 76/65(t1); 36/85(t2) 53/68(t1); 33/88(t2) 50/91(t1); 33/71(t2) 27/59(t1); 47/61(t2) Calculation as shown in ***table 3***
DSLICE_TI_DEPTH(Data Slice 1 to 5)	01	4 OFDM symbols
DSLICE_TYPE (Data Slice 1 to 5)	0	Type 2
DSLICE_CONST_CONF (Data Slice 1to 5)	0	Data slice changes between C2 frames
DSLICE_LEFT_NOTCH (Data Slice 1 to 4)	0	No notch
DSLICE_NUM_PLP	0000 1000 0000 0010	8 PLPs per Data Slice in Data Slice 1 to 4 2 PLPs per Data Slice in Data Slice 5
PLP_ID (Data Slice 1 to 4), 8 PLPs PLP_ID (Data Slice 5), 2 PLPs	0000 00000….0000 0111 0000 00000/0000 0001	PLP identifier ("0/"1"/"2"/"3"/"4"/"5"/"6"/"7") PLP identifier ("0"/"1")
PLP_BUNDLED(Data Slice 1 to 4)	0	Not bundled with PLPs of other Data Slices
PLP_TYPE (Data Slice 1 to 6)	10	Normal data PLP
PLP_PAYLOAD_TYPE (Data Slice 1 to 5)	00010	GSE
PLP_GROUP_ID	00	No association with common PLP
PSI/SI_ REPROCESSING	0	IP data, Video service information is transmitted as SD+S on IP level

12.1.5.2 "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.

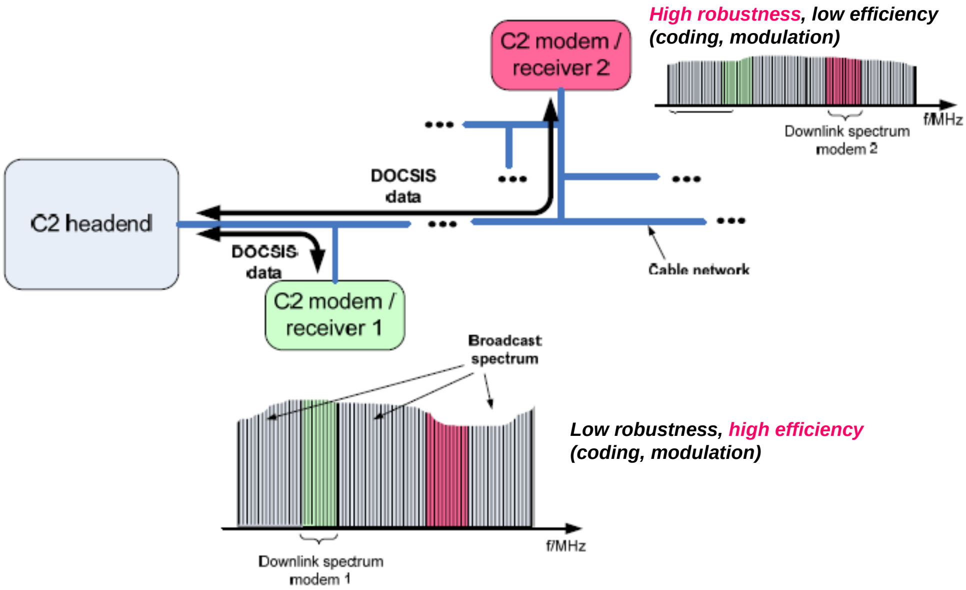

Figure 71: Principle of Adaptive Coding and Modulation (ACM)

Figure 71 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 27. 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 1024-QAM combined with a FEC code rate of 3/4. The remaining outlets support a higher SNR value of 32 dB including 2,5 dB margin, allowing an application of 1024-QAM and FEC code rate of 8/9.

Figure 72 shows the block diagram of the DVB-C2 downstream part for ACM.

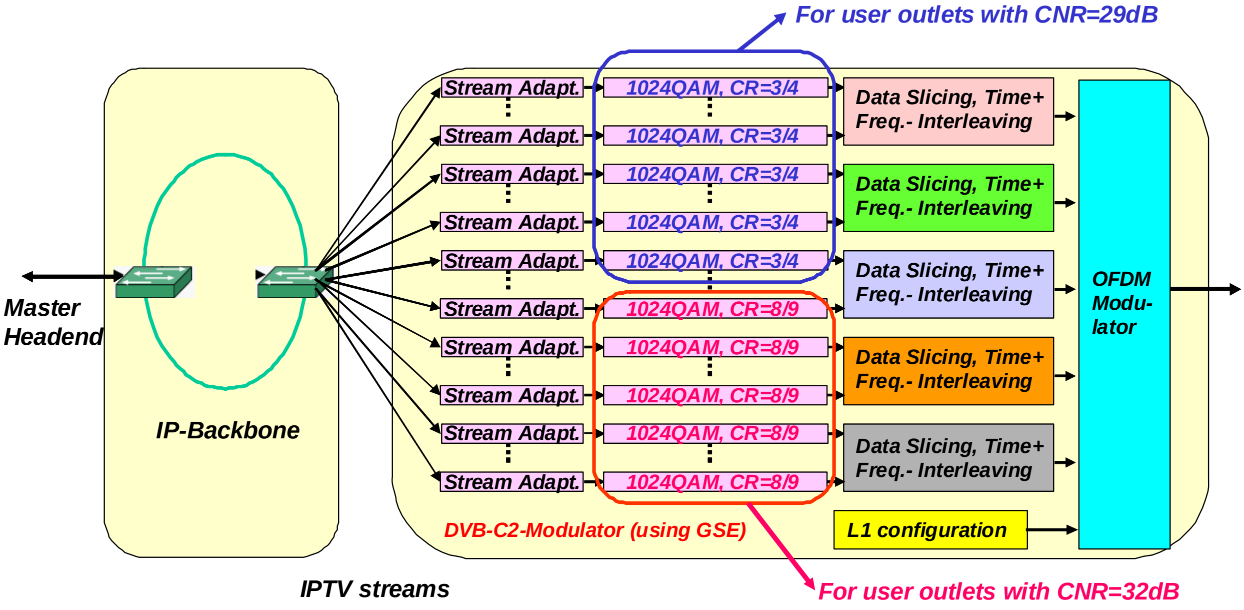

Figure 72: Block Diagram of the downstream configuration for Adaptive Coding and Modulation (ACM) using Variable Coding and Modulation (VCM) in the DVB-C2 Modulator

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) must be applied. In this case generally the robust modulation and coding scheme (1024-QAM, CR = 3/4) must be chosen to assure QoS for the complete network. Therefore the possible data rate is reduced according to the equation below:

$$R = \frac{14160}{14336} * \frac{448us}{10*(128/129)*(3/4)*(11712/11880)*(95/96)*0.99} = 226.65 Mbit/s \quad (96)$$

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:

$$R_{robust} = 0,5 * \frac{14160}{14336} * \frac{448us}{10*(128/129)*(3/4)*(11712/11880)*(95/96)*0.99} = 113.32 Mbit/s \quad (97)$$

for the services with robust coding and modulation and

$$R_{eff} = 0,5 * \frac{14160}{14336} * \frac{448us}{10*(128/129)*(8/9)*(14232/14400)*(95/96)*0.99} = 134.65 Mbit/s \quad (98)$$

for the services with efficient coding and modulation.

The total bit rate therefore is $R_{tot} = 113.32 Mbit/s + 134.65 Mbit/s = 248 Mbit/s$

In this case the capacity increase compared to CCM is about 21 Mbit/s.

Table 28 summarizes the results above.

Table 28: Data rates of the IP transmission example for different operation modes (CCM, VCM/ACM)

	100 % high efficient Mod/Cod (CCM) (1024-QAM, CR = 8/9)	100 % robust Mod/Cod (CCM) (1024-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	226,65Mbit/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 must be changed from "high efficient" (CR = 8/9) to "robust" (CR = 3/4) for this service.

12.1.5.3 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 73 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 part 2 signalling bundled PLP must be indicated (BUNDLED_PLP="1") in table 27. The remaining structure of table 27 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.

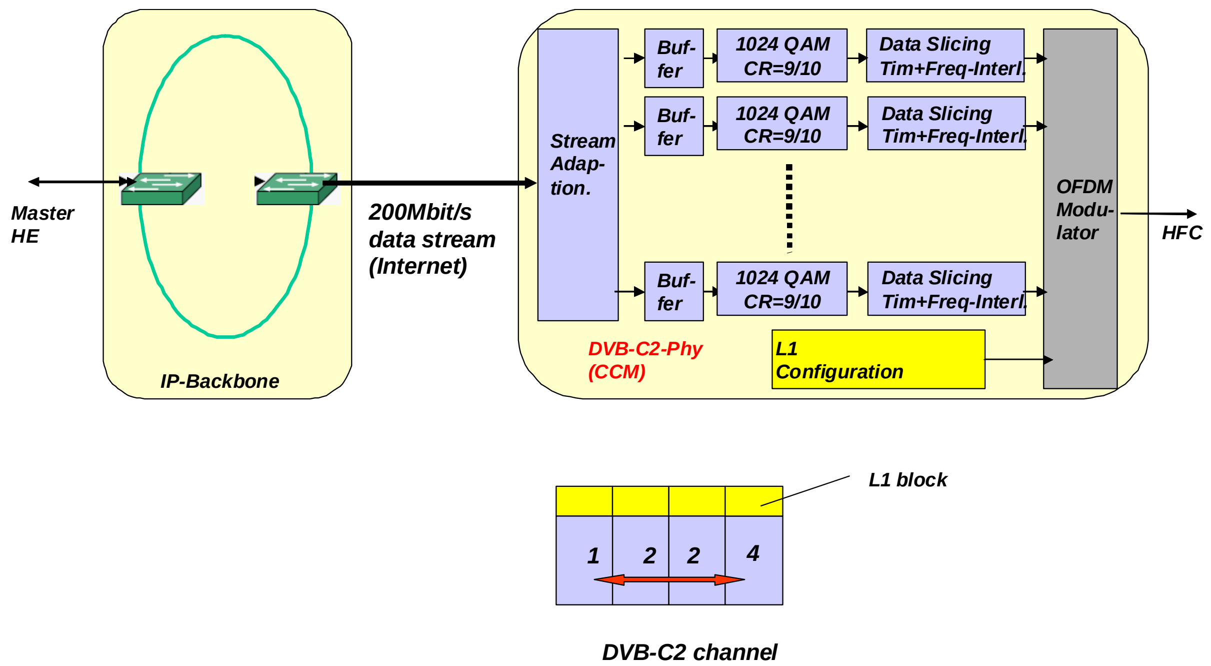

Figure 73: Application example for PLP bundling; transmission of a high data rate MPEG2 transport stream via several bundled Data Slices in a DVB-C2 channel

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.

12.1.6 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 × 2,232 kHz = 104,9 kHz within 8 MHz channel bandwidth. Those narrow notches are called 'Narrowband Notches' in the DVB-C2 system.

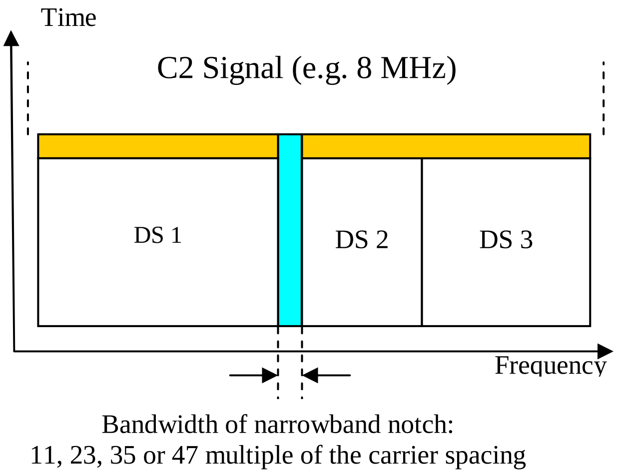

Figure 74: Narrowband Notch with a maximum bandwidth of 104,9 kHz

Figure 74 shows schematically a cable TV channel (8 MHz). The blue bar is a notched frequency band with a maximum bandwidth of 104,9 kHz.

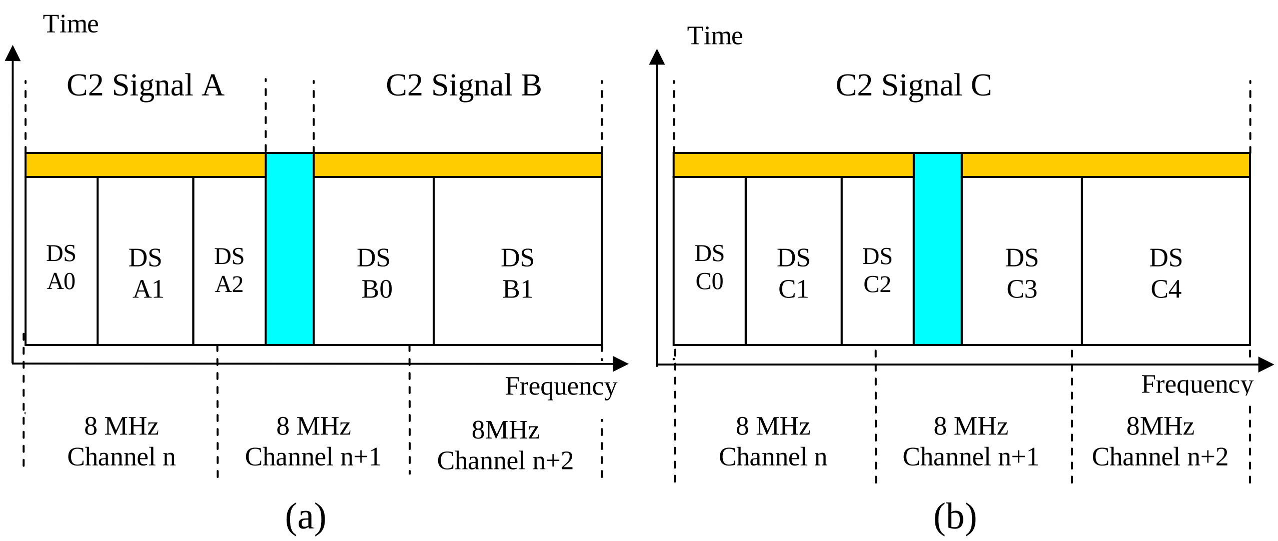

Figure 75: Two options for the implementation of Broadband Notches in a DVB-C2 signal

NOTE: Figure 75shows two options for the implementation of Broadband Notches in a DVB-C2 signal: (a) Broadband Notch placed between 2 adjacent DVB-C2 signal spectra and (b) Broadband Notch placed within a single DVB-C2 signal spectrum.

Figure 75 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 75 (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 75 (b).

NOTE: Only in case of PLP static mode, 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.1].

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.

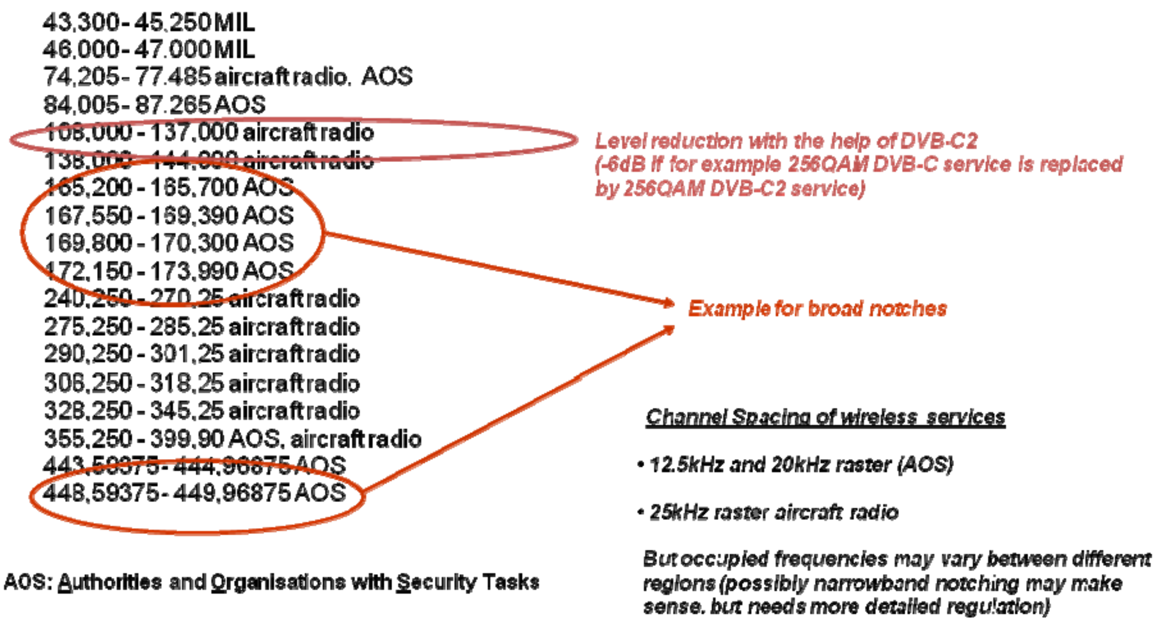

Figure 76: Example of the used opportunities from DVB-C2 to reduce the interference to wireless services

Figure 76 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), there is a 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

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 µs (6 MHz) and 7/64 µs (8 MHz), respectively (see according to [i.1], table 37). 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.1] table 38 which is also included in the present document as table 29.

Table 29: OFDM parameters for 6 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 L1 Block K_{L1}	3 408	3 408	3 408	3 408
Bandwidth of L1 Signalling Block (see note)	*5,71 MHz*	*5,71 MHz*	*7,61 MHz*	*7,61 MHz*
Duration T_U	4 096T	4 096T	4 096T	4 096T
Duration T_U µs (see note)	*597.3*	*597.3*	*448*	*448*
Carrier spacing $1/T_U$ (Hz) (see note)	*1 674*	*1 674*	*2 232*	*2 232*
Guard Interval Duration Δ/T_u	*64T*	*32T*	*64T*	*32T*
Guard Interval Duration Δ/T_u µs (see note)	*9,33*	*4,.66*	*7*	*3,5*
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 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 29) of the two system configurations will probably be taken into account as well.

Figure 77 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.

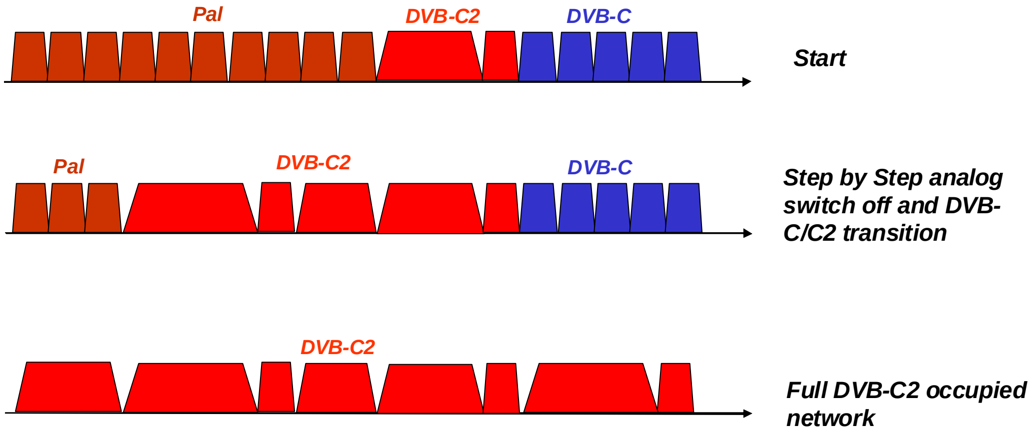

Figure 77: Spectrum occupation during the transition from analogue TV and DVB-C to DVB-C2

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 must 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 78 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 1024-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 4096-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 65K in combination with the remaining DVB-C2 signal processing. However, these constellation schemes are not defined yet by the current version of the standard but may be subject to investigation for future updates.

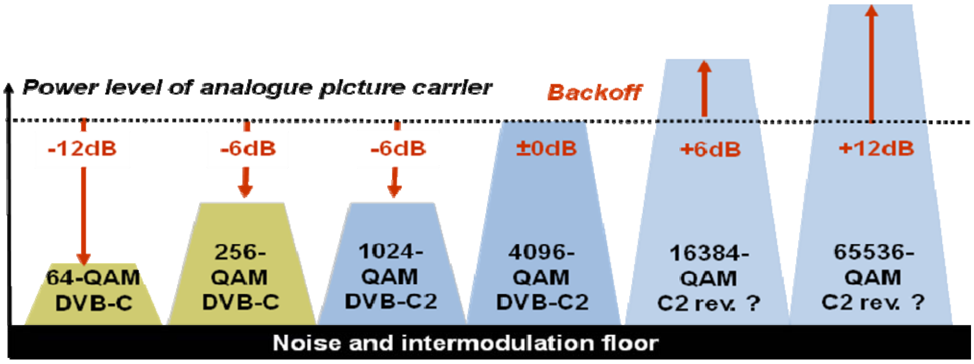

Figure 78: Examples for power levels applicable to DVB-C and DVB-C2 signals in cable networks

12.2.3 Non-backward compatibility to DVB-C

The signal processing defined by the DVB-C2 standard 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.

Annex A (informative): Examples for the calculation of payload capacity of the DVD-C2

Annex A gives two examples for calculated payload capacities of DVB-C2_systems with different parameters.

A.1 Examples for payload capacity using Guard Intervall 1/128

The following set of parameters applies for table A.1: 1 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
1024-QAM 3/4	55,20	56,61	57,09	57,32	57,68
1024-QAM 5/6	61,39	62,97	63,49	63,75	64,15
1024-QAM 9/10	66,28	67,98	68,55	68,83	69,26
4096-QAM 5/6	73,67	75,56	76,19	76,51	76,98
4096-QAM 9/10	79,54	81,58	82,26	82,60	83.11

A.2 Examples for payload capacity using Guard Intervall 1/64

The following set of parameters applies for table A.2: 1 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
1024-QAM 3/4	54,22	55,62	56,08	56,31	56,66
1024-QAM 5/6	60,31	61,86	62,37	62,63	63,02
1024-QAM 9/10	65,11	66,78	67,34	67,62	68,04
4096-QAM 5/6	72,37	74,23	74,85	75,16	75,62
4096-QAM 9/10	78,13	80,14	80,81	81,14	81,64

Annex B (informative): Bibliography

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