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Digital Video Broadcasting (DVB); DVB-SH Implementation Guidelines

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Reference

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Contents

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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).

The present document is complementary to EN 302 583 [1], which provides the waveform-level specifications for the use of TV Broadcast in DVB Satellite to Handhelds applications.

The present document is complementary to TS 102 585 [2], which provides the system-level specifications for the use of TV Broadcast in DVB Satellite to Handhelds applications.

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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Founded in September 1993, the DVB Project is a market-led consortium of public and private sector organizations in the television industry. Its aim is to establish the framework for the introduction of MPEG-2 based digital television services. Now comprising over 200 organizations from more than 25 countries around the world, DVB fosters market-led systems, which meet the real needs, and economic circumstances, of the consumer electronics and the broadcast industry.

Introduction

The present document gives the first guidelines for setting up networks and services using the Digital Video Broadcasting - Satellite to Handheld (DVB-SH) specifications. Updates to the present document will be produced when more results become available.

Document summary

Clause 4 gives an outline of the DVB-SH system, the main hybrid satellite/terrestrial architectures, the distinction between SH-A and SH-B as well as between SFN and non-SFN, the possible approaches for frequency planning, the frequency bands and their regulatory aspects, the characteristics of satellite and terrestrial propagation channels, the receivers constraints, and other system considerations.

Clause 5 gives a discussion of possible DVB-SH system configurations and likely scenarios of their deployment sequence.

Clause 6 introduces the elements of the DVB-SH specifications at the link layer and above. These are, at the link layer, time-slicing and LL-FEC and, at above this layer, the supports of VBR/Statmux and of Mobility. Compatibility with DVB-H link layer is discussed.

Clause 7 clarifies the physical layer elements of the DVB-SH specifications that are new to the DVB families of standards, namely the 3GPP2 turbo codes, the long physical time interleaver and its implication on receivers (Class 1 and Class 2 receivers), the 1K-FFT mode for OFDM and the diversity combining techniques (MFN).

Clause 8 discusses services and usage scenarios. This clause presents concepts such as service categories and attributes and highlights the consequences of the hybrid architectures on a personal mobile TV service offering.

Clause 9 is devoted to DVB-SH network configuration. The following issues are discussed: Network configurations, Service Information, Handover.

Clause 10 provides preliminary information about the DVB-SH reference terminals and reference receivers.

Clause 11 is devoted to Network coverage planning. The main purpose is to present reference data and methodologies for transmitter network (both satellite and terrestrial) sizing as a function of coverage.

Finally four annexes are included:

- Annex A documents the methodology used by TM-SSP and presents the simulation results that support the recommendations given.
- Annex B details the implications for "convergence terminals" in which broadcast (DVB-SH, DVB-H) and telecommunication (GSM/UMTS) technologies co-exist.
- Annex C documents the methodology and formulas for spectrum efficiency analysis and throughput calculations.
- Annex D provides with complementary bibliography.

1 Scope

The present document provides guidelines for the use and implementation of ETSI Digital Video Broadcasting–Satellite to Handheld (DVB-SH) standard EN 302 583 [1] in the context of providing an efficient way of carrying multimedia services over digital hybrid satellite/terrestrial broadcasting networks to handheld terminals.

The document should be read in conjunction with the:

- DVB-SH waveform specifications [1];
- DVB-SH system specifications [2];
- DVB-CBMS (to be completed);
- DVB-GBS specifications (to be completed).

Objective

The present document draws attention to the technical questions that need to be answered when setting up DVB-SH services and networks and offers some guidance in finding answers to them. It does not cover in detail, issues linked to the content of the broadcasts such as Coding Formats, Electronic Programme Guides (EPG), Access Control (CA), etc.

Target readers

The present document is aimed at the Technical Departments of organizations that are considering implementing digital hybrid satellite/terrestrial broadcasting to handheld devices. It assumes that readers are familiar with digital satellite and terrestrial networks.

Contributors

The present document was prepared by members of the Ad-hoc group TM-SSP from the DVB Project. Members include broadcasters, network operators and professional as well as domestic equipment manufacturers.

2 References

References are either specific (identified by date of publication and/or edition number or version number) or non-specific.

- For a specific reference, subsequent revisions do not apply.
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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 indispensable for the application of the present document. For dated references, only the edition cited applies. For non-specific references, the latest edition of the referenced document (including any amendments) applies.

[1] ETSI EN 302 583 (V1.1.1): "Digital Video Broadcasting (DVB); Framing Structure, channel coding and modulation for Satellite Services to Handheld devices (SH) below 3 GHz".

[2] ETSI TS 102 585 (V1.1.1): "Digital Video Broadcasting (DVB); System Specifications for Satellite services to Handheld devices (SH) below 3 GHz".

[3] ETSI EN 302 304 (V1.1.1): "Digital Video Broadcasting (DVB); Transmission System for Handheld Terminals (DVB-H)".

[4] Void.

[5] Void.

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[7] Void.

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[22] ETSI TS 102 472: "Digital Video Broadcasting (DVB); IP Datacast over DVB-H: Content Delivery Protocols".

[23] ISO/IEC 13818-6: "Information technology -- Generic coding of moving pictures and associated audio information; Part 6: Extensions for DSM-CC".

[24] ETSI EN 300 468: "Digital Video Broadcasting (DVB); Specification for Service Information (SI) in DVB systems".

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[26] IEEE 802.11 (2007): "Standard for Information technology-Telecommunications and information exchange between systems-Local and metropolitan area networks-Specific requirements; Part 11: Wireless LAN Medium Access Control (MAC) and Physical Layer (PHY) Specifications".

[27] ITU-R Recommendation P.526 (2007): " Propagation by diffraction".

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[33] ITU-R Recommendation P.680 (1999): "Propagation data required for the design of Earth-space maritime mobile telecommunication systems".

2.2 Informative references

The following referenced documents are not essential to the use of the present document but they assist the user with regard to a particular subject area. For non-specific references, the latest version of the referenced document (including any amendments) applies.

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[i.23] ETSI TR 102 376 (V1.1.1): "Digital Video Broadcasting (DVB) User guidelines for the second generation system for Broadcasting, Interactive Services, News Gathering and other broadband satellite applications (DVB-S2)".

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NOTE: Available at http://www.ero.dk/documentation/docs/doc98/official/pdf/ECCDEC0505.PDF.

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3 Definitions and abbreviations

3.1 Definitions

For the purposes of the present document, the following terms and definitions apply:

cell: contiguous coverage area obtained by aggregating the coverages of adjacent terrestrial transmitters using the same frequency configuration to broadcast exactly the same content in the area

NOTE: A single Cell_ID is announced in the area.

common content (or Service): content available via the SC and also obligatorily transmitted in each cell of the CGC

Complementary Ground Component (CGC): terrestrial leg of a hybrid network that uses a portion of the frequency allocated to that network, under a special regulatory regime decided at regional or national levels, to provide a complementary cellular coverage of the regional/national territory

NOTE: It can not exist independently of the SC it is associated to. Also called ATC in the US.

hybrid frequency: frequency used by a terrestrial transmitter carrying the Common content

NOTE: A Non-Hybrid frequency carries only Local content.

Inter-Burst FEC (IFEC): is a new FEC at link layer standardized in DVB-SH to combat erasures of several complete Bursts

NOTE: It usually operates over protection periods of 10 s or more. It uses the syntax of the conventional MPE-FEC (see EN 301 192 [9]), but extends this technique with spreading and interleaving, to achieve the stated improved protection.

late/early decoding: "Late decoding" refers to the technique of buffering the incoming channel data so that delivery to the source decoders occurs only when all data has been received that ensure maximum FEC protection

NOTE: On the contrary, "Early decoding" refers to the technique of delivering data to the source decoders before all data has been received, for the purpose of improving zapping time. This introduces jitter, in adverse reception conditions, that has to be taken into account by the presentation layer.

Link Layer FEC (LL-FEC): FEC implemented at the Link layer. It usually includes a time interleaver and relies on independent framing and error detection mechanisms to correct long strings of erased bits

NOTE: In DVB-SH, link-layer FEC are constructed using the MPE syntax.

local content: content not available via the SC and the not guaranteed to be available in every cells of the CGC

modulation: process of varying a sine wave to convey a message

NOTE: Four digital modulations are used in DVB-SH: QPSK, 8PSK(TDM only), 16APSK(TDM only) and 16QAM(OFDM only).

nomultiplex: group of services transmitted with the same set of physical parameters (frequency, bandwidth, waveform, modulation, error protection, guard-time, etc.)

OFDM (waveform): waveform composed of a large number of closely packed sine waves each modulated at low speed

NOTE: This waveform is optimized for terrestrial propagation, but can also be used with satellite. This waveform allows SFN operation by several transmitters.

partially available Transport Stream: ransport Stream that carries at least one DVB service that is not transmitted in all cells where the TS is transmitted

NOTE: This is signalled by the service_availability_descriptor in the SDT.

receiver: functional module of a terminal from antenna(s) to IP baseband output. TS 102 585 [2] currently specifies two *classes* of receivers, according to their capability and methods to process time diversity elements in the waveform (see clause 10)

Satellite Component (SC): satellite leg of a hybrid network, using a portion of the frequency allocated to satellite services at international level, and which appears as an umbrella cell to the DVB-SH receivers

sub-band: contiguous segment of the allocated spectrum usually dedicated to the transmission of a multiplex

NOTE: In DVB-SH, sub-bands can have 1,7 MHz; 5 MHz; 6 MHz; 7 MHz and 8 MHz of bandwidth.

TDM (waveform): waveform composed of a sine wave modulated at high speed and with a modulation designed to produce a low temporal envelop variation

NOTE: This waveform optimized for use with satellite. It does not allow SFN operations.

terminal: device comprising several functional modules: receiver, video/audio processing, battery/power, display and optionally (non-DVB-SH) bidirectional radio functions

NOTE: For network planning, several terminal *categories* are distinguished in DVB-SH (see clause 10).

time interleaver: physical layer operation introduced for mitigating long fades

NOTE: At the transmitter it breaks one turbo code word into several pieces (also called Interleaver Units) and spread these over time (from several hundred ms to more than 10 s). In the receiver, these interleaver units are collected, reordered and forwarded to the turbo decoder.

Transport Stream (TS): multiplex based on the MPEG2 packet protocol and complete with its EN 300 468 [24] compliant SI signalling

NOTE: A TS has a unique identification through the Transport_Stream_ID and the Original_ network_ID. By extension, two multiplexes having the same unique identification but transmitted in parallel in the SC and the CGC are considered in DVB-SH as forming a unique TS.

waveform: shape and form of a transmitted signal. DVB-SH has defined two waveforms: TDM and OFDM

3.2 Abbreviations

For the purposes of the present document, the following abbreviations apply:

16APSK	16-state APSK
16QAM	please change to 16-QAM
3G	Third Generation
8PSK	please change to 8-PSK
A3GH	Above 3 GHz (frequency band)
ACLR	Adjacent Channel Leakage Ratio
ACS	Adjacent Channel Selectivity
ADT	Application Data Table
ASI	Asynchronous Serial Interface
ATC	Ancillary Terrestrial Component (US)
AWGN	Additive White Gaussian Noise
B3GH	Below 3 GHz (frequency band)
BFN	Beam Forming Network
BW	Bandwidth
C/I	Carrier-to-Interference ratio
C/N	Carrier-to-Noise ratio
CBMS	Convergence of Broadcast and Mobile Systems (DVB specifications)
CDF	Cumulative Distribution Function
CEPT	Conference Européenne des Postes et Telecommunications
CGC	Complementary Ground Component
CRC16	16-bit CRC
CU	Capacity Unit (DVB-SH)
dBic	(Gain) in dB relative to Isotropic with Circular polarization
DFL	DataField Length
DTT	Digital Terrestrial Transmission
DU	Dense Urban
DVB-GBS	DVB Generic data Broadcasting & Service information protocols
DVB-SH-IG	DVB-SH Implementation Guidelines
DVB-T/H	This is a short form for « either DVB-T or DVB-H”
ECC	European Community Commission
EIRP	Equivalent Isotropic Radiated Power
ESG	Electronic Services Guide
ESR(5)	Performance index computed based on ESR5
ESR5	Errored-Second-Ratio at % in 20 s (ITU)
FCC	Federal Communications Commission (US)
FDT	FEC Data Table
FEC	Forward Error Correction
FER	Frame Error Rate
G/T	Receiver fiure of merit
GBBF	Ground Based Beam Forming
GBS	should be DVB-GBS
GEO	Geosynchronous or Geostationary (orbit)
GI	Guard Interval (for OFDM)
GMR1 and 2	Geo-Mobile Radio 1 and 2 (ETSI standards)
GSE	Generic Stream Encapsulation (IETF)
HEO	Highly Elliptical Orbit
Hyb	Hybrid
iFDT	Interburst FEC Data Table

IFEC	Should be always MPE-IFEC
INT	IP Notification Table (DVB-GBS)
IPDC	IP DataCasting (DVB specifications)
ITS	should be LMS-ITS
IU	(Time) Interleaver Unit (DVB-SH)
Ka-band	Fequency band (defined by ITU) in the range of 18-30 GHz
Ku-band	Frequency band (defined by ITU) in the range of 10-14 GHz
LEO	Low-Earth Orbit
LHCP	Left hand circular polarisation
LL-FEC	Link-Layer FEC
LL-FEC	Link-Layer Forward Error Correction
LLR	Log-Likelihood Ratio
LMS	Land Mobile Satellite (channel)
LMS-ITS	LMS- Intermediate Tree Shadowing
LMS-SU	LMS- SubUrban
LNA	Low Noise Amplifier
LNB	Low noise block converter
LOS	Line Of Sight
MAC	Medium Access Control (layer)
MBRAI	Mobile Radio Access Interface (specifications)
MEO	Medium Earth Orbit
MER	Modulation Error Ratio
MFER	see DVB-H implementation guidelines
MFN	Multi Frequency Network
MPE-FEC	see DVB-H implementation guidelines
MPEG TS	please change to MPEG-2 TS
MPEG2 TS	please change to MPEG-2 TS
MPEG2	please change to MPEG-2
MPEG-TS	please change to MPEG-2 TS
MPE-IFEC	MPE Inter-burst Forward Error Correction (DVB-SH)
MRP	Milestone review procedure
MSB	most significant bit
Msec	millisecond
MSS	Mobile Satellite System
NF	Noise Figure
NIT	Network Identification Table (DVB)
nm	Nano meter
NPR	Noise Power Ratio
OBO	Output Back-Off
OFDM	Orthogonal Frequency Division Multiple Access
OFDM/TDM	OFDM or TDM
PAT/PMT	Program Association Table/ Program Map Table
PDA	Personal Digital Assistant
PDF	Probability Density Function
PDP	Power-Delay Profile
PFD	Power Flux Density
PHY	Physical Layer
PID	Program Identifier (MPEG)
PL SLOTS	Physical Layer Slots
PL	Physical Layer
Pps	(one) pulse-per-second
PSI/SI	Program Specific Information / Service Information
PVR	Personal Video Recording
QoS	Quality of Service
QPSK	Quadrature Phase Shift Keying
RF	Radio Frequency
RHCP	Right Hand Circular Polarisation
RNPT	Radio Network Planning Tool
RS	Reed-Solomon (FEC)
Rx	Receive
Sat	Satellite
SC	Satellite Component (DVB-SH)

SDARS	Satellite Digital Audio Radio System (US)
SDT	Service Description Table (DVB)
SFN	Single Frequency Network
SGC	should be CGC
SH	Satellite to Handheld (specification)
SH-A	SH- waveform configuration A (OFDM/OFDM)
SHA	Should be SH-A
SH-B	SH- waveform configuration B (TDM/OFDM)
SHB	Should be SH-B
SH-FRAME	A baseband frame defined in EN 302 584
SHIP	SH frame Information Packet (DVB-SH)
SHIP	Should be SHIP
SHWav	To be changed to EN 302 584
SI	Service Information (DVB)
SLA	Service Level Agreement
SMS	Short Messages Service
SNORE	Signal to Noise Ratio estimator
SNR	Signal-to-Noise Ratio
SRRC	Square-Root Raised-Cosine
SU	Sub-Urban
S-UMTS	Satellite UMTS
TDL	Tapped Delay Line
TDM	Time-Division Multiplex
TDMA	Time Division Multiple Access
TPS	Transmission Parameter Signalling
TR(x)	Terrestrial Repeater, type x (DVB-SH)
TS PER	Transport Stream Packet Error Rate
TS	Transport Stream
TU6	Typical Urban channel with 6 taps
TWTA	Travelling-Wave Tube Amplifier
U	Urban
VBR	Variable Bit Rate (video)
VoD	Video on Demand
VSWR	Voltage Standing Wave Ratio
w.r.t.	With respect to
WER	word error rate

4 System outline

4.1 DVB-SH system capability outline

DVB-SH system provides an efficient and flexible mean of carrying broadcast services over an hybrid satellite and terrestrial infrastructure operating at frequencies below 3 GHz to a variety of portable, mobile and fixed terminals having compact antennas with very limited or no directivity. Target terminals include handheld defined as light-weight and battery-powered apparatus (e.g. PDAs, mobile phones), vehicle-mounted, nomadic (e.g. laptops, palmtops, etc.) and stationary terminals.

The broadcast services encompass streaming services such as television, radio programs as well as download services enabling for example Personal Video Recorder services.

The DVB-SH system coverage is obtained by combining a Satellite Component (SC) and, where necessary, a Complementary Ground Component (CGC) to ensure service continuity in areas where the satellite alone can not provide the required QoS. The SC ensures wide area coverage while the CGC provides cellular-type coverage. All types of environment (outdoor, indoor, urban, suburban and rural) can then be served. It should be noted that the area served by a beam of currently planned multibeam satellites is in the order of 600 000 Km^2.

DVB-SH can provide the following theoretical total capacity per 5 MHz (typical) satellite bandwidth and per beam.

Table 4.1: DVB-SH typical and maximum net bit rates (Mbps) in Satellite-only coverage (SAT) and in Terrestrial coverage (TER)

Waveform		**SH-A**				**SH-B**	
Hybrid network frequency configuration		SFN		MFN		MFN	
		Typ	Max	Typ	Max	Typ	Max
Multibeam satellite system with 3x5 MHz spectrum assigned, 5 MHz allocated to each satellite beam in a 3-color reuse pattern	SAT/beam TER/beam	2,5 10,0	10,0 30,0	2,5 7,5	10,0 20,0	2,66 7,42	10,64 20,53
Multibeam satellite system with 4x5 MHz spectrum assigned, 5 MHz allocated to each satellite beam in a 4-color reuse pattern	SAT/beam TER/beam	2,5 13,74	10,0 40,0	2,5 11,24	10,0 30,0	2,66 11,13	10,64 30,43
NOTE 1: Values are indicative as different optimizations allowed by the standard lead to slightly different results. NOTE 2: The TER capacity includes the repetition of the SAT capacity.							

The total system capacity for a multibeam system is obtained by multiplying each number above by the number of beams. For example a hypothetical 9-beam satellite with 20 MHz bandwidth can offer, in terrestrial coverage typically 100 Mbps total (MFN/SH-B/Typ) and up to 360 Mbps total (SFN/SH-A/Max) when there is no power constraint. The same satellite can offer, in satellite-only coverage typically 24 Mbps total (MFN/SH-B/Typ) and up to 96 Mbps total (MFN/SH-B/Max) when there is no power constraint.

Currently planned multibeam satellite systems target a minimum of 6,6 Mbps total per 5 MHz of satellite bandwidth per beam (of which 2,2 Mbps capacity when in satellite-only reception conditions). Higher total capacity can be achieved through more satellite frequency-reuse, more powerful satellites, more CGC transmitters density or more advanced terminal technologies (e.g. antenna diversity).

4.1.1 System overview

A typical DVB-SH system (see figure 4.1) is based on a hybrid architecture combining a SC and, where necessary, a CGC consisting of terrestrial repeaters fed by a broadcast distribution network of various kinds (satellite [DVB-S/S2], terrestrial [fibre, xDSL, etc.]). Three kinds of terrestrial repeaters are envisaged:

- TR(a) are broadcast infrastructure transmitters which complement reception in areas where satellite reception is difficult, especially in urban areas; they may be collocated with mobile cell sites or standalone. Local content insertion at that level is possible, relying on adequate radio frequency planning and/or waveform optimizations.
- TR(b) are personal gap-fillers of limited coverage providing local re-transmission, on-frequency and/or with frequency conversion; typical application is indoor coverage provision, locally repeating the satellite signal available outdoor. No local content insertion is foreseen.
- TR(c) are mobile broadcast infrastructure transmitters creating a "moving complementary infrastructure" on board moving platforms (cars, trains, bus). Depending on waveform configuration and radio frequency planning, local content insertion may be possible.

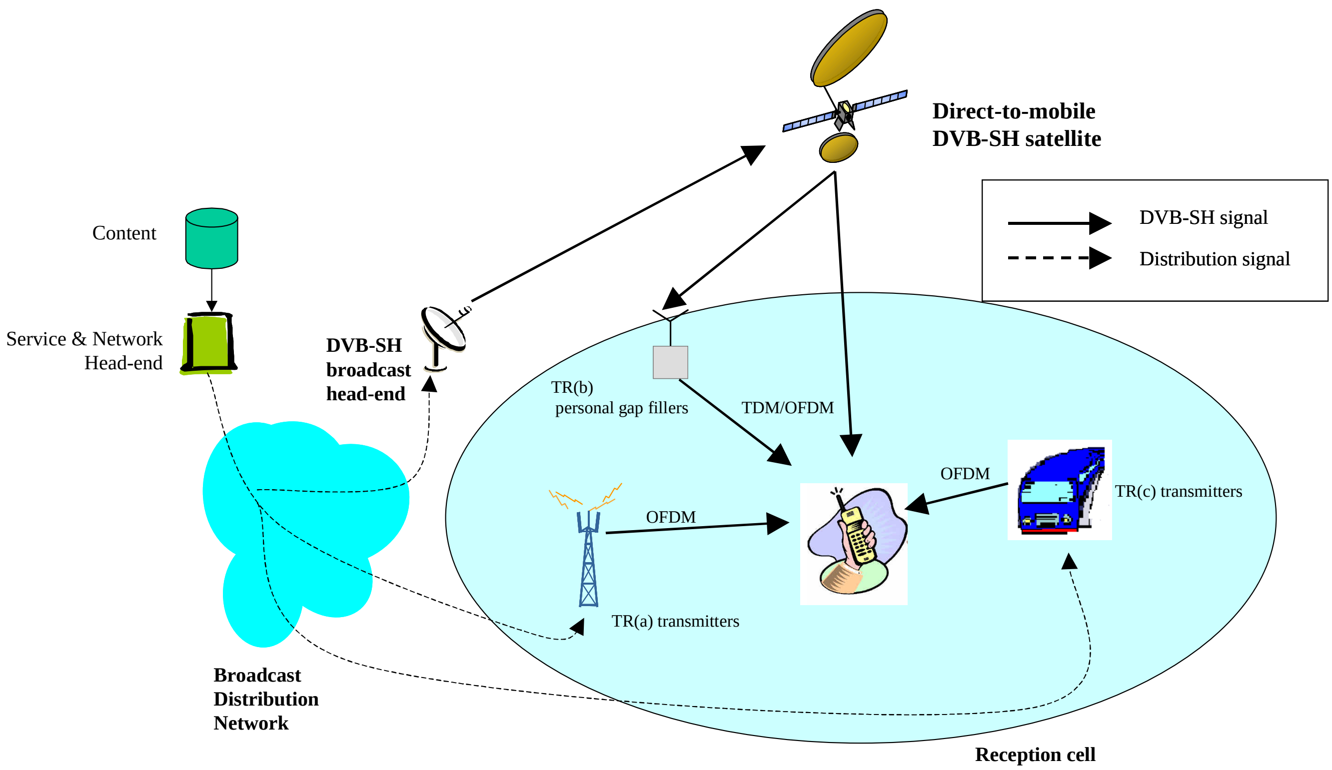


Figure 4.1: Overall DVB-SH transmission system reference architecture

The DVB-SH transmission system key features are the following:

- seamless service continuity between SC and CGC coverages;
- support of all reception conditions associated to portable and mobile terminals: indoor/outdoor, urban/suburban/rural, static/mobile conditions. Typical mobility conditions covers pedestrian as well as land vehicular scenarios;
- possible implementation of power saving schemes to minimize the power consumption of battery activated terminals in order to maximize autonomy;
- local insertion of broadcast services on CGC;
- use different kinds of distribution network to feed the CGC repeaters, such as Satellite (DVB-S/S2) and/or terrestrial (Optical fibre, Wireless Local Loop, xDSL, etc.) resources;
- possible offering of interactive services by inter-working with mobile or wireless systems at service, network head-ends and user terminal levels.

Two main different physical layer configurations are supported by the DVB-SH waveform standard [1]:

- SH-A exploiting OFDM transmission mode for both SC and CGC. SH-A allows (but does not impose) a Single Frequency Network (SFN) between the SC signal and the CGC signal carrying the same content. If SFN is implemented then FEC, Channel Interleaver, modulation and Guard Interval cannot be optimized separately for the SC and CGC (by definition of SFN). If it is desirable to optimize these parameters separately for the SC and the CGC, a distinct frequency channel for the SC and the CGC can also be used in SH-A, leading to reduced spectrum efficiency and some handover complications for receivers having only one RF front-end. If the receiver has two RF-front-ends, soft combining (see clause 4.2.6) and easier handover (see clause 9), can also be implemented with SH-A.
- SH-B exploiting TDM transmission mode for the SC and OFDM transmission mode for the CGC. SH-B requires a distinct frequency band for the SC and the CGC since they transmit signals based on two different physical layers. The impacts on the system frequency plan for this physical layer configuration are discussed in the next clause. Each component of the transmission system can be optimized separately to its respective transmission path.

The OFDM waveform is known to exhibit a larger peak-to-average signal envelope fluctuation compared to the TDM waveform. Therefore, SH-A is recommended for spectrum limited systems while SH-B is of interest in power limited satellite systems. Differently from SH-A in SFN configuration, SH-A in *non-SFN configuration* and SH-B also allow an independent optimization of the waveform and LL-FEC parameters for the SC and CGC.

4.1.2 Frequency planning aspects

Being a hybrid satellite/terrestrial system, the frequency planning for DVB-SH is typically more complex than for a pure terrestrial system or a pure satellite system. In fact due to the large geographical coverage provided by the satellite and the possible linguistic segmentation of the market, multibeam satellite may be used in certain areas (e.g. Europe). The use of a multibeam satellite with linguistic areas and/or nation-wide broadcasting requires the segmentation of the DVB-SH downlink spectrum. If the number of beams, N_b , is large enough (e.g. greater than 3), then frequency reuse among satellite beams can be exploited. The multibeam solution brings the following advantages:

a) higher satellite antenna gain providing higher user terminal C/N for the same in-orbit RF power. The higher receiver C/N translates in higher spectral efficiency and/or link margin;

b) frequency reuse between non-adjacent beams in the SC (independently of the existence of the CGC);

c) frequency reuse between the SC and CGC.

To facilitate understanding we will assume in the subsequent that the bandwidth allocated for the satellite transmission is the same for each beam and denote this unit of resource as a frequency "sub-band". Then the overall satellite downlink spectrum is partitioned into f_R sub-bands, with $f_R = 1$ for a single-beam and typically $f_R = 3$,for a multibeam as shown in figure 4.2.

The same sub-band can be reused only by non adjacent satellite beams with a certain reuse pattern that depends on the antenna design. Typically a three-color frequency reuse scheme is sufficient to overcome interference between adjacent beams. The entire available spectrum (F_G in the global beam case) is then split into 3 frequency sub-bands denoted F_A, F_B and F_C. It should be kept in mind that, in the most general case frequency reuse may require up to (but no greater than) four colours.

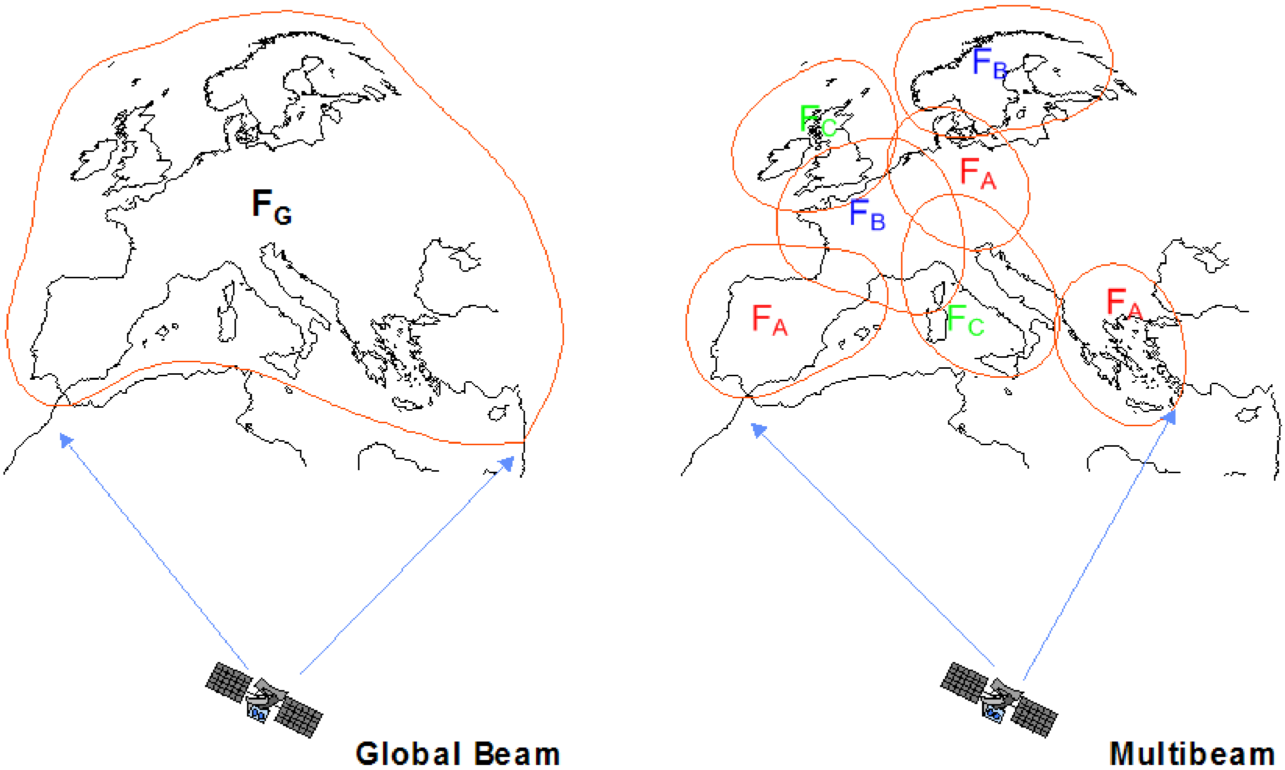


Figure 4.2: Example of satellite single and multibeam

The apparent drawback due to satellite spectrum segmentation is compensated by the fact that the satellite power can be focused over the wanted (and smaller) regions in a multibeam scenario. Furthermore, the global beam scenario would require broadcasting the whole multilingual multiplex to the global footprint, wasting of bandwidth and power in areas where reception of the signal content is not required (or even forbidden).

The resulting spectral efficiency improvement factor is given by:

$$\gamma(N_{beams}, f_R) = \frac{N_{beams}}{f_R}$$

It is observed that for 6 beams there is a doubling of spectral efficiency for frequency reuse factor 3 while for a reuse factor 4 the improvement is limited to 1,5.

The system frequency plan and spectrum efficiency are also impacted by the decision to use or not the SFN technique to complement the satellite coverage with the CGC coverage.

If SFN is used by the CGC to complement the satellite coverage, the common content is transmitted by the CGC using the same sub-band as the satellite signal. By definition of SFN, no additional local content can be added in this sub-band. However, within each beam, the remaining $f_R - 1$ sub-bands, which cannot be used by satellite due to adjacent beams, may be used by the CGC to carry local content.

Alternatively, if SFN is not used, the spectrum must be divided into at least 2 sub-bands: one for the direct-to-user transmission from satellite and the other for the complementary and indirect transmission from the CGC. In this case, this second sub-band may be used to carry additional local content by the CGC since the SFN constraint does not apply. If $f_R > 2$, the remaining $f_R - 2$ sub-bands may be used by the CGC to carry additional local.

When planning additional local content transmission it should be kept in mind that strong interference from satellite into terrestrial may occur in quite large *exclusion zones* (see for example, in figure 4.2, the overlap of the beam centred on France to the south of UK territory). This means that, for sub-bands not reserved for satellite in each beam, the terrestrial transmission parameters and the corresponding terrestrial coverage/capacity cannot be identical within a territory and must be carefully engineered, in coordination with the satellite operator.

If Multi Frequency Network (MFN) is used, different physical-layer configurations will be possible in different areas of a territory (e.g. higher spectral efficiencies modulation and FEC and/or guard-times and/or hierarchical modulation). Therefore MFN allows optimization for different transmitter densities, for different terrestrial coverages and different capacity requirements. In SFN instead, the terrestrial signal complementing the satellite signal has to use exactly and uniformly the common configuration with the satellite signal.

For the various reasons above DVB-SH does not impose the use of SFN for SH-A.

4.1.3 Other considerations for spectrum efficiency

A satellite customer traditionally leases a satellite beam with an associated satellite frequency sub-band. To convert the leased resource into a program offer, he needs a measure of system performance called spectrum efficiency which is given in bits/s/Hz. For example an efficiency of 0,5 bits/s/Hz means that, using 5 MHz satellite bandwidth the customer can build a bouquet of programs with data rates summing up to 2,5 Mbits/s (8 to 10 Mobile TV programs).

In DVB-SH, this customer also gets a frequency "bonus" which is the frequency resource available for potential use terrestrially. This by-product is a virtual asset which becomes a real resource when (i) the customer obtains the authorization from his government authority to use this spectrum terrestrially, and (ii) he deploys/leases the appropriate CGC infrastructure. Therefore in DVB-SH, three spectrum efficiencies can be distinguished, all normalized to the *leased satellite bandwidth*:

- The $\eta[COM]$ is the spectrum efficiency for the "common" programs, per Hz of satellite bandwidth in a beam.
- The $\eta[LOC]$ is the spectrum efficiency for the "local" programs, per Hz of satellite bandwidth in a beam.
- The $\eta[TOT]$, or system spectrum efficiency, is the sum of the above efficiency parameters.

The $\eta[TOT]$ gives an indication of the total programs offer that can be provided with satellite and terrestrial transmissions combined, normalized to the leased satellite bandwidth.

Note that in the "exclusion zones", coexistence of terrestrial transmissions with satellite transmissions in adjacent beams can impose lower $\eta[LOC]$. Also, in MFN between satellite and CGC, the OFDM guard-time can be engineered differently from one local zone to another, resulting in different $\eta[LOC]$. Therefore $\eta[LOC]$ and consequently $\eta[TOT]$ should be used as local performance indexes and not a nation-wide index.

To evaluate and/or optimize DVB-SH system spectral efficiency as defined above, first the following must be determined/computed.

Table 4.2

Parameter/ Questions	Explanation	Main Dependencies	General target values
N_{beams}	The number of satellite beams	Number of countries/linguistic regions, satellite antenna sub-system cost, inter-beam interferences	>=3 (for regional coverage) (for large countries, Nbeams = 1)
f_R	Satellite frequency reuse over the beams (also called number of "colours")	N_{beams}, national/ linguistic boundaries, satellite power, frequency allocation and channelization.	3
Is there regulatory limitations for local content?	Local content may be forbidden or its amount restricted.	Regulatory context	Country dependent
Is SFN used by CGC to complement satellite?	For SH-B, the answer is negative by definition. For SH-A it is possible to choose between SFN and MFN	SH-A/SH-B, handover complexity, chipset availability	Equally likely
Does the satellite allow nonlinear operation of TWTA?	In general yes for single beam satellite or in markets where in-orbit re-configurability of power and coverage is not a must.	Market definition	Equally likely
$\eta_{OFDM}[SAT]$	Spectral efficiency of the OFDM waveform when used by satellite, given minimum terminal characteristics and Quality of Service.	Satellite effective EIRP, Terminal characteristics, required link margin (QoS.), allowed end-to-end max delay.	0,5 b/s/Hz
$\eta_{OFDM}[TER]$	Spectral efficiency of the OFDM waveform when used by CGC, given minimum terminal characteristics, coverage type (indoor/outdoor) and Quality of Service.	CGC network cost, indoor-outdoor operations, terminal types, required link margin (QoS.)	0,7 b/s/Hz
Total payload equivalent signal power degradation D_{TOT}	Total payload losses due to payload nonlinear effects (output back off, signal distortion, intermodulation noise)	Payload characteristics, physical layer settings, required margin (QoS), interference scenario	1 dB to 1,5 dB for single carrier TWTA operation, 2,5 dB to 3,5 dB for multi-carrier TWTA operation

Several frequency plans are possible depending on the SH-A or SH-B architecture and SFN or MFN configurations. Clause C.1 provides examples and a methodology for spectral efficiency computations and comparisons.

4.1.4 Frequency bands and their regulatory aspects

4.1.4.1 Frequency Allocations and coordination procedures

Frequency spectrum is regulated at the international level by a binding treaty called the Radio Regulations. The Radio Regulations deal with two aspects: frequency allocation and regulatory procedures for accessing the spectrum/orbit resources. The binding character of the Radio Regulations implies that national or regional (e.g. European) regulations shall be defined within the framework of this treaty.

Frequencies between 1 GHz and 3 GHz are the most suitable, considering the satellite, terminal and mobility constraints. Within this frequency range, the following bands are candidates for the provision of multimedia services based on DVB-SH (the uplink bands are given for completeness and their use are out of the scope of the DVB-SH specifications).

Table 4.3

<table>
<tr><th>Frequency band designation</th><th>Frequency range</th><th>Others common names</th></tr>
<tr><td rowspan="3">L band</td><td>1 626,5 MHz to 1 660,5 MHz, 1 668 MHz to 1 675 MHz (uplink)
1 518 MHz to 1 559 MHz (downlink)</td><td>MSS GEO
L band</td></tr>
<tr><td>1 610 MHz to 1 626,5 MHz (up and downlink)</td><td>MSS Big LEO
L band</td></tr>
<tr><td>1 452 MHz to 1 492 MHz (downlink)</td><td>S-DAB band</td></tr>
<tr><td>2 GHz S band</td><td>1 980 MHz to 2 010 MHz (uplink)
2 170 MHz to 2 200 MHz (downlink)</td><td></td></tr>
<tr><td>S-DARS S band</td><td>2 320 MHz to 2 345 MHz (downlink)</td><td></td></tr>
<tr><td rowspan="2">2,5 GHz S band</td><td>2 670 MHz to 2 690 MHz (uplink)
2 500 MHz to 2 520 MHz (downlink)</td><td></td></tr>
<tr><td>2 520 MHz to 2 670 MHz (downlink)</td><td></td></tr>
</table>

4.1.4.1.1 L band

Frequency allocations

In the field of satellite communications, the term "L band" encompasses three frequency bands:

- the bands 1 626,5 MHz to 1 660,5 MHz and 1 668 MHz to 1 675 MHz (uplink)/1 518 MHz to 1 559 MHz (downlink): these bands are allocated almost worldwide to the mobile-satellite service and are mainly used by incumbent MSS operators using geostationary satellites to provide two-ways mobile satellite communications;
- the band 1 610 MHz to 1 626,5 MHz (up and downlink): this band is allocated worldwide to the mobile-satellite service and is used by incumbent MSS operators using non-geostationary satellite systems in a low Earth orbit (LEO) to provide the same type of services as in the previous bands. It should be noted that one non-geostationary system uses this band only for its uplinks (its downlink using the band 2 483,5 MHz to 2 500 MHz);
- the band 1 452 MHz to 1 492 MHz (downlink): this band is allocated worldwide (with the only exception of the USA) to the broadcasting-satellite service but only the upper 25 MHz (i.e. 1 467 MHz to 1 492 MHz) can actually be used since the lower 15 MHz are not usable until a decision by a future ITU World Radiocommunications Conference. This band is also allocated on a co-primary basis with the broadcasting-satellite service to the terrestrial fixed, mobile and broadcasting services. The use of this band by the broadcasting-satellite and terrestrial broadcasting services is limited to digital audio broadcasting.

Access to the spectrum/orbit resources

In the bands 1 518 MHz to 1 559 MHz, 1 610 MHz to 1 626,5 MHz, 1 626,5 MHz to 1 660,5 MHz and 1 668 MHz to 1 675 MHz, geostationary satellites and non-geostationary systems are on an equal footing with regards to spectrum access (see provision No. 9.11A of the Radio Regulations [20]). The basic principle governing the coordination under No. 9.11A [20] is "first come, first served".

In the band 1 467 MHz to 1 492 MHz (since the band 1 452 MHz to 1467 MHz is de facto unusable), geostationary satellites have regulatory priority with regards to non-geostationary systems (see provision No. 22.2 of the Radio Regulations [20]). Among themselves, geostationary satellites are required to coordinate according to Radio Regulations provision No. 9.7 [20] ("first come, first served" but only between geostationary satellites). Moreover, all systems are required to coordinate with terrestrial services under provision No. 9.11 [20]. It should be noted that the protection of terrestrial services in the United States of America renders operations of a geostationary satellite in visibility of the US territory very challenging if not unfeasible.

4.1.4.1.2 2 GHz S band (1 980 MHz to 2 010 MHz/2 170 MHz to 2 200 MHz)

Frequency allocations

At the international level, the band 2 170 MHz to 2 200 MHz is allocated worldwide to the fixed, mobile and mobile-satellite (space-to-Earth) services on a co-primary basis. It means that, from a regulatory point of view, these three services enjoy the same status. It does not mean that applications provided through these services are necessarily compatible from a technical point of view (i.e. that they can use the same band without interfering each other). On the contrary, numerous studies have shown that terrestrial mobile applications and satellite mobile applications are not compatible if they are operated independently in the same geographical area. Complementary ground components (CGC) are compatible with the provision of satellite applications to mobile terminals since they are operated in a coordinated manner with the satellite system.

It should be noted that in the Americas, the adjacent band 2 160 MHz to 2 170 MHz is also allocated, inter alia, to the mobile-satellite service (space-to-Earth).

Some countries have chosen to favour terrestrial systems in these bands: in Algeria, Benin, Cape Verde, Egypt, Iran, Mali, Syria and Tunisia, the use of these bands by the mobile-satellite service shall neither cause harmful interference to the fixed and mobile services, nor hamper the development of those services prior to 1 January 2005, nor shall the former service request protection from the latter services. While this has not a direct impact on the overall design and operations of MSS systems intended to cover Europe, it may have an impact on the design of satellite antennas compatible with the protection of terrestrial services in these bands.

The fact that a country is not listed above does not imply that it intends to give priority to satellite services but simply that it has chosen to keep satellite and terrestrial services on an equal footing from the point of view of international regulations. National regulators may chose to restrict nationally the use of these bands to a specific service. For more information on the regulatory status of these bands, the national table of frequency allocations shall be consulted.

NOTE: National tables of frequency allocations of most of European countries can be accessed in: www.efis.dk.

Access to the spectrum/orbit resources at the international level

In these bands, geostationary satellites and non-geostationary systems are on an equal footing with regards to spectrum access (see provision No. 9.11A of the Radio Regulations [20]). The basic principle governing the coordination under No. 9.11A [20] is "first come, first served".

Moreover, in the band 2 170 MHz to 2200 MHz, a coordination procedure between transmitting space stations and terrestrial services is established through the provision No. 9.14 of the Radio Regulations [20]. The PFD thresholds used to determine whether there is a need to coordinate with terrestrial services in any country are:

Table 4.4

	For geostationary satellites	**For non-geostationary satellites**
For $0° \leq \delta \leq 5°$	-128 dBW/m²/MHz	-123 dBW/m²/MHz
For $5° \leq \delta \leq 25°$	-128 + 0,5(δ - 5) dBW/m²/MHz	-123 + 0,5(δ - 5) dBW/m²/MHz
For $25° \leq \delta \leq 90°$	-118 dBW/m²/MHz	-113 dBW/m²/MHz
NOTE: where δ is the angle of arrival (degrees).		

It should be noted that these pfd levels are not compatible with MSS operations towards handheld terminals. Therefore this coordination threshold will always be exceeded over the territories of countries within the service area of the MSS system. It will not be a constraint as such, since it is widely recognized that such systems can not share the same geographical area with terrestrial services. Therefore countries willing to encourage the provision of MSS services in these bands over their territory have taken steps to free the bands from any terrestrial systems. However, the same does not apply in countries not belonging to the MSS service area. On their territories, these administrations may wish to continue to deploy terrestrial services as planned and may require protection from MSS transmitting satellites. This request for protection has to be explicit: following the publication of the MSS system special section by the ITU, any administration considering itself as potentially affected has four months to explicitly request to be included in the coordination process. Otherwise, it is deemed to be unaffected. This mechanism substantially reduces the number of administrations actually included in the process of coordination with respect to their terrestrial services.

4.1.4.1.3 Sound Broadcasting Satellite (SDARS) S band (2 310 MHz to 2 360 MHz)

The band 2 310 MHz to 2 360 MHz is allocated to the sound broadcasting-satellite service in only three countries: India, Mexico and the USA. Only in the latter is the band currently used for such service. More particularly, only 25 MHz are available in the USA for such a service (2 320 MHz to 2 345 MHz). This 25 MHz band was divided into two blocks of 12,5 MHz and auctioned by the FCC in April 1997. At the time of writing these guidelines, the two blocks were fully used by their respective owners for the provision of digital radio via satellite supplemented by ancillary ground repeaters in areas where the satellite signal is too weak to be received with the required quality. These ground repeaters are using part of the frequencies available in the 12,5 MHz blocks.

4.1.4.1.4 2,5 GHz S band (2 500 MHz to 2 690 MHz)

The band 2 500 MHz to 2 690 MHz includes various allocations to satellite services. Of particular interest for the introduction of DVB-SH based services are the following:

- 2 500 MHz to 2 520 MHz (downlink)/2 670 MHz to 2 690 MHz (uplink): these paired bands are allocated worldwide to the mobile-satellite service (see note 1) and are shared, inter alia, with terrestrial (fixed and mobile) services. As for the bands 1 980 MHz to 2 010 MHz/2 170 MHz to 2 200 MHz (see above), coordination under No. 9.11A [20] (i.e. "first come, first served") applies in these bands and therefore geostationary satellites and non-geostationary systems are on an equal footing with regards to spectrum access. Similarly to the band 2 170 MHz to 2 200 MHz, in the band 2 500 MHz to 2 520 MHz, a coordination procedure between transmitting space stations and terrestrial services is established through the provision No. 9.14 of the Radio Regulations [20]. The pfd thresholds used to determine whether there is a need to coordinate with terrestrial services in any country are -128 MHz/-118 dBW/m²/MHz depending on the elevation angle.

NOTE 1: With the exception of Azerbaijan, Bulgaria, Kyrgyzstan and Turkmenistan.

- 2 520 MHz to 2 670 MHz: this band is allocated to the broadcasting-satellite service and is also shared, inter-alia, with terrestrial (fixed and mobile) services.

It should be noted that the European Conference of Postal and Telecommunications administrations (CEPT) has decided in 2005 that the frequency band 2 500 MHz to 2 690 MHz is designated for terrestrial IMT-2 000/UMTS systems only (see note 2). This band is therefore not suitable for the provisions of satellite services within Europe.

NOTE 2: See ECC Decision of 18 March 2005 on the harmonised utilisation of spectrum for IMT-2000/UMTS operating within the band 2 500 MHz to 2 690 MHz (ECC/DEC/(05)05 [i.25]).

4.1.4.2 Regulatory aspects

4.1.4.2.1 Europe

European Union

L band and 2,5 GHz S-band:

No specific regulatory instruments on the designation of the frequency bands for satellite use have been adopted by the European Union in these bands.

2 GHz S-band:

On 14th February 2007, the European Commission adopted a Decision "on the harmonized use of radio spectrum in the 2 GHz frequency bands for the implementation of systems providing mobile satellite services" (see note) which gives priority to the development of MSS in the 2 GHz bands. Pursuant to the decision, Member States will be obliged to designate and make available as of 1 July 2007 the 2 GHz S-bands for systems providing mobile satellite services, which can include the use of complementary ground components. Contrary to ECC Decisions (see the following clause), a European Decision is mandatory for all Member States of the European Union. It gives therefore more regulatory certainty for the use of these bands.

NOTE: This Decision 2007/98/EC [i.26] was published in the Official Journal of 14th February 2007.

European Conference of Postal and Telecommunications administrations (CEPT)

L band:

CEPT has decided in 2004 that the frequency bands 1 518 MHz to 1 525 MHz and 1 670 MHz to 1 675 MHz are designated to the mobile-satellite service from 1 April 2007 (see note 1). While no specific CEPT instruments designate the frequency bands 1 525 MHz to 1 559 MHz and 1 6101 MHz to 660,5 MHz to the mobile-satellite service, other CEPT decisions on free circulation and use or on licence exemption of terminals imply that these bands are, de facto, designated to such service.

NOTE 1: See ECC Decision of 12 November 2004 on the designation of the bands 1 518 MHz to 1 525 MHz and 1 670 MHz to 1 675 MHz for the Mobile-Satellite Service (ECC/DEC/(04)09 [i.27]).

CEPT has decided in 2003 that the frequency band 1 479,5 MHz to 1 492 MHz is designated for use by satellite digital audio broadcasting systems (see note 2). It should be noted that the band 1 467 MHz to 1 479,5 MHz is part of the Maastricht Plan for the introduction of terrestrial T-DAB systems in L band. The use by satellite systems of this lower band in Europe is generally not compatible with T-DAB systems.

NOTE 2: See ECC Decision of 17 October 2003 on the designation of the frequency band 1 479,5 MHz to 1 492 MHz for use by Satellite Digital Audio Broadcasting (S-DAB) systems (ECC/DEC/(03)02 [i.28]).

No specific regulatory instruments have been adopted by the CEPT in these bands concerning complementary ground components.

2 GHz S band:

In December 2006, the European Communications Committee (ECC) adopted two Decisions and one Recommendation designed to foster the regulatory certainty of the 2 GHz bands.

- Band designation for MSS systems, including those supplemented by a Complementary Ground Component (CGC) (see note 1): the bands 1 980 MHz to 2 010 MHz and 2 170 MHz to 2 200 MHz are designated for systems of the mobile-satellite service. Since no other ECC Decision exists to designate these bands for other services, it means that these bands are exclusive for MSS systems. The same Decision explicitly mentions the possibility for MSS systems in these bands to incorporate a complementary ground component (CGC). This explicit mention gives more regulatory certainty for deploying such ground-based infrastructure across the whole Europe.

NOTE 1: See Decision ECC/DEC/(06)09 [i.29].

- Harmonized regulatory conditions for complementary ground components: within the same Decision, CGC is defined as follows: "CGC is an integral part of a mobile satellite system and consists of ground based stations used at fixed locations to improve the availability of the mobile satellite system in zones where the communications with one or several space stations cannot be ensured with the required quality. CGC uses the same portions of the mobile-satellite service frequency bands (1 980 MHz to 2 010 MHz/2 170 MHz to 2 200 MHz) as authorized for the associated space station(s)" Moreover, some conditions are attached to the use of CGC:
  - the frequency band to be used by the CGC of a particular satellite system shall be accommodated within the same frequency band used by the satellite component of that satellite system;
  - the use of CGC shall not increase the spectrum requirement of the satellite component of that particular mobile satellite system;
  - the CGC shall only be deployed in the geographical areas where the mobile earth stations of the associated MSS system are authorized to operate;
  - the same direction of transmission by CGC and the satellite component shall be used so as to decrease the number and complexity of compatibility issues;
  - the satellite segment shall be re-established as soon as possible in case of failure of the satellite segment, and no later than 18 months after such a failure, unless justified otherwise on considerations based on reasonableness and/or proportionality. Otherwise, CGC shall cease operation;
  - compatibility with terrestrial IMT-2 000/UMTS operational systems in adjacent bands should be ensured;

- the CGC shall not operate independently from the satellite resource/network management system.

- Spectrum reframing of the 2 GHz bands (see note 2): in addition to the Decision designating the 2 GHz bands for MSS, the ECC also agreed on a Decision on "transitional arrangements for the fixed service and tactical radio relay systems in the bands 1 980 MHz to 2 010 MHz and 2 170 MHz to 2 200 MHz in order to facilitate the harmonized introduction and development of systems in the mobile satellite service including those supplemented by a complementary ground component". This Decision is threefold:
  - finalizing the transitions of existing fixed service systems and tactical radio relay systems operating in or overlapping the 1 980 MHz to 2 010 MHz and 2 170 MHz to 2 200 MHz bands;
  - not implementing any new FS networks and tactical radio relay systems which are incompatible with MSS operations in the bands 1 980 MHz to 2 010 MHz and 2 170 MHz to 2 200 MHz;
  - not requiring any longer coordination of MSS systems with respect to fixed service and tactical radio-relay systems (see section 4.1.3.1.2, coordination under No. 9.14 [20]).

NOTE 2: See Decision ECC/DEC/(06)10 [i.30].

- Recommendation on Milestone Review Process (see note 3): at its December 2006 meeting, the ECC adopted a Recommendation establishing a Milestone Review Process (MRP) in order to gather precise information about the development of the various satellite projects intending to use the bands. This ECC Recommendation lists the milestones for which MSS systems proponents are invited to submit information in order to monitor the development of projects in these bands. It was initially envisaged to use this MRP as a binding process to choose MSS systems eligible for a frequency authorization in these bands; however the lack of legal power of CEPT to enforce such a process was deterrent to its implementation. The MRP is now only of an informative nature. All declarations are based on good faith and will not be checked against. Any binding decision should be transferred to the European Commission. The milestones consists in evidence of submission of ITU request for co-ordination, satellite manufacturing contract, completion of the Critical Design Review of the MSS system, satellite launch agreement, gateway earth stations procurement, successful satellite mating, successful satellite launches, completion of frequency co-ordination as well as actual provision of satellite service within the territories of CEPT countries.

NOTE 3: See Recommendation ECC/Recommendation (06)05 [i.31].

2,5 GHz S band

As mentioned above, CEPT has decided in 2005 that the frequency band 2 500 MHz to 2 690 MHz is designated for terrestrial IMT-2 000/UMTS systems only (see note). This band is therefore not suitable for the provisions of satellite services within Europe.

NOTE: See ECC Decision of 18 March 2005 on the harmonised utilisation of spectrum for IMT-2 000/UMTS operating within the band 2 500 MHz to 2 690 MHz (ECC/DEC/(05)05 [i.26]).

4.1.4.2.2 North America (all bands)

In the United States of America, the available bandwidth is limited by the Table of Frequency Allocations to 20 MHz (i.e. 2 180 MHz to 2 200 MHz coupled with the uplink band 2 000 MHz to 2 020 MHz). In these bands, ancillary terrestrial components may be operated in conjunction with MSS networks, subject to the Commission's rules for ancillary terrestrial components (ATC) and subject to all applicable conditions and provisions of the MSS authorization. The FCC generally imposes to an MSS licensee that wishes to include ATC to meet five requirements (see note 1):

- geographic coverage: an MSS licensee must provide space-segment service across the entire geographic area where the ATC can be deployed;

- coverage continuity: MSS operators must maintain space station coverage over the relevant geographic area, which implies timely replacement of satellites in the event coverage should degrade as a result of satellite failure. In order to implement this condition, a non-geostationary MSS system licensee is required to maintain an in-orbit spare, while a geostationary MSS system licensee is required to maintain a spare satellite on the ground within one year of commencing operations and launch it into orbit during the next commercially reasonable launch window following a satellite failure;

- commercial availability: the MSS service via satellite must be commercially available as a perquisite to any offering of the ATC service;

- an integrated offering: MSS licensees must offer an integrated service. MSS licensees must make an affirmative showing to the FCC that demonstrates that their ATC service offering is truly integrated with their MSS offering. As an example, MSS licensees that wish to provide ATC services could demonstrate that they use a dual-mode handset to provide the proposed ATC service; and

- in-band operation: the ATC operations must remain limited to the precise frequency assignments authorized for the MSS system.

NOTE 1: For more information, see Federal Communications Commission - Report and Order and Notice of Proposed Rulemaking 03-15 - In the matter of Flexibility for Delivery of Communications by Mobile Satellite Service Providers in the 2 GHz Band, the L-Band, and the 1,6/2,4 GHz Bands (released 10th February 2003).

In Canada, the available bands for mobile-satellite services are 2 165 MHz to 2 200 MHz (space-to-Earth) and 1 990 MHz to 2 025 MHz (Earth-to-space). Industry Canada has issued generic regulations for Ancillary Terrestrial Components (ATC) (see note 2) that can be used in conjunction with MSS systems in the frequency range 1 GHz to 3 GHz. These regulations are based on the following principles:

- the ATC mobile service will be an integral part of MSS service offerings. A substantial level of mobile-satellite services will be provided with the ATC service;

- the frequencies used for the ATC system will be within the assigned spectrum for a particular MSS network and the ATC service will be limited to the satellite serving areas. The use of the MSS spectrum for ATC operation will be subordinate to the spectrum being available for mobile-satellite service;

- the ATC mobile service will be required to cease operation, within a reasonable period, should the mobile-satellite service or network be discontinued;

- the ATC operation will be authorized such that it will neither cause harmful interference to, nor claim protection from, MSS services and other primary radio services operating in adjacent bands. ATC operations will be subject to technical and operational requirements considered appropriate to mitigate potential interference;

- complete applications as radiocommunication carriers will need to be submitted to seek authorization to operate an ATC mobile system as an integral and indefeasible part of the MSS service offerings. Specific information will be required as part of the applications to demonstrate adherence to policy, operational and regulatory principles;

- spectrum area licences will be issued for ATC systems and will be subject to spectrum fees.

NOTE 2: See Industry Canada - Notice No. DGTP-006-04 - Spectrum and Licensing Policy to Permit Ancillary Terrestrial Mobile Services as Part of Mobile-Satellite Service Offerings (May 2004).

4.2 Main issues addressed by DVB-SH

4.2.1 Differences in propagation characteristics between satellite and terrestrial channels

For frequency below 3 GHz, the only significant effects are those caused by the environment close to the user terminal as atmospheric effects are negligible (see ITU-R Recommendations P.676 [25], ITU-R Recommendation P.618 [31], ITU-R Recommendation P.531 [32] and ITU-R Recommendation P.680 [33]).

For DVB-SH, three main mobile environments may be considered:

- *rural environment* : the propagation is mainly affected by the vegetation. The coverage is mainly provided by the SC;

- *urban environment* : the propagation is mostly affected by dense buildings or other constructions with height of 4 storeys or more. The coverage is mainly provided by the CGC; and

- *suburban environment* : representing an intermediate case with medium density of buildings, lower structures (2 to 3 storeys) and roads which are wider than in an urban environment. The SC and the CGC contribute to the desired coverage. Small villages may be treated as suburban areas where, if the population density is low, satellite may be the only source for service provision.

For both LMS and terrestrial mobile channels, the effects are conventionally divided into three types according to the scale of distances to be considered:

- *path loss at large scale* (very slow fluctuations): the signal suffers variations due to modifications in the geometry of the propagation path. This loss is usually assumed to be proportional to d^n, d being the distance from the transmitter to the user terminal and n being an empirical exponent, based on theory and measurements;

- *shadowing at mid-scale* (slow fluctuations): it corresponds to amplitude variations due to nearby obstacles on the ground. The signal suffers variations due to obstructions caused by buildings, trees etc., the scale here being similar to the dimensions of these obstacles; and

- *multipath fading at small scale* (fast fluctuations): the scale of the variations of the signal is about one wavelength, as a result of the constructive or destructive addition of multiple paths. For wideband systems, it is necessary to consider the multipath fading as frequency selective. The satellite propagation channel in DVB-SH is generally non-frequency selective. This assumption is very accurate for 1,5 MHz, pretty correct for 5 MHz and fairly correct for 8 MHz channelization. It means that the satellite propagation channel can be considered as a single complex multiplicative process. Instead the terrestrial repeaters propagation channel is considered to be frequency selective because of their respective coherence bandwidths. A frequency selective fading is classically characterized through a Power Delay Profile (PDP) which gives the relative time of arrival, the relative power and the type (Ricean or Rayleigh distribution, spectrum) of each group of unresolved echoes (also called tap). These PDP are then used to parameterize Tapped Delay Line (TDL) models. For SFN operation between satellite and terrestrial OFDM (SH-A, SFN), all contributions have to be taken into account.

In summary, the LMS channel has very different characteristics compared to the terrestrial one. In particular, the channel can be considered as non frequency selective and link margins are not as large as for terrestrial networks. More specifically, the satellite distance from the user is basically constant within the beam and the satellite power limitations make the link margin bounded to a value typically ranging from 5 dB to 15 dB.

4.2.2 LMS channels (in the L and S bands)

The LMS channel models can be grouped into three classes:

1) *empirical models* are obtained from experimental data. They are very close to reality for the environment type in which the measurements have been done but are difficult to generalize to other environment types;

2) *statistical models* are based on the use of canonical statistical distributions. Like empirical models, statistical models are applied to environment classes (rural, suburban, urban, etc.). The subsequent classification problem is not straightforward, since an environment classified as urban in some countries may look a little more like small town elsewhere. Therefore statistical models are also difficult to generalize; and

3) *physical models* rely on a deterministic modelling of the propagation phenomena (reflection, diffraction, refraction), but also of the considered environment. These models have been efficiently used for planning purposes in terrestrial radio-communication or broadcast networks.

In the following the main models belonging to the two first categories are presented, the third being mainly used for terrestrial networks.

4.2.2.1 Empirical Models

The ERS (Empirical Road Side) model was elaborated based on numerous experiments performed since 1983 by J.Goldhirsh and W.J.Vogel [i.2]. This model mathematically describes attenuation from tree-lined roads, multipath in mountainous and tree lined roads. It also provides a frequency scaling relationship for static tree attenuation from UHF to K Band. This model is recommended by ITU for rural environment (see ITU-R Recommendation P.681-4 [12]). Examples of ERS model link margin calculations for S-band (2,2 GHz) are reported in table 4.5.

Table 4.5: Additional link margin versus elevation angle for rural landscapes according to the ERS model, at f_c = 2,2 GHz

Elevation angle	Margin for 80 % availability [dB]	Margin for 95 % availability [dB]	Margin for 99 % availability [dB]
41,5°	5	12	20
39°	6	13,5	21,5
30°	9,5	17,5	26,5
25°	12,5	20	29

It can be seen that by using an ERS model, calculation based on instantaneous signal availability results in a link margin that, for availability greater or equal than 95 %, is incompatible with practical satellite capabilities. As shown in clause A, by using the DVB-SH waveform features (powerful FEC, long time interleaving at physical layer or inter-burst upper layer FEC) one can achieve high availability with smaller margins. This is possible as short term link unavailability is recovered through physical FEC and LL-FEC. Thus the DVB-SH link margins shall be based on a more elaborated satellite channel model able to represent the first and the second order statistics of the fading/shadowing process. Therefore the ERS model is not recommended for DVB-SH network planning.

4.2.2.2 Statistical models (for narrow-band signals)

The statistical channel models illustrated in the following clause are capable of generating time series of the LMS signal amplitude fading/shadowing process through a propagation model that can be readily implemented in a computer simulation or in a laboratory test environment. The statistical LMS propagation model is characterized by parameters that have been derived through synthetic time series matching with experimental data obtained in different LMS propagation environments. The experimental data matching corresponds to first and second order statistics of the signal attenuation, thus is in principle suitable to our DVB-SH performance assessment scope.

4.2.2.2.1 Single-state models

Statistical models for satellite channels make the assumption that the received signal is composed of two parts: a coherent part associated with the LOS path and a diffuse part arising from multipath components. Table 4.6 summarizes the best known statistical satellite mobile channels in increasing order of complexity.

Table 4.6: Statistical satellite mobile channel models

Model	Coherent part	Correlation	Diffuse part
Rice	Constant	Zero	Rayleigh
Loo [i.4]	Log-normal	Variable	Rayleigh
Corazza [i.5]	Log-normal	Unity	Log-normal, Rayleigh
Hwang [i.6]	Log-normal	Zero	Log-normal, Rayleigh

The Rice model assumes that the LOS component is only affected by Rayleigh distributed multipath and is characterized by the carrier-to-multipath ratio (C/M or Rice factor K). The Loo model assumes that the LOS is affected by lognormal shadowing (typical for tree shadowing) which makes the instantaneous C/M time variant. The Corazza model assumes that both the LOS and the multipath components are affected by the same lognormal shadowing. Thus in the Corazza model the C/M is constant although the LOS component is lognormal distributed. The Hwang model also considers lognormal process affecting both the LOS and the multipath component but with total decorrelation between them. The Hwang model has been shown to include the Rice, Loo and Corazza models as special cases.

As DVB-SH is supposed to operate in different mobile environments, where channel conditions are time variant, a single-state channel model is considered inadequate.

4.2.2.2.2 Two-state (Lutz) model

The Lutz model [i.3] is a statistical model which represents the channel by a two states Markov chain. The "good" state occurs when a dominant line-of-sight component is received. In that state, the channel can be considered as a Ricean channel. The "bad" state occurs when no LOS component is received. In this state, the channel can be modelled as a Log-normal Rayleigh channel without coherent component. The parameters associated with each state and the transition probabilities from one state to another are empirically derived. This approach allows the generation of time series representing different environments. However, two-state channels were considered not enough to represent the variety of propagation conditions experienced in the different environments.

4.2.2.2.3 Three-state (Fontan) Model

A further refinement of the Lutz model is represented by the Fontan model [16] which includes a three-state Markov chain:

- state 1: line-of-sight (LOS);
- state 2: moderate shadowing; and
- state 3: deep shadowing.

The statistical model adopted for all three states is a Rice/lognormal distribution with parameters that have been experimentally derived from propagation campaigns performed in different European locations. The model parameters are reported in [16] for different environments, elevation angles and locations. Being recognized as the most accurate statistical LMS channel model available today, encompassing the widest set of environments and elevation angles, this model has been adopted for the DVB-SH performance evaluation detailed in annex A.

Figure 4.3 shows a number of time series generated by using the Fontan's model for various environments and satellite elevations angles at S-band. For open area the received signal power fluctuations are rather limited except for the 20 degrees elevation.

The situation is radically different for S-band intermediate tree-shadowed (ITS) environment. In this case the multi-state channel nature is evident in particular up to 40 degrees elevation. When no LOS signal is present large shadowing fluctuations on top of multipath fading are visible. Suburban areas shows similar two-state nature with longer LOS condition compared to the ITS case. Urban environment keeps a remarked on-off channel nature even at high satellite elevation angles due to the presence of close buildings.

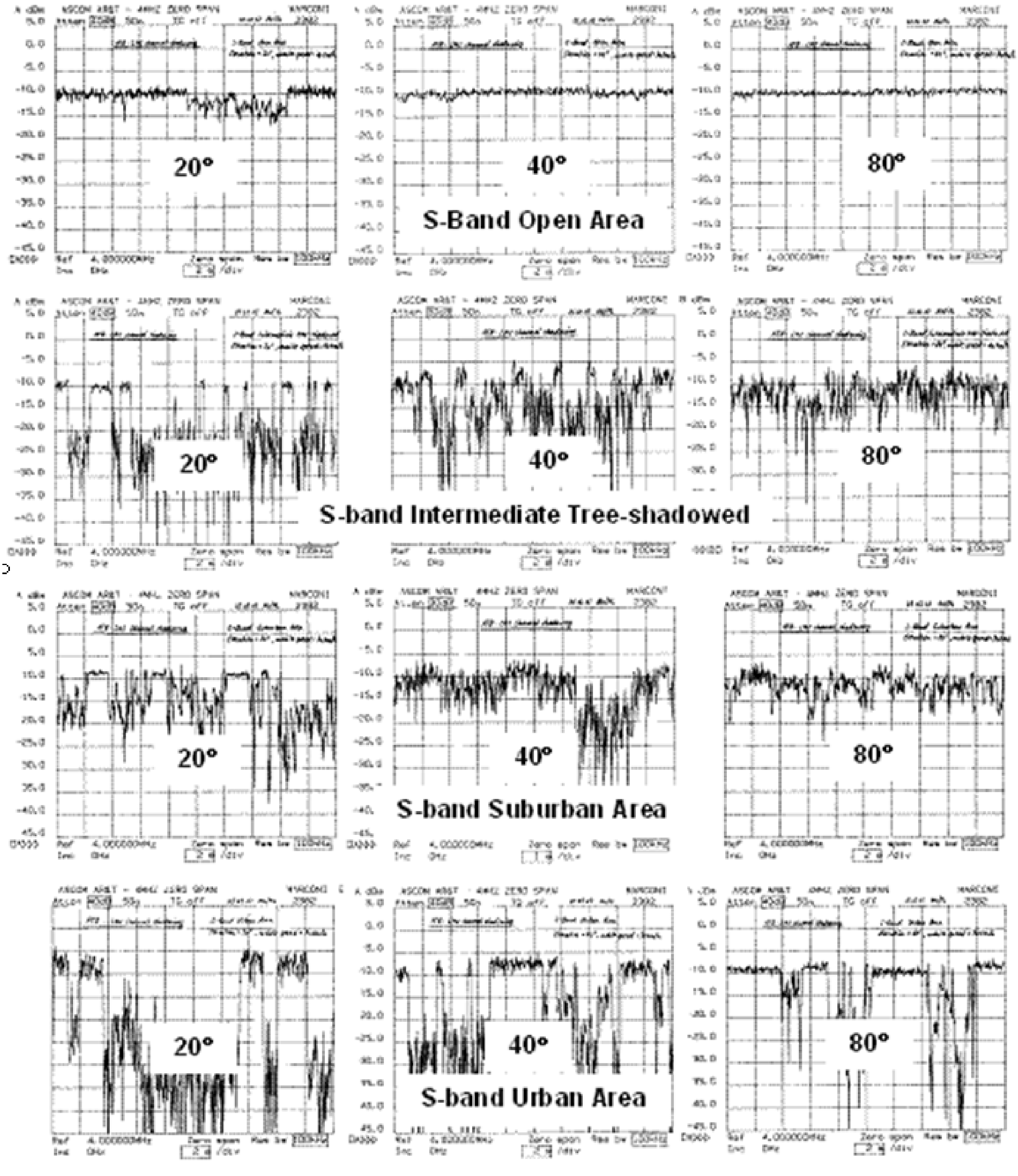


Figure 4.3: Examples of Fontan's channel model time series

4.2.2.2.4 Quasi-static Channel Model

The use of DVB-SH for direct satellite operation reception by hand-held terminals operating in quasi static conditions with user cooperation calls for an extra channel model. This channel model corresponds to the line of sight state of the Fontan model described before. Being the terminal quasi static, the resulting channel can be described by a very slow Ricean fading process (see annex A). Being the fading correlation time larger than the physical layer or upper layer decorrelation capabilities, link availability can be computed according to first order fading statistics as described in clause 11.

4.2.3 Terrestrial Channels

Path loss: several models have been developed to describe path loss for a single terrestrial transmitter in both urban and suburban environments in the past [i.1]. Two of them have been used in contexts close to the DVB-SH one:

a) the COST 231-Hata classically used for terrestrial broadcast network planning;

b) the Xia Bertoni model used for the assessment of UMTS performances (see TR 101 112 [i.24] and ITU-Recommendation M.1225 [11].

Shadowing effects: for a single terrestrial transmitter, shadowing is generally considered to be lognormal distributed. This lognormal law is characterized by its median value (corresponding to the path loss) and by its standard deviation or location variability, σ_L. Location variability is generally higher in suburban areas than in urban areas. For outdoor users in urban and suburban environments, the value of σ_L recommended by ITU-Recommendation M.1225 [11] is 10 dB at 2,2 GHz. However, in urban environment, this value can be slightly pessimistic: models described in [i.1] give a value of about 8,1 dB in urban environment and of 9,5 dB in suburban environment. Nevertheless, other references that are used by Mobile Telephony Operator also propose 8 dB for suburban areas (see clause 11). On the other hand, it is acknowledged that the value of 5,5 dB for σ_L is commonly used by broadcasters [13], [19] and [i.21]. So the proposed values for network planning in clause 11 are σ_L = 5,5 dB and σ_L =8 dB, for the "broadcasters' approach" and the "cellular approach" respectively. TM-SSP plans to publish refreshed recommendations on this parameter as soon as experimental results are made available.

Physical layer performance assessment:

Single cell environment: for physical layer performance assessment in terrestrial environment DVB-SH in common to DVB-H has selected the TU6 model [18]. The Typical Urban 6-paths model (TU6) is proven to be representative of the typical mobile reception with Doppler frequency above 10 Hz. In TU6 the channel model consists of tapped delay line (TDL) made of 5 delayed plus one non delayed path each affected by Rayleigh distributed fading with different average power. The Rayleigh fading bandwidth is dependent on the mobile speed.

Although very popular the TU6 channel shows limitations when used to represent SFN type of networks, indoor operations and low speed mobile conditions. Low fading effects (also called large scale fading) modelled for example with lognormal fading has not been included in the simulation setups for the evaluation of the terrestrial reception. These effects are covered by the margin added by the network planning. For the satellite reception the slow and very slow fading effects are covered by the LMS models. The evaluation of the fading characteristics for terrestrial networks typical for DVB-SH is subject of several on-going projects and field trials. In the future these models may be used to provide complementary performance data.

It should be remarked that Rayleigh and Rice channel models definitions are also used in the context of DVB-T/H but they refer to two different contexts as explained detailed in clause A.7.

Multi-cell environments: the DVB-SH terrestrial re-transmitter network is characterized by Single Frequency Network (SFN) architecture. This characteristic must be taken into account when assessing the system performances in realistic propagation conditions when coverage is fully or partly ensured by gap-fillers as it is the case in urban and suburban environments. Contrary to the satellite-only case, the terrestrial propagation channel must be considered as frequency selective. Therefore, the assessment of waveform performances must be performed using a TDL model parameterized with a Power Delay Profile giving the stationary impulse response of the propagation channel. This PDP for such SFN configurations can be obtained through two types of methods:

- for a given site for which a 3D representation is available, a site-specific deterministic tool can be used (ray launching or ray tracing tools). For a generic urban/suburban context, a macro-cellular geometric configuration is set for a given cell radius. Then, given a position of the user terminal, the channel SFN Power Delay Profile (PDP) is obtained by combining elementary PDP coming from each gap-filler taking into account the path losses from the gap-fillers to the user terminal. This second approach allows the cell radius to be tuned and requires an elementary PDP repeated for each gap-filler. Two standardized PDP corresponding to a macro-cellular single emitter case have been used in contexts close to the DVB-SH one:
  - ITU Vehicular A recommended by ITU-Recommendation M.1225 [11]. This PDP has been used for UMTS performance assessment;
  - GSM-TU6 as defined in [18]. This PDP has also been used for DVB-H performance assessment;

- for path loss, the Xia-Bertoni model gives average results between dense urban and suburban cases of the COST 231-Hata model. Therefore this model is preferably used for such an analytical approach that requires a unique cell radius for the whole SFN configuration.

A SFN channel model for DVB-SH physical layer performance assessment is for further study.

4.2.4 Satellite specific issues

Another important element in the DVB-SH transmission chain to the user is the satellite payload that together with the land mobile satellite channel greatly contributes to the end-to-end system performance.

4.2.4.1 Satellite Payload architectures

Satellite payloads for mobile broadcasting systems comprise two major elements: a feeder-link receive section and a high power S-Band transmit section. The outstanding EIRP requirements obviously make the last element the most demanding and worth of a dedicated discussion.

The payload architecture of a high power transmit section largely depends on the required service performance (coverage, EIRP, number of beams, etc.) and on the satellite constellation typology (GEO/HEO). Systems aiming at wide coverage regions, are based on medium/large on-board antenna apertures which are composed of light-weight deployable rigid reflectors (these technologies are typically available up to about 5 meters of reflector diameter at S-band). On the contrary, regional or multi-regional coverage areas as seen from a GEO/HEO are of very limited angular extension and called for a large deployable reflector of typically 12 meters projected aperture diameter, at S-band. EIRP requirements and the difficulty of embarking large deployable antennas lead in most cases to a demanding RF power requirement that generally exceeds the generation capabilities of single amplifiers available for space applications and "power combining" must be adopted. The approach consists in adding up a plurality of amplified replicas of the desired signal individually amplified by highly-efficient/high-power devices such as Travelling Wave Tubes Amplifiers (TWTAs).

The problem dimensionality is further increased if power re-configurability is required to match varying and un-predictable market demand of transmitted capacity (i.e. EIRP) among the different beams, or to adjust in-orbit the beam coverages with the use of Beam Forming Network (BFN) or Ground Based Beam Forming (GBBF) antennas.

To meet these objectives, the distributed amplification concept plays a fundamental role. The general concept foresees several power amplifiers working together not only to achieve the total amount of power otherwise non-achievable with a single source, but also to constitute a single power pool to which each signal can get its own amount in a variable and reconfigurable manner. This leads to the question of whether the satellite can or cannot be used in non-linear mode as further discussed in the next clause. More details about typical satellite payload architectures are provided in clause A.1.

4.2.4.2 Nonlinear Payload Distortion Effects

For assessing payload distortion impacts on DVB-SH waveforms, the optimal operating point (OBO) of the TWTA must be evaluated. Details about the satellite nonlinearity impact and the TWTA operating point optimization methodology is for further study.

For this, two main satellite payload configurations should be distinguished (see clause A.1):

1) single multiplex amplification per TWTA (network and polarization combining);

2) multiple multiplexes amplification per TWTA (antenna spatial combing and multibeam satellite).

In the first case, the satellite TWTA can be driven close to saturation if the TDM waveform is selected i.e. SH-B. In this case, SH-B may have a power advantage of 1 dB to 1,5 dB with respect to SH-A.

In the second case, since each satellite TWTA is used to amplify several multiplexes, these amplifiers have to be backed-off away from saturation in order to avoid the creation of intermodulation products. Therefore the use of SH-B is not so advantageous in terms of link-budget for these class of payloads.

4.2.4.3 Phase Noise Aspects

In the DVB-SH direct path (i.e. satellite-to-user) system phase noise is dominated by the contributions from the satellite local oscillators used for on-board frequency conversion and the user terminal RF front-end Local Oscillators. This is because of the usually high frequency band used for the feeder link, the need of frequency conversion agility and the space environment. Table 4.7 reports a typical/worst-case Ka- to S-band satellite transponder DVB-SH phase noise mask which has been used to derive the impact of phase noise on DVB-SH waveform.

Table 4.7: S-band DVB-SH phase noise mask (Satellite)

	10 Hz	100 Hz	Hz	10 kHz	100 kHz	1 MHz	10 MHz
Phase noise (dBc/Hz)	-29	-59	-69	-74	-83	-95	-101

There exist simplified approaches to compute the impact of the phase noise residual phase error on the digital demodulator i.e. [i.7] for OFDM or [i.8] for TDM. In practice the phase noise that can not be tracked is dominated by the high frequency components of the phase noise i.e. the components from the equivalent phase estimator phase noise bandwidth B_N to the baud rate.

In SH-B two pilots groups having length $L = 80$ symbols and separated by $L_{TOT} / 2 = 1088$ symbols being L_{TOT} the PL slot length. Consequently the equivalent feedforward phase estimator can be computed as $B_N = \frac{R_s}{L_{TOT}}$ kHz which corresponds to $B_N =$ 1,8 KHz for the 4 Mbaud case. The DVB-SH phase noise mask of table 4.7 produces a typical amount of 4,5° rms untrackable phase noise due to the almost flat phase noise PSD between 1 KHz and 100 KHz.

The impact of phase noise error depends also on its dynamic compared to the physical layer FEC block size. Typically for the same standard deviation "fast" phase noise has much less impact compared to "slow" phase noise being the speed related to the FEC block duration. For this reason it is recommended to simulate the phase noise impact on the end-to-end system rather than just basing impact estimation on the simplified approach described above.

4.2.4.4 Satellite induced Doppler Shift

Satellite induced Doppler shift manifests itself as a change in signal frequency which cycles through a range of frequencies and which reaches a given ± maximum value. This cycling results in a smooth and readily predictable frequency shift as a function of time. This effect should not be confused with the Doppler spread induced by multipath and the receiver speed, *except in SFN mode*. In SFN mode, the satellite-induced Doppler must be corrected (locally) by the CGC, otherwise it manifests itself as a Doppler spread in the hybrid reception coverage. Clause 7.6.1 gives recommendations for this correction.

In considering Doppler shift here it is tacitly implied that any Doppler shift in the uplink can be pre-compensated for at the transmitting station. Unfortunately such pre-compensation approaches are not feasible for the downlink where the Doppler shift varies as a function of geographical location within the coverage footprint.

For GEO satellites the Doppler shift is very small and presents no problems for the receiver. For the case of a GEO system with an inclined orbit (up to around 5 degrees) the Doppler Shift may be of the order of ±250 Hz.

For HEO and MEO satellite constellations the downlink Doppler shift at S-band can be of the order of tens of kHz and requires that the receive can handle any initial frequency shift at acquisition and can track the frequency shift during the orbit pass. Figure 4.4 shows typical ground tracks of HEO systems targeted for European service based upon the parameters given in table 4.8.

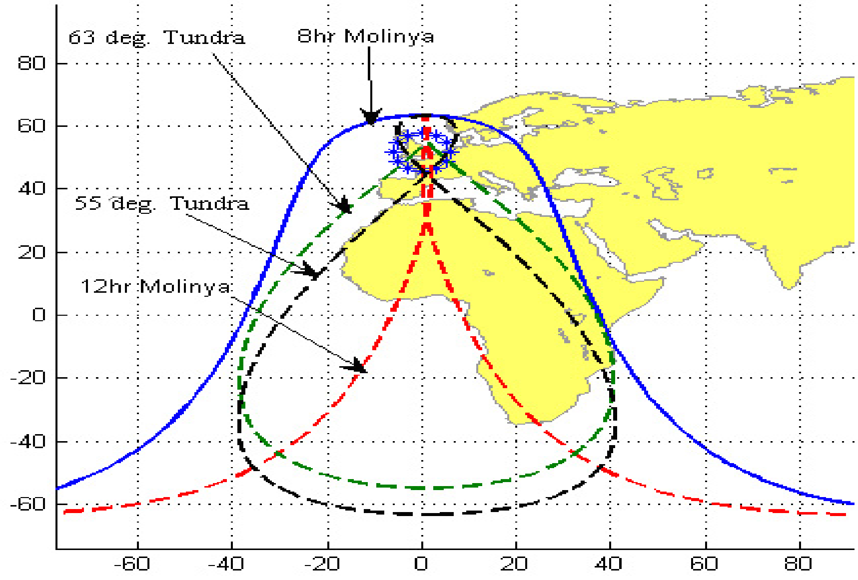


Figure 4.4: Ground track of typical HEO orbits

Table 4.8: Representative parameters of typical HEO Orbits

Orbit	Orbit Period (hrs)	Semi-major Axis (km)	Apogee Radius (km)	Apogee height (km)	Perigee Radius (km)	Perigee height (km)	eccentricity	Inclination (degs)	Argument of Perigee (degs)	Range of Elevation Angles (degs)	Mininum number of satellites
8 hr Molnya	8	2 027	33 163	26 785	7 378	1 000	0,636	63,435	270	55 to 90	8
12 hr Molnya	12	26 561	45 804	39 426	7 307	929	0,7249	63,4	270	60 to 90	4
63 degs Tundra	12	42 164	53 480	47 102	30 847	24 469	0,2684	63,4	270	at least 60	3
55 degs Tundra	24	42 164	55 656	49 278	28 672	22 294	0,32	55	270	65 to 90	3

The maximum value of S-band downlink Doppler shift and the maximum rate of change of this Doppler shift with time are given in table 4.9.

Table 4.9: Typical maximum Doppler shift parameters for downlink at S-Band

Orbit	Min elevation angle (degs)	Maximum Doppler shift (KHz)	Maximum rate of change of Doppler shift, (Hz/s)
8 hr Molnya	50	17	2,0
	60	15	2,0
12 hr Molnya	50	22	1,8
	60	21	1,8
63 degs Tundra (24 hr)	50	3,9	0,35
	60	3,9	0,3
55 degs Tundra (24 hr)	50	4,8	0,3
	60	4,3	0,3
MEO	10	11,0	4,5
GEO, stable orbit	Any	Very small	Very small
GEO, inclined orbit (5^0)	Any	0,25	0,02

4.2.5 DVB-SH Impairments Countermeasures

4.2.5.1 LMS Mobile Channel Impact Mitigation

To mitigate the impact of LMS propagation impairments, two specific techniques are recommended: exploit diversity and optimize demodulator robustness.

4.2.5.1.1 Exploitation of Time and Space Diversity

The DVB-SH standard provides inherent methods for both time and space diversity:

Method 1: Time diversity

It is supported in DVB-SH by the highly scalable Physical time interleaver. In addition, a solution based on LL-FEC, with possible splitting of the interleaving function between the physical and the link layers is further detailed in clauses 6 and 7.

Method 2: Space diversity (simultaneous reception of SC and CGC)

For the SH-A (SFN), the space diversity is intrinsic in the demodulation process as the isofrequency satellite and terrestrial OFDM signals will be combined at the user terminal antenna.

For the SH-B and SH-A (MFN) cases, the user terminal needs to separately demodulate the satellite and terrestrial signals and may implement different combining options depending on its capability (clause 7.6).

In addition, the DVB-SH standard supports space and time diversity from "extrinsic" methods:

Method 3: Multiple satellite space diversity

The same combining strategy for MFN of Method 2 applies to this case too. For SFN combining there are restrictions.

Method 4: Multiple receive antenna space diversity

It has particular benefits in terrestrial conditions, mainly indoor (quasi-static). For the satellite direct reception antenna diversity may bring valuable performance improvement for (very) slow mobility conditions if the separation is sufficiently large (e.g. 1 meter or more) which is often difficult to support in practice, except for vehicular applications. Polarization diversity (e.g. horizontal/vertical linear or left-handed/right-handed circular) may be beneficial as soon as the reflected components still carry sufficient power.

4.2.5.1.2 Robust Demodulator Operations

In a mobile fading channel, it is important to have very robust signal re-acquisition after signal loss as well as a fast and reliable signal state tracking. Although this aspect is also common to terrestrial standards such as DVB-H some specific aspects must be addressed by DVB-SH:

- satellite operations with limited link margins compared to terrestrial systems;
- demodulator operation at lower SNR than terrestrial systems.

Guidelines for the demodulator implementations and performance for this aspect are given in clause 10.

4.2.5.2 Nonlinear Channel Impairments Countermeasures

In case of multiple-carriers amplification, the use of linearized TWTA can bring appreciable advantages. Similar performance improvement can be obtained by on-ground pre-distortion to compensate for the satellite TWTA nonlinear characteristic.

In case of single-carrier amplification, the use of linearized TWTAs provides limited performance improvement in particular for QPSK with low code rate as the TWTA will be operated in the vicinity of its saturation (where the linearizer has little effect). However, linearized TWTA may provide appreciable benefits for 16APSK in particular if the operating point corresponding to a given modulated signal drive is not negatively impacted by the linearizer presence.

For single carrier TWTA operation, 16-APSK being multi-ring can benefit from simple static pre-distortion techniques which are able to mitigate the constellation warping effect [i.16]. Dynamic pre-distortion techniques are particularly suitable to reduce the clustering effects of higher order modulations such as 16APSK. Their benefit for QPSK and 8PSK is instead limited.

4.2.6 Receiver characteristics

This clause discusses, in general terms, the architectures of DVB-SH receivers. The detailed Reference Designs for the terminal classes and their recommended performance are given in clause 10.

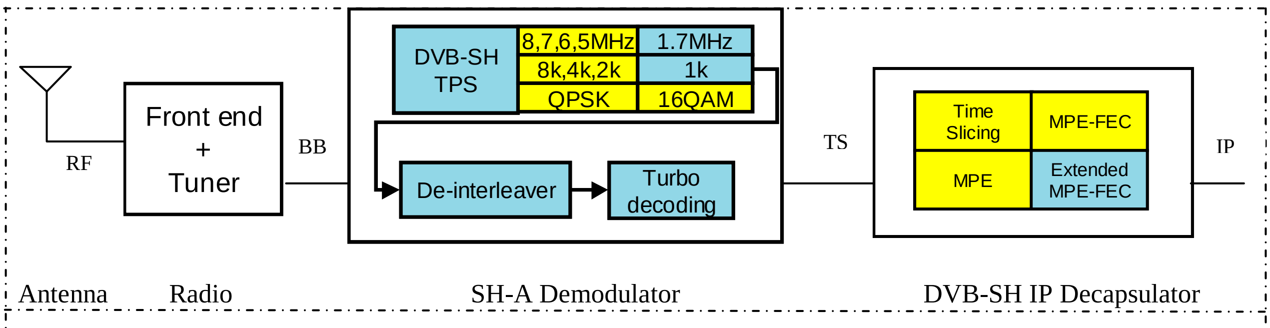
DVB-SH receivers must be able to:

- receive seamlessly a signal when moving between satellite(s) and CGC(s) coverages;
- implement battery saving schemes when necessary;
- exploit diversity mode with several receiver branches in parallel when appropriate;
- co-exist with other active radio functions (2G, 3G, IEEE 802.11 [26], Bluetooth, etc.) embedded in a terminal; and
- be relatively immune to high-level signals from various cellular network base stations (2G, 3G) or/and ISM systems.

Two different receiver architectures can be distinguished according to the DVB-SH waveform options, respectively the OFDM/OFDM (SH-A see figure 4.5) and the TDM/OFDM (SH-B see figure 4.6) system architectures.

The SH-B architecture encompasses the SH-A architecture in the sense that SH-B receivers can be used in an SH-A configuration. The converse is obviously not true.

Receivers designed for the SH-A architecture are in general intended to be used with SFN between satellite and CGC. However, it is recommended that they should be also compatible with an MFN configuration (transmissions of the OFDM satellite signal and its OFDM CGC counterpart on two different sub-bands).



NOTE: The dual tuner may have a common front-end and downconversion.

Figure 4.5: SH-A Receiver architecture

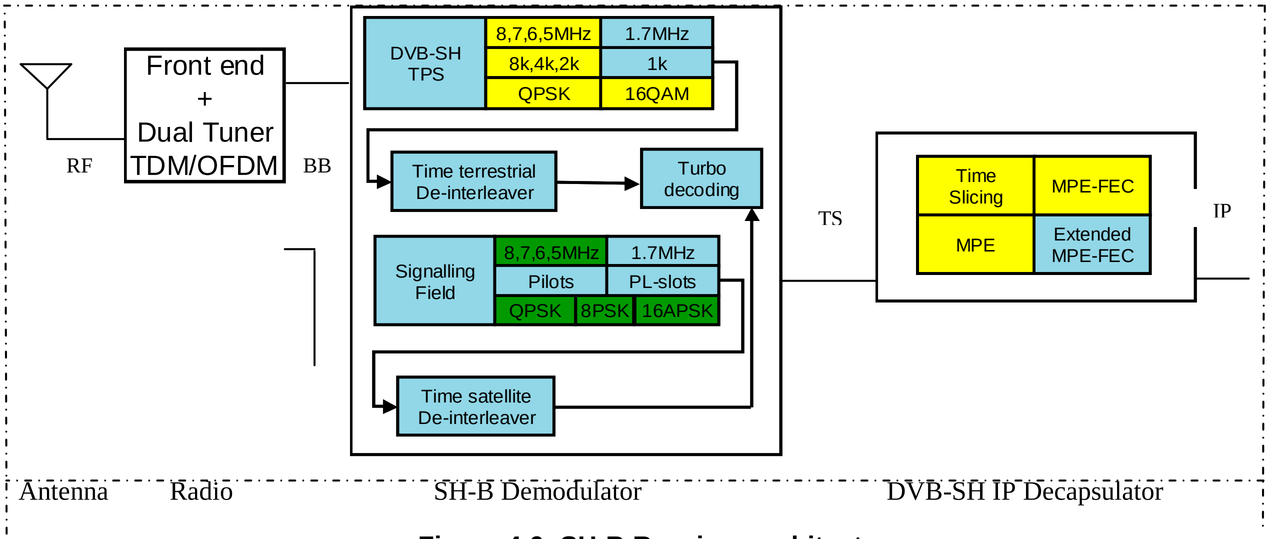


Figure 4.6: SH-B Receiver architecture

4.2.6.1 Vehicular reception constraints

Vehicular receivers must work at high moving speeds and stay longer in a satellite-only reception mode. However, vehicular receivers can exploit the following features not generally possible for receivers used in a handset:

- the terminal has adequate power supply able to support more complex receiver processing;
- the terminal allows better antenna diversity (order 2 or more) implementation (form factor and antenna spacing);
- one antenna can be optimized for satellite reception (directivity and matching polarization);
- Low Noise Amplifiers (LNAs) can be integrated with the antenna(s) to reduce sensitivity loss (the underlying assumption is that the vehicular receiver does not coexist with embedded Telecom modems in the same electronic equipment and therefore the RF filtering protection for the LNA is removed or relaxed, therefore improving noise factor);
- larger memory can be embedded in the receiver so that longer Physical Layer interleaving can be supported;
- the higher speed of the terminal allows a better exploitation of time diversity (either at Physical Layer or at Upper Layers), at equal memory resource.

4.2.6.2 Handheld reception constraints (pedestrian)

In most common cases the channel conditions are that produced by a pedestrian user (< 3 kmph). Due to the relatively low speed, continuity of service is in general achieved by increasing the link margin, rather than by increasing the time interleaving depth. When in satellite-only reception mode, some cooperation may be required from the user, i.e. to maintain good LOS with the satellite.

Some challenges are associated to the specificities of handheld terminals. These include:

- antenna diversity (order more than 2 is very challenging);
- small battery requires an efficient power saving management;
- antenna gain is in general low (can be less than -3 dBi);
- antenna polarization is most often linear and not optimized to satellite reception;
- embedding telecom modems like GSM or 3G inside terminal without reducing the satellite receiver sensitivity;
- RF filtering, antenna design rules and compactness constraints have an impact on the achievable receiver sensitivity and immunity to high level blockers coming from the terminal;

- memory limitation may, in some architectures, not allow the support of a large Physical Layer interleaver.

4.2.7 Terminal High-level architecture

In theory, all types of terminal can embed large memory and all options defined above. Furthermore, no limitation is imposed on the number of embedded multi-band or/and multimode 2G, 3G communication modem(s) or other wireless features like Bluetooth or WiFi. Also GPS could be integrated in a terminal. In practice, compactness, supply source and price considerations split terminal in categories. Categories chosen in DVB-H Implementation Guidelines [i.21] and in EICTA/TAC/MBRAI-02-16 [14] are kept (see clause 10). Figure 4.7 shows the terminal high-level architecture for all classes of terminals. Terminal sub-systems will be designed according to their terminal categories and manufacturer marketing requirements.

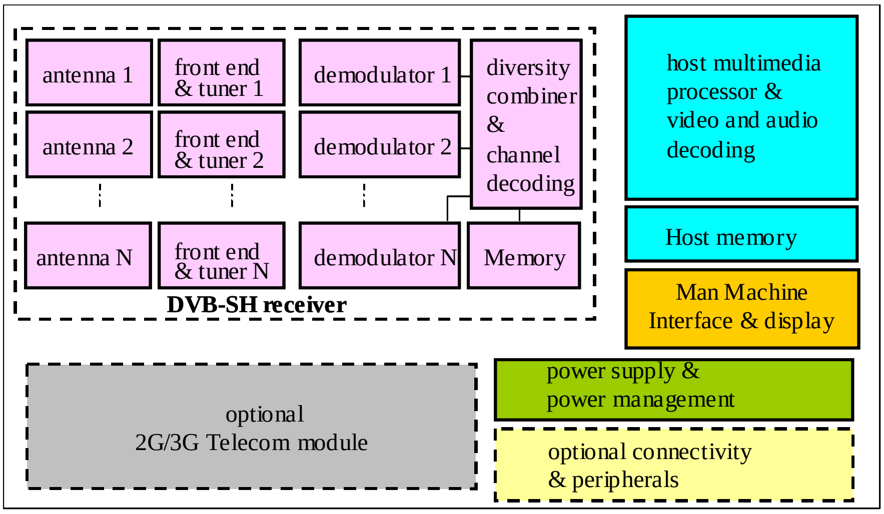


Figure 4.7: Terminal high-level architecture

4.2.8 Compatibility with Existing DVB-H System Features

4.2.8.1 Battery-powered receivers

This clause deals with receivers embedded in handheld "convergence" terminals and pocketable digital TV receiver devices (see clause 10 for definitions of receiver categories). For these devices, battery saving is essential.

4.2.8.1.1 Time slicing

The DVB-H group has defined time slicing as a key mechanism for battery saving purposes. DVB-SH receivers also use this mechanism for the same purposes. Burst size and off time, maximum burst duration, power saving calculations are based on same assumptions than DVB-H.

4.2.8.1.2 Low-power Microelectronics

State-of-the-art technologies like the newest RF BiCMOS or SiGe for tuner and CMOS 65 nm process for channel decoder and sometimes full CMOS process for one chip solutions have significantly reduced the contribution of the front-end in the overall terminal power consumption, compared to what it used to be some years ago.

4.2.8.1.3 Antenna Diversity (for S-band)

Operation at S-band enables the use of antenna diversity, even in small handheld terminals.

Antenna diversity does increase power requirements since more than one receiving chains operate in parallel. However, this power consumption increase can be kept acceptable compared to the gain in QoS improvement with the use of new low-power technology and time-slicing mechanism as stated above. In addition, a signal quality indicator could be used to turn off or reduced diversity processing when this is not required.

4.2.8.2 Service Switching time (Zapping time)

Zapping time is an important subjective quality criterion for the user in a multi-program offer. Due to the combined effect of time-slicing for battery-life conservation and time-diversity techniques to overcome long fadings, possible degradation of zapping time may be incurred if appropriate counter-measure techniques are not applied. In DVB-SH it should be noted first that zapping time may be different depending on the coverage (CGC or Satellite) and the particular length and structure of the time interleaver chosen for the propagation channel in each coverage.

In the CGC coverage, DVB-SH zapping time, normally, does not exceed 2 s on average.

In the satellite coverage, one can distinguish between good reception conditions (e.g. about > 5 dB above threshold) and critical reception conditions. A "smart" terminal can optimize zapping time according to its measured reception margin. In good reception conditions, such terminals can immediately play out any source data without waiting for maximum FEC protection to be fulfilled. This strategy is made possible by special sending arrangements of FEC parity and source data allowed by DVB-SH. Techniques for transitioning between "immediate" play out strategy (in good reception) and maximum error protection (in mobility) may be needed to be implemented in the terminal. Clauses 6 and 7 describe the use of these techniques, for the link and the physical parameters respectively, and also discuss their possible side implications.

4.2.8.3 Support for Variable Bit Rate (VBR) and Statistical Multiplexing (Statmux)

The use of VBR and statistical multiplexing (Statmux) of the different services transmitted in a single TS allow either bandwidth saving or better video quality. The actual gain provided by VBR depends on the system configuration, on the types of contents and on the number of services multiplexed statistically on the TS.

The main element impacting VBR implementation with DVB-SH is the combination of the Time interleaver together with Time Slicing:

- with short interleaving (i.e. slightly exceeding the length of the Burst), VBR is supported with time slicing thanks to the "delta-T" mechanism defined for the same purpose in the DVB-H standard;
- when long interleaving is used, Burst content (and hence "delta-T" information) is spread over several Bursts. This adds complexity in handling of Time Slicing, but does not preclude the use of VBR nor reduce its bandwidth saving gain.

VBR implementation is described in more detail in clause 6.3.4.

4.2.8.4 Multiple QoS support

4.2.8.4.1 Introduction

This clause deals with the support of variable quality of service in a DVB-SH environment. The overall quality of service results from the combination of different layers. Streaming protection is mainly achieved by physical and link layers, whereas physical and application layers achieve file delivery protection.

Depending on the chosen systems configurations, different options are possible at different layers as presented in table 4.10.

Table 4.10: Configuration options for the different classes

Terminal option		Parameters	Relevance
Class 1	Physical layer	Rate	++
	Link layer	Rate; depth	++++
	Application layer	Rate, repetition	++
Class 2	Physical layer	Code rate; interleaver	++++
	Link layer	Code rate; interleaver	+
	Application layer	Rate, repetition	++

4.2.8.4.2 At physical layer

DVB-SH does not provide variable quality of service at physical layer. The following parameters must be configured according to the expected performance of the most demanding service and as a function of the worst channel model scenario:

- code rate;
- interleaver configuration, in particular interleaving depth.

4.2.8.4.3 At link layer

DVB-SH does provide at link layer a per-service protection based on MPE sections. It is possible to configure a streaming service with a long interleaving while a "push" service may be configured with no protection at the link layer since the protection is provided at the application level.

4.2.8.4.4 At application layer

DVB-SH does not provide any specificity for the application layer since the interface with the "application world" is IP. However, the DVB-SH application layer is fully compliant with all CBMS specifications and, in particular, with the file delivery specified in document [Content Delivery Protocols]. It is also expected that other techniques may be applied to the streaming services such as scalable video encoding. Scalable video encoding must be coordinated with the link layer protection since the former is based on a differential protection applied to two parallel streams making the video stream.

4.2.8.4.5 Recommendations

The procedure displayed in figure 4.8 should be followed when configuring the parameters of the FEC protection.

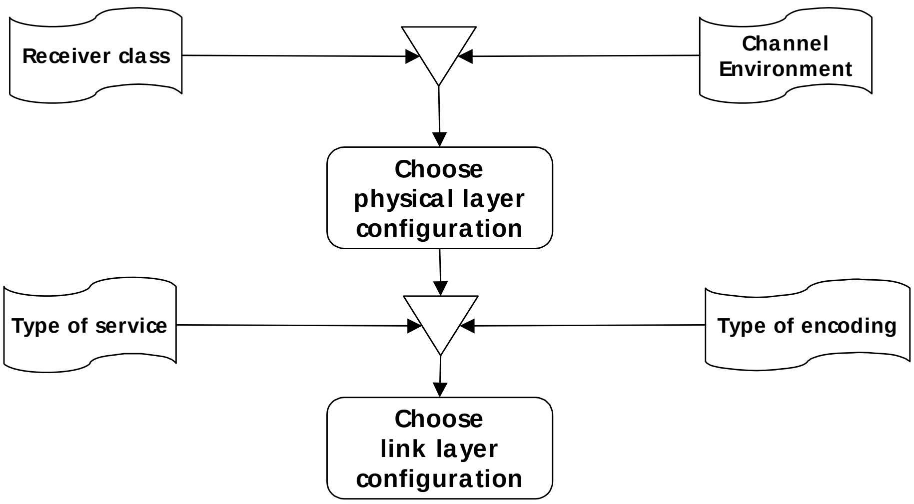


Figure 4.8: Procedure for choosing FEC protection parameters

Several configurations are possible while offering the same useful throughput but not always the same performance:

- protection at physical layer only compared to joint physical and link layer protections;
- different splits of the joint protection.

The network operator, depending on the deployed network and receivers, will choose the physical layer configuration. The service operators will then be able to choose if they want to activate the link layer and application layer protections, trading performance versus available bandwidth.

Differentiation between services at application and link layer may usually happen in the following circumstances:

- when different kinds of services are broadcasted in parallel: this is for instance the case between the streaming application (which will highly benefit from the link layer protection) and the file download application (which may benefit better from the content delivery protection and then will less use the link protection);
- when, for a specific service, different levels of protection are required: this is the case of scalable video encoding which requires two levels (or more) of protection to protect better the main stream than the others.

4.2.8.5 Service announcements in hybrid networks

This clause deals with announcement of services within the DVB-SH context.

4.2.8.5.1 Introduction

Signalling within the DVB-SH system follows the layered approach described in figure 4.9.

Figure 4.9: Layered signalling approach

According to this layered approach, a terminal learns all information about the network and services at first switch-on time in the following steps:

- radio configuration: the terminal can discover the physical layer parameters by scanning relevant frequencies and recovering either TPS (in OFDM waveform) or Signalling Field (in TDM waveform). These fields will deliver all information about the baseband configuration like FEC ratio, interleaver depth, etc.;
- TS configuration: the terminal is then able to decode the MPEG2 flow. If required it will discover the SH services structure by reading the SHIP packets;
- section configuration: the terminal will read the PSI/SI tables that will give all system related information to be able to derive from a target IP address the relevant PID: Given a target IP address, the IP Notification Table (INT) will deliver the relevant component tag of the target elementary stream;
- then using the PMT and PAT, the relevant PID will be found;
- programme configuration: the terminal is then able to read to the bootstrap ESG IP address to discover the list of ESG providers, choose one and then listen to the relevant IP address to discover its electronic service guide. The ESG has the format of an XML file describing what content can be watched when, typically <name_of_the_programme><IP_address>. Then, using PSI/SI the relevant PID is fetched by the terminal.

Compared to DVB-H, the only difference is the introduction of the SH service layer. The service announcement is performed, as in DVB-H, according to the Electronic Service Guide specification of the DVB-CBMS protocol suite. In general, there is nothing specific compared to the ESG used for the DVB-H, in particular for the SFN case that is very similar to the DVB-H case (no need for SH services). In the MFN case, the announcement of the local content and of its availability at particular locations will require the use of the "SH service" concept (see clause 6).

4.2.8.6 Handover issues

This clause describes in general terms the handover issues in DVB-SH. DVB-SH has been designed to take advantage of the mechanisms for handover that have been defined in the DVB-H framework (see TS 102 470 [21]). In particular, the following principles are maintained in DVB-SH:

- **use of physical layer signalling:** for instance, the TPS provides information such as cell_id to quicken the detection of the target network. A similar signalling mechanism has been defined for the DVB-SH TDM waveform;
- **use of spare time in time-slicing for scanning available frequencies:** this is also possible in DVB-SH, even with long physical layer interleaving;

- **PSI/SI signalling:** the same philosophy of signalling has been maintained, with some adaptations to cater for the hybrid architecture. The NIT describes the full network in a beam, including the target frequencies and waveforms. The INT describes the availability of the session on the target transport stream.

Nevertheless, DVB-SH introduces also some specificities that are worth keeping in mind:

- presence of the satellite "umbrella" cell that covers all the terrestrial cells;
- terrestrial cells can be much smaller than in DVB-H, especially in S-band (less than a Km in radius);
- terrestrial cells and satellite umbrella cell may use different time interleaver lengths for the same service (for SH-B and MFN-SH-A;
- terminals with diversity reception have more than one front-end and demodulator;
- coherency of PSI/SI signalling must be addressed carefully when code diversity combining (with SH-B only) is used.

A detailed analysis of handover in DVB-SH is given in reference [i.9].

4.2.9 Support of Other Frequency Bands

The DVB-SH applicability should not be limited to above mentioned bands. The standard should be scalable to other frequency bands and channelization schemes that could be allocated to satellite broadcasting services.

Possible other frequency bands to be supported by DVB-SH are the following:

1) Ku band: non planned FSS, planned FSS, planned BSS: 10,7 GHz to 12,75 GHz;
2) Ka band: non planned: 17,3 GHz to 20,2 GHz. BSS: 21,4 GHz to 22 GHz;
3) other bands: e.g. C-band.

A high-level survey about the standard applicability to the above listed frequency bands above 3 GHz led to the following preliminary conclusions:

- the SH-A configuration is definitely not suitable for A3GH applications as the OFDM waveform will not be able to cope with the higher Doppler spread of A3GH systems (in particular Ku and Ka-band) unless a drastic reduction in mobile user speed can be accepted;
- the satellite TDM component of the SH-B configuration can also be exploited for A3GH bands with no expected impact on performance. In fact the mobile channel countermeasures adopted for SH-B are also valid for A3GH operations. The possible Ku/Ka-band receiver different phase noise mask compared to B3GH applications must be compatible with SH-B TDM pilots structure;
- for the CGC other solutions must be used (different waveform compared to the SH-B) or different frequency bands (e.g. B3GH whereby SH-B OFDM waveform can be exploited).

5 System Configurations and possible Deployment

5.1 Definitions and concepts

System: combination of physical elements (satellite, CGC, terminals, etc.) implementing the DVB-SH specifications.

Radio configuration: defines the selected parameters of the DVB-SH waveform. Among available options we find: SH-A (OFDM/OFDM) and SH-B (OFDM/TDM), interleaving (physical, link), FEC ratios, modulation mode, hierarchical modulation.

Frequency configuration: spectrum allocation to the satellite and terrestrial component. For the terrestrial transmitters, it is useful to distinguish between hybrid and non-hybrid frequency sub-bands. A non-hybrid frequency sub-band is defined as one that carries only local content. One main option in the frequency configuration is the choice between SFN and non SFN. In a SFN configuration, the hybrid sub-bands carry only the Common content. In a non SFN configuration, hybrid sub-bands include also those that carry a mix of Common and Local content.

Topological configuration: used in conjunction with frequency configuration, it defines where the spectrum is available, on a satellite ("beam") or terrestrial ("cell") basis. For satellite, it gives the location and shape of the beams. For a terrestrial cell, it gives the centre (in latitude, longitude) and extension of the cell.

Local Content Insertion configuration: defines the way the local content is inserted by the CGC transmitters, in the context of a non-SFN network.

CGC Transmitter configuration: hardware options placed on the transmission site; it concerns the location (co-sited with a 3G base station, in a new specific site), the antenna (reused on new), the power.

Receiver configuration: hardware options placed on the receiver; it concerns essentially the available memory for physical and link layer interleaving (using class 1 and class 2 categories) and the antenna options (diversity, gain, etc.).

System configuration: set of options retained for a DVB-SH system; this includes *radio, frequency, topological, content, receiver and transmitter configurations.*

System transition: evolution of a system when one of its configurations changes. Not all transitions are relevant and/or foreseen in the system deployment.

Compatibility: a receiver is *compatible* with "radio configuration X" when it can successfully decode the useful information using only a sub-set of the allowed waveform elements in this configuration. A lower reception margin or a more restricted coverage area is implied compared to that of a compliant receiver.

Compliancy: a receiver is *compliant* with "radio configuration X" when it can exploit ALL the waveform elements allowed in configuration X to maximize reception margin and/or coverage area.

5.2 Configurations options definitions

This clause discusses in more details on the options that are offered.

5.2.1 Radio configuration

The air interface has many parameters available. Three of them are discussed in this clause.

Choice of TDM (SH-B) or OFDM (SH-A) for the satellite signal: this choice is dictated mainly by spectral efficiencies considerations (satellite, terrestrial, overall) and the satellite nonlinearity characteristics as discussed in clauses 4, 4.1.3, 4.2.3.2 and C.1.

Choice of between physical layer or link layer techniques to combat long fadings: comparison of these techniques is given in clauses 6 and 7. The choice is dictated by the performance required in a long fading environment; however, the cost and required foot-print of the memory to implement a long interleaver at the physical layer must be considered. In the short-term, the combination of a short physical interleaver with a long link layer interleaver is preferable, especially for handheld terminals (see clause 10). Note that, in code diversity, since only the FEC packets must be correlated, it is possible to resort to different interleavers (physical and link) for the satellite and terrestrial signals.

Choice of Modulation and FEC: guidelines for this choice are given in clause 7.

5.2.2 Frequency configuration

It is important to plan how the frequency can be used and, more importantly, reused.

- firstly, the bandwidth of a frequency sub-band must be decided between the values allowed in the standard (1,7 MHz, 5 MHz, 6 MHz, 7 MHz and 8 MHz). This bandwidth is most of the time implied by the frequency band selection;

- secondly, the choice between SFN and non-SFN must be made, on a beam-by-beam basis. In SFN case, a minimum of 1 sub-band is needed per beam, whereas in the non-SFN case this minimum is 2 sub-bands. Since the sub-band that repeats the Common content in the non-SFN CGC may need to be at different frequencies in different places, this minimum may be higher than 2;

- thirdly, the remaining sub-bands can be allocated to local content transmission, on a per cell basis.

5.2.3 Topological configuration

This consists in allocating the sub-bands to the topological elements, i.e. the satellite beam(s) and the terrestrial cell(s). Adjacent beams must not have common frequencies for interfering avoidance reasons. This is the same for adjacent cells. However it is possible to reuse these frequencies:

- a frequency used in a beam/cell can be reused in another beam/cell if this beam/cell is sufficiently separated from the first;

- terrestrial cells can reuse an adjacent beam frequency provided they these cells are sufficiently far away from this beam.

Actual frequency allocation to sub-bands may be quite complex due to possible interferences from the satellite adjacent spot beams. In particular, for reusing adjacent spot beam frequency, terrestrial reuse is sought in the centre of the spot (far from the borders) but this is not always possible.

5.2.4 Local content insertion configuration

The choice of the method for Local content insertion depends on the ratio between the Local content and Common content bit rates.

- If this ratio is greater than 2, the *hierarchical modulation* method can be used: content is split into 2 TS. The First TS is input to the primary interface of the terrestrial modulator; this TS is exactly the same as the one going to the satellite transmitter. The Second TS is input to the secondary interface of the terrestrial modulator to carry local contents.

- Otherwise, the *content removal* method can be used: a single TS is generated and transmitted to all transmitters, either satellite or terrestrial. Using the SHIP synchronization, the transmitters will forward the only relevant part of the TS. The satellite transmitter removes all the local content. The terrestrial transmitters remove the part of the local content they need not forward.

NOTE 1: Hierarchical modulation can be used also for providing graceful degradation (see note 2). For example, it can be used to transmit, on a low-protected TS, additional data that can be exploited to improve content quality, either by sending enhanced video coding layer (as in scalable video coding), or by sending additional MPE-IFEC protection using a second source (see clause 6). The benefit on the signal quality will depend mainly on the position of the terminal under the satellite-only coverage:

  - under good reception quality, the terminal will benefit from both TS signals, the low priority that conveys usual video and IFEC protection, and the high priority signal that conveys enhanced layer and/or additional MPE-IFEC source;

  - under worse reception quality, the terminal will benefit only from the high priority giving the basic video layer and primary source of MPE-IFEC protection.

NOTE 2: The detailed specifications for this configuration is for further study.

5.2.5 CGC transmitters configuration

There are two main approaches for configuring the CGC network of transmitters:

- the "high-density", "low-power" approach attempts to reuse, all or partially, existing 3G/2G transmitter sites, or to build an equivalent low to medium height type of transmitters network. These networks are characterized by transmitter towers of typically 30 meter high delivering from 200 W to 1 kW ERP for dense urban coverage in the range of 0,5 km (deep indoor) to 2 km (outdoor).

- the "low-density", "high-power" approach attempts to reuse existing digital terrestrial TV transmitter sites, or to build an equivalent high-altitude transmitters networks. These networks are characterized by transmitter towers of 100 to 300-meter high, delivering 1 kW to 4 kW ERP for typical coverage in the range from 5 Km to 7 Km.

5.2.6 Receiver configuration

The standard identifies two receiver classes based on their capability to process **time diversity**. Refer to clause 10 for the memory sizing guidelines for each Class. Class 2 receivers offer maximum flexibility in terms of time diversity. Class 1 receivers are compatible with the time diversity transmitted for Class 2. Refer to clauses 6 and 7 for the different trade-offs between memory complexity and time-diversity performance.

5.2.7 Configuration naming

Of the various possible configurations described above, 4 are considered critical in system deployments: Waveform, Receiver Class, SFN/MFN, Local content handling.

We propose the following nomenclature for a configuration.

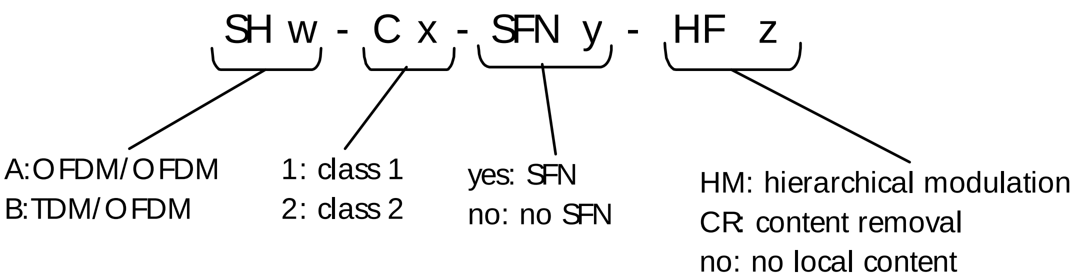


Figure 5.1: DVB-SH system configuration classification

The rules for referencing the configurations are the following:

- when the option is not set, the character is set to this value w, x, y, z; for instance if we want to refer to SH-A configurations in SFN, whatever the other parameters may be (classes and local hybrid frequency), we refer to "SH-A-C-x-SFN-yes-HF-z";
- wildcards are always 1 character long and lower cap; example of wildcards are "w", "x", "y", "z".d;
- exact values are either 1 character long in higher cap or several characters long; example are "A", "B", "1", "2", "hm", "cr".

5.3 Examples of system deployment

Below is a non-exhaustive list of relevant transitions:

1) from **SH-w-C-1-SFN-y-HF-z** to **SH-w-C-2-SFN-y-HF-z:** this transition corresponds to a terminal upgrading in terms of memory, in particular by replacing the long upper layer with physical layer long interleaving;

2) from **SH-w-C-x-SFN-no-HF-z** to **SH-A-C-x-SFN-yes-HF-no:** this transition corresponds to a bandwidth upgrading by moving from non SFN conditions to SFN conditions;

3) from **SH-A-C-x-SFN-yes-HF-no** to **SH-w-C-x-SFN-no-HF-z:** it would make non sense downgrading overall bandwidth by changing from SFN to non-SFN unless terrestrial network is dense enough to enable important local content insertion that mitigates bandwidth loss. So this transition corresponds to an increase of local content by a strengthening of the CGC network.

These transitions will be used in the following descriptions of system scenarios.

5.3.1 SH-A deployment scenario

Below is one typical SH-A deployment.

Step 1: early deployment of CGC

- SH-A-C-1-SFN-yes-HF-no configuration enables to deploy receivers with the constraint of limited memory, both on physical and link layers. Antenna diversity and physical short time diversity offer quite good performance under CGC coverage. Frequency reuse is possible between all CGC cells.

Step 2: launch of a satellite with limited power: we have 2 basic options

- no transition:
  - LL-FEC is used for time diversity;
  - depending on the link layer memory deployed in Step1, early terminals may remain compliant;
- transition to a new configuration:
  - change to a physical layer protection: transition to SH-A-C-2-SFN-yes-HF-no:
    - early terminals are only "compatible" (inside CGC coverage, in good satellite only reception conditions);
  - change to a non SFN configuration: go to SH-A-C-x-SFN-no-HF-cr:
    - early terminals continue to receive as before under the CGC and correctly receive satellite only under good reception conditions;
  - a transition to SH-B leads to the replacement of early terminals. The viability of this transition is therefore dependent on the latter installed base;
  - in the case of NGSO satellite, SFN-yes may no longer be possible.

Step 3: launch a second generation terrestrial network we have 2 basic options

- no transition;
- increase local content insertion by increasing ratio between satellite and terrestrial bandwidth on the hybrid frequency: go to SH-A-C-x-SFN-no-HF-hm.

Step 4: launch of a satellite with extended power

- different Modulation and FEC parameters may be fine tuned to take advantage of the increase satellite margin. Note that SFN operation may limit the freedom to tune these parameters.

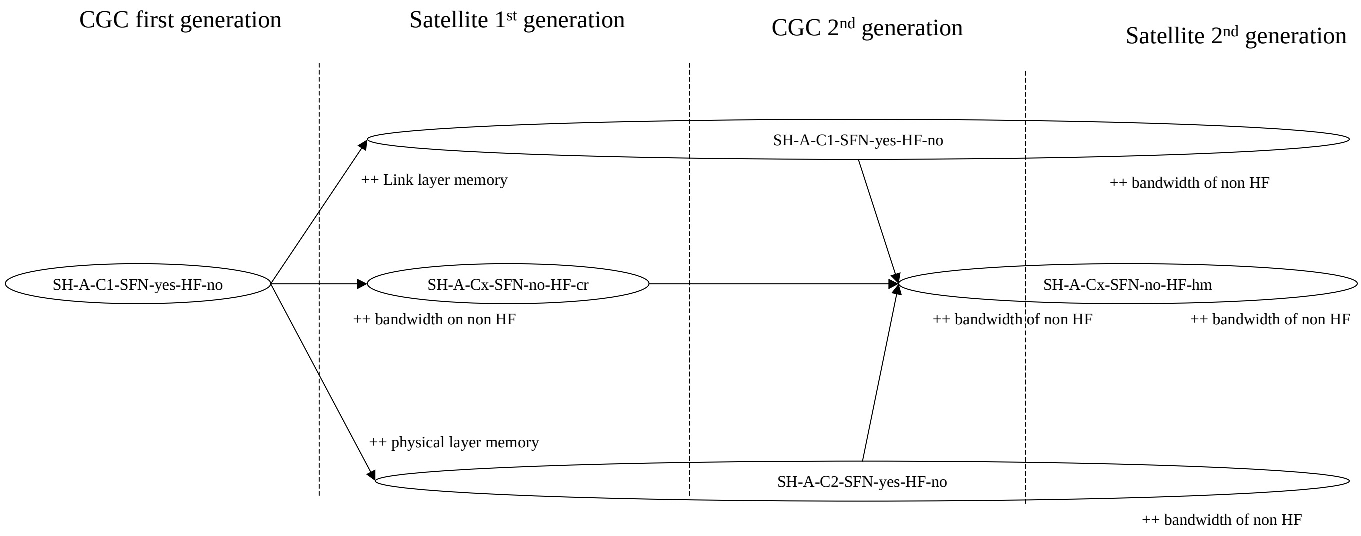


Figure 5.2: Typical SH-A deployment sequences

5.3.2 SH-B deployment scenario

In a typical SH-B deployment, we have the following steps:

Step 1: early deployment of CGC

- SH-B-C-1-SFN-no-HF-z configuration enables to deploy receivers having limited memory constraints, both on physical and link layers. Physical short time diversity offer quite good performance under CGC coverage.

Step 2: launch of a satellite with limited power capacity; 3 possible configurations are possible

- no transition:
  - LL-FEC is used for time diversity;
  - depending on the link layer memory deployed in Step1, early terminals may remain compliant;
- transitions to a new configuration with better time diversity: transition to SH-B-C-2-SFN-no-HF-z:
  - early terminals are only "compatible" (inside CGC coverage, in good satellite only reception conditions);
- transitions to a new configuration with better local content insertion: transition to SH-B-C-x-SFN-no-HF-cr:
  - early terminals are still compliant; more local content is available.

Step 3: launch a second generation terrestrial network we have 2 basic options

- no transition;
- increase local content insertion by increasing ratio between satellite and terrestrial bandwidth on the hybrid frequency and go to SH-B-C-x-SFN-no-HF-cr.

Step 4: launch of a satellite with extended power capacity

- different Modulation and FEC parameters can be easily fine tuned to take advantage of the increase satellite margin;
- depending on the relative satellite power, it could also be envisaged to transition to SH-A if local content insertion is not sufficient and satellite power is large enough.

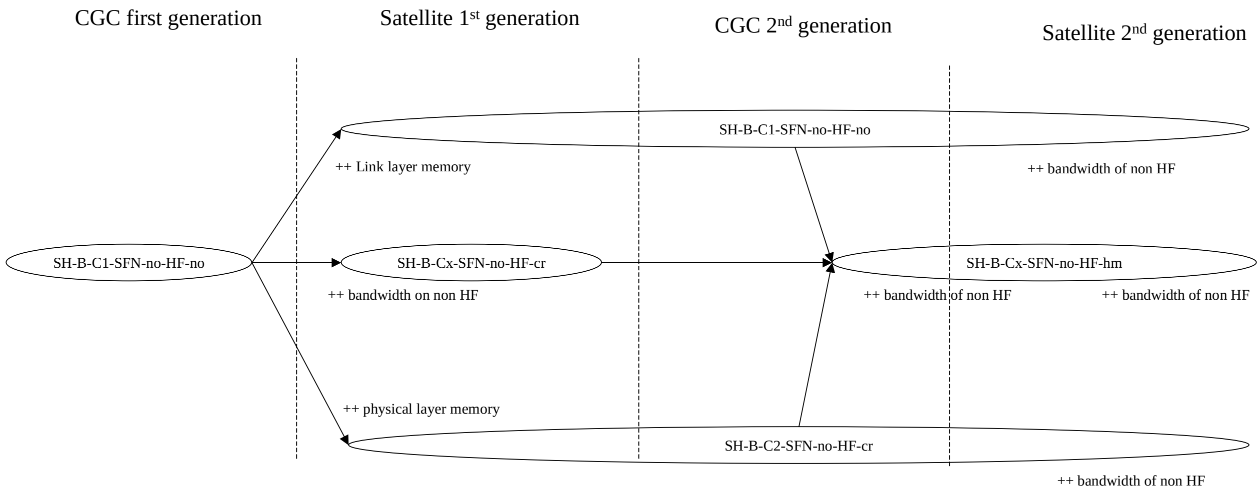


Figure 5.3: Typical SH-B deployment sequences

5.3.3 Less likely transitions

These are listed hereafter:

1) from **SH-w-C-2-SFN-y-HF-z** to **SH-w-C-1-SFN-y-HF-z:** this transition is relevant only when the network/market/use cases/receiver technology removes the need of long physical interleaver introduced in the earlier stage;

2) from **SH-B-C-x-SFN-no-HF-z** to **SH-A-C-x-SFN-no-HF-z:** the underlying logic is that a transition from SH-B to SH-A is accompanied by a transition from non-SFN to SFN. Note however that a transition from non-SFN to SFN is not always easy if the early CGC network takes advantage of the non-SFN characteristic and is not configured uniformly;

3) from **SH-w-C-x-SFN-no-HF-hm** to **SH-w-C-x-SFN-no-HF-cr:** it makes non sense decreasing terrestrial/satellite bandwidth ratio since the terrestrial network density can generally be improved over time and its bandwidth increased.

5.3.4 Forbidden configurations

Configurations that are not technically possible are listed below:

1) **SH-B-C-x-SFN-yes:** it is not possible with the SH-B radio configuration to have SFN since the modulation is not the same;

2) **SH-A-C-x-SFN-yes-HF-hm:** it is not possible in the SFN case to have two transport streams in hierarchical modulation, one being global and the other local; hierarchical modulation is possible but all local TS must be the same;

3) **SH-A-C-x-SFN-yes-HF-rm:** it is not possible in the SFN case to have different radio configurations between the satellite and terrestrial transmitters and so enable content removal.

6 Elements at Link and Service layers

6.1 Introduction

DVB-H presents a layered system structure that is one reason of its success: equipments operating on a specific layer can easily interconnect to equipments operating on an adjacent layer. Acknowledging this successful approach, the DVB-SH reuses to the most extent the DVB-H link and service layer in order to achieve seamless interoperability with DVB-H and to benefit from all available DVB-H link layer features as well as the already developed DVB-H ecosystem. This layered approach is presented in figure 6.1:

- a set of IPDC servers deliver IP streams, including the video streams;
- these IP streams are encapsulated by a DVB-SH IP encapsulator; the latter performs IP to MPE encapsulation according to EN 301 192 [9], PSI/SI insertion and MPE-IFEC protection and delivers an MPEG2 TS for the DVB-SH modulator;
- the DVB-SH modulators delivers a radio signal ultimately received by the DVB-SH receiver which performs baseband demodulation and decoding and processes the MPEG2 TS in the link layer client;
- the latter processes sections, MPE, MPE-FEC, MPE-IFEC, PSI/SI, and delivers an IP stream to the IPDC client;
- the IPDC client processes the IP streams, for example to deliver the ESG, the security decryption and the video and audio play out.

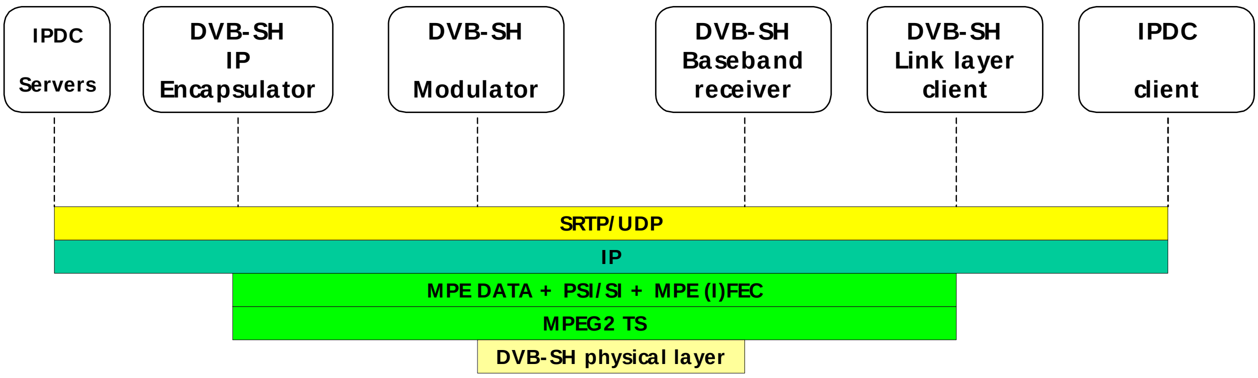


Figure 6.1: DVB-SH layered approach

Key features of DVB-SH link and service layers are:

- Support of Multi-Protocol Encapsulation:
  - DVB-H provides an IP multicast transport on top of MPEG2 Transport Streams (TS). To encapsulate the IP datagrams over MPEG2 TS, Multi-Protocol Encapsulation (MPE), (see EN 301 192 [9], clause 7) is applied. As the DVB-SH physical layer is also MPEG2 TS based, DVB-SH reuses MPE for the transport of IP datagrams over DVB-SH physical layers;
  - MPEG2 TS-based transport and MPE enable to reuse most signalling concept of DVB-H also for DVB-SH;
  - note that DVB-SH physical layer also supports the GSE but this is for further study.
- Support of time slicing:
  - DVB-H uses the real-time parameters, specifically the Delta-t information, conveyed within MPE and MPE-FEC headers in order to inform the start of the next burst. DVB-SH re-uses this concept: each MPE, MPE-IFEC and MPE-IFEC section carried by the MPEG2 TS over DVB-SH physical layer includes the same Delta-t information;

- this mechanism enables to power off the terminal during periods where no relevant bursts for this service are transmitted. This also enables hand-over even for receivers with a single demodulator in case the infrastructure provisions to appropriately synchronizes the transmitted TS;
- in addition, time slicing enables the efficient support of variable bit rate services since Delta-t can be adapted for each burst size. This is one way to efficiently support statistical multiplexing.

- Support of link layer protection:
  - DVB-H permits the use of link layer protection by applying MPE-FEC (see EN 301 192 [9], clause 9) to counteract terrestrial fading. DVB-SH also supports the use of MPE-FEC;
  - in addition, MPE-IFEC [reference] may be required in satellite coverage, especially with class 1 receivers;
  - by doing so, individual protection for each service is enabled. Depending on the service requirements and the physical layer performance, the transmitter can select from a variety of link layer parameters, e.g. using single burst MPE-FEC or multi-burst MPE-IFEC. Each FEC protection scheme can be fully configured to the service requirements thanks to a number of parameters;
  - simultaneous use of MPE-FEC and MPE-IFEC on the same elementary stream is for further study and therefore not allowed.
- Support of IPDC features:
  - DVB-SH is fully compatible with the DVB IPDC specifications, enabling a fast deployment of services on top of DVB-SH physical and link layers through the reuse of the IPDC protocol stack. Recommendations on the use of IPDC specifications in a DVB-SH context are for further study;
  - DVB-SH uses updated PSI/SI to convey system and program parameters. This enables smooth transition scenarios between DVB-SH and DVB-H networks, in particular for handovers: dual-mode receivers may receive content on one or the other technique seamlessly, provided PSI/SI are coherently signalled. The PSI/SI updates to support DVB-SH signalling are under study.

As DVB-SH relies to the most extent on DVB-H link layer technologies, this clause will only highlight the additions and differences of DVB-SH link layer compared to DVB-H. In case information on specific aspects on the DVB-SH link layer can not be found in this clause, the DVB-H link layer specification has to be used.

To address the modification and additions of DVB-SH link layer compared to DVB-H link layer, this clause is organized in three parts:

- the first part introduces of a new link layer forward error correction referred to as MPE-IFEC. This new technology is motivated by the fact that DVB-SH networks without terrestrial repeaters can results in challenging propagation channels ("Land Mobile Satellite"). Signal outages in the order of several seconds may be quite frequent, in particular due to blockage. These outage durations generally exceed conventional physical or link layer error correction capabilities. The interleaving depth of such systems is generally in order to several milliseconds for DVB-T or at most several hundreds of milliseconds with MPE-FEC. In order to counteract signal outages of several seconds, it is necessary to extend the protection capabilities of the forward error correction schemes. On may either use physical layer protection with very long interleavers or may use a link layer forward error correction scheme. DVB-SH provides both solutions: when transmitting to class 1 receivers, one should rely on a combination of short physical layer FEC and extended link layer FEC, referred to as IFEC and introduced in clause 6.2 of this clause; when transmitting to class 2 receivers, one may rely exclusively on a physical layer FEC with extended interleaving as described in clause 7;

- the second one addresses the backward compatibility of the DVB-SH physical layer with existing DVB-H time slicing and related legacy features. These include Time-slicing signalling, power saving, variable bit rate support and statistical multiplexing support. Clause 6.3 describes what level of compatibility such legacy DVB-H features is introduced by the DVB-SH physical layer for both terminal classes;

- the third part deals with mobility aspects and IPDC features (clause 6.4).

6.2 MPE-IFEC

6.2.1 Framework description

The MPE-IFEC has been designed taking account:

- the necessity to have a protection scheme to counteract the disturbances in DVB-SH transmission and reception environments;
- to achieve this, the MPE-IFEC, contrarily to MPE-FEC, encodes over several time-slice bursts, which enables to increase the interleaver duration;
- to address legacy requirements to existing DVB-H equipment;
- to enable a flexible solution which permits service specific adjustments;
- to provide the option to do further adjustments and optimizations during deployments.

The features supported by the framework are presented below.

- **Compatibility with DVB-H link layer (MPE sections):** MPE-IFEC is introduced in a way that it does not modify MPE section format, but only introduces one new sections type, the MPE-IFEC section. As MPE sections, MPE-IFEC sections convey almost identical real-time parameters. MPE-IFEC conveys exactly the same Time-slicing information as MPE. In addition, MPE-IFEC sections introduce additional information on the respective positions of MPE sections within the burst via the use of `MPE_boundary` bit (`frame_boundary` bit signals the end of the MPE-IFEC section part inside the burst).
- **Support of MPE-FEC:** permits the concurrent use of MPE-FEC and MPE-IFEC sections in one burst. However the support of simultaneous MPE-FEC and MPE-IFEC decoding in DVB-SH is for further study and therefore the parallel sending of MPE-FEC and MPE-IFEC on the same elementary stream is not allowed. This choice can be done on a per elementary stream bases so that MPE-FEC can be applied, on the same transport stream, on one elementary stream while another has MPE-IFEC. MPE-IFEC also relies on the same Application Data Tables (ADT) as the MPE-FEC and the transmitter may use the ADT of the MPE-FEC as the ADST of the MPE-IFEC. As a consequence, the number of rows of MPE-IFEC ADST is the same as the number of rows of MPE-FEC ADT. In case of losses of MPE sections, the receiver may use MPE-FEC sections, (exclusive) or MPE-IFEC sections, depending on the time_slice_fec_identifier signalling, but not both simultaneously, to recover the missing data.
- **Support long interleaving:** MPE-IFEC allows significantly enhanced performance (as measured by ESR(5)) in LMS channels when compared to MPE-FEC as demonstrated by simulations. This performance benefits can be achieved as the MPE-IFEC encoding process spans several time-slice bursts. MPE-IFEC requires a burst numbering which is signalled in MPE-IFEC headers.
- **Support of different service requirements:** The MPE-IFEC can be configured to enable a variety of configurations providing flexibility for the network operator. Guidelines are provided later in this clause on how to select these parameters.
- **Support fast zapping:** inter burst FEC protection is adversely affecting latency, and therefore also has influence on channel zapping times. This is due to the fact that for making use of all MPE-IFEC sections protection, the receiver must perform a late decoding and wait for reception of all MPE sections and MPE-IFEC sections relating to the encoding matrix to which a certain MPE section is assigned to. However, by sending MPE sections in each burst, immediate access and processing of these MPE sections is possible in case no errors have occurred. As reception continues, additional parity is received and MPE-IFEC protection can progressively protect the data by doing an early decoding. After some time, late decoding is possible.
- **Support a variety of FEC codes:** the framework can support a variety of codes, currently Reed Solomon is the only code supported by SSP, other FEC codes are for further study.

The MPE-IFEC provides an inter burst protection. Compared to MPE-FEC, this is achieved by either:

- the increase of the encoding matrix to sizes larger than one burst (one encoding matrix is filled during several successive bursts instead of a unique burst as in the DVB-H case);

- the parallelization of the encoding mechanism (instead of using only one matrix, data are distributed over a number of B parallel matrices, among M possible);
- the combination of the above both concepts.

The framework encoding process is shown in figure 6.2.

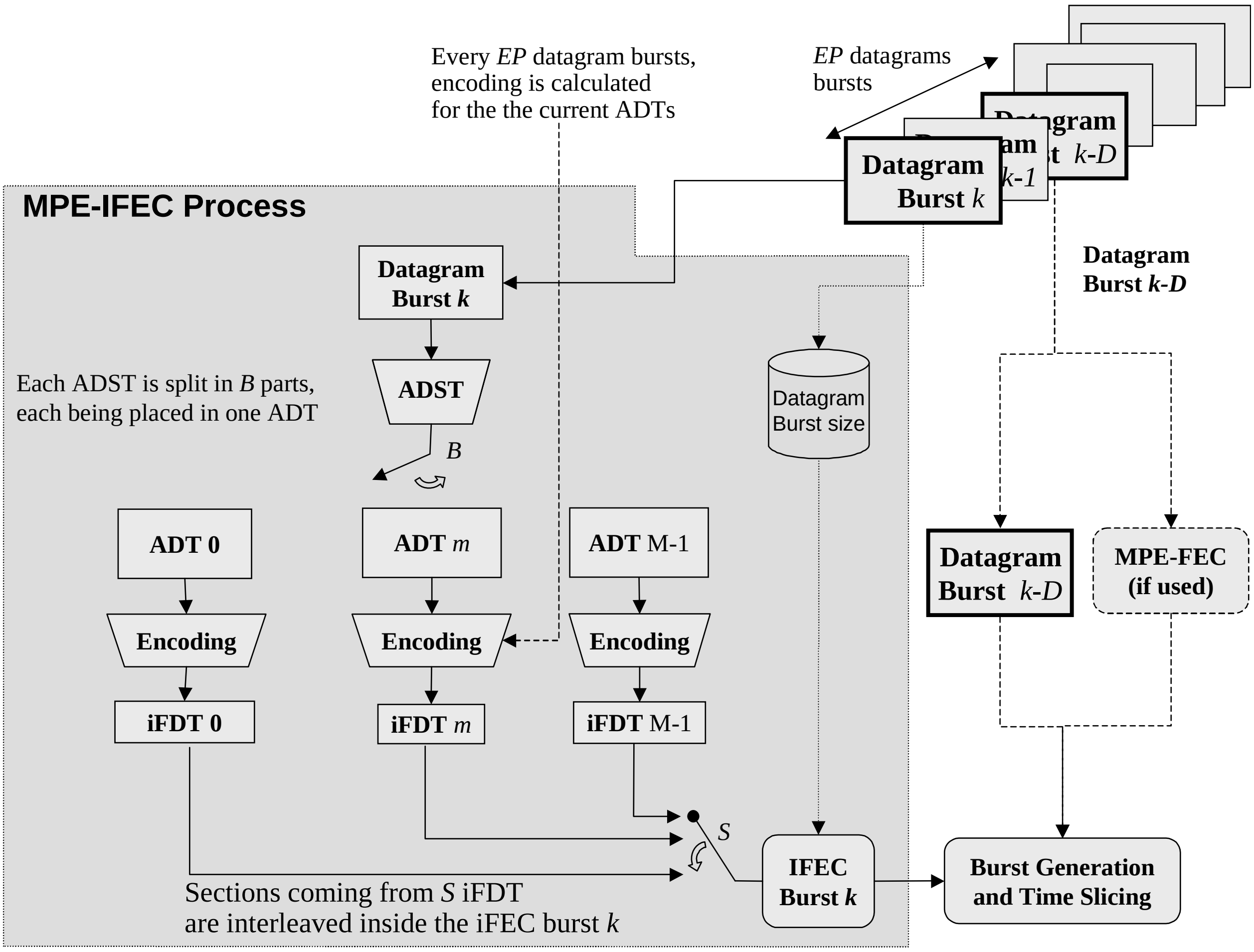


Figure 6.2: MPE-IFEC encoding process

The MPE-IFEC encoding process operates on a sequence of datagram bursts. Datagram bursts consist of a collection of datagrams, i.e. generally IP datagrams. Each datagram burst may be of different size and may comprise a different number of IP datagrams. For each datagram burst, a corresponding IFEC burst is created which encompasses MPE-IFEC sections from iFDTs of S encoding matrices. An MPE-IFEC section is comprised of a header, the parity symbols from one or multiple columns from the same iFDT, and a checksum. This IFEC burst is then multiplexed with all MPE sections of an original datagram burst, including the corresponding MPE-FEC sections if present. This multiplexing is done with the datagram just received in case the delay parameter D is set to D=0, or with a previously received datagram when D>0. This collection of MPE, MPE-FEC, and MPE-IFEC sections forms an MPE-IFEC time slice burst that is further processed. The sections within the IFEC time-slice burst are mapped to MPEG2 TS packets and then encapsulated in the DVB-SH encapsulation frames. Each datagram burst is mapped by an ADST (Application Data Sub Table) function on to the B ADTs (Application Data Tables) selected out of M parallel encoding matrices. Every EP bursts, an encoding process is carried out on one of the encoding matrices, generating from the ADT the corresponding iFDT (IFEC Data Tables) by applying an FEC encoding function. The resulting protection is achieved at the cost of some latency for generating and receiving parity data since the parity data is computed and spread over different bursts instead of one single burst in the MPE-FEC case.

The link layer encoding key parameters are given below:

- **EP:** this is the *encoding period* of the FEC process expressed in burst units. It refers to the frequency with which FEC is computed: an EP of 1 means that the encoding process occurs at every burst whereas an EP greater than 1 means that the encoding process occurs every EP bursts and the encoding matrix capacity is EP times greater so it takes EP times longer to fill the matrix with data. This EP normalizes all other parameters except D.

- **B:** this is the *encoding parallelization* expressed in encoding matrix units. Every burst is split into B parts distributed over B parallel encoding matrices. So actual interleaving depth is B*EP bursts, or a computed FEC has correction capabilities spanning EP*B successive bursts. The B matrices used for the encoding are updated every EP bursts and the list is completely refreshed after EP*B bursts.

- **S:** this is the depth of the FEC spreading factor; it means that produced FEC is interleaved over S iFDT. This enables to better protect the produced FEC.

- **D:** this is the delay applied to the data. Since FEC is computed with the received data, the normal sending order (send data immediately) would imply sending the FEC after the data. This parameters influences the zapping quality but has impact on end-to-end latency (see clause 6.2.4).

These four parameters enable to configure the encoding process in a flexible way. A typical configuration is obtained by setting EP=1 and B>1. This results in a sliding encoding process which enables to reuse several components of an MPE-FEC implementation: every received data burst is interleaved over B encoding matrices and one iFDT is computed applying the MPE-FEC Reed Solomon code.

To familiarize the reader with the MPE IFEC concept, figure 6.3 provides an example encoding and sending process with EP=2, B=9, S=2, and D=0. The horizontal axis represents the M=11 encoding matrices, numbered from 0 to 10, and the vertical axis represents the continuous datagram and MPE-IFEC time slice burst numbers (right column from 0 to 33).

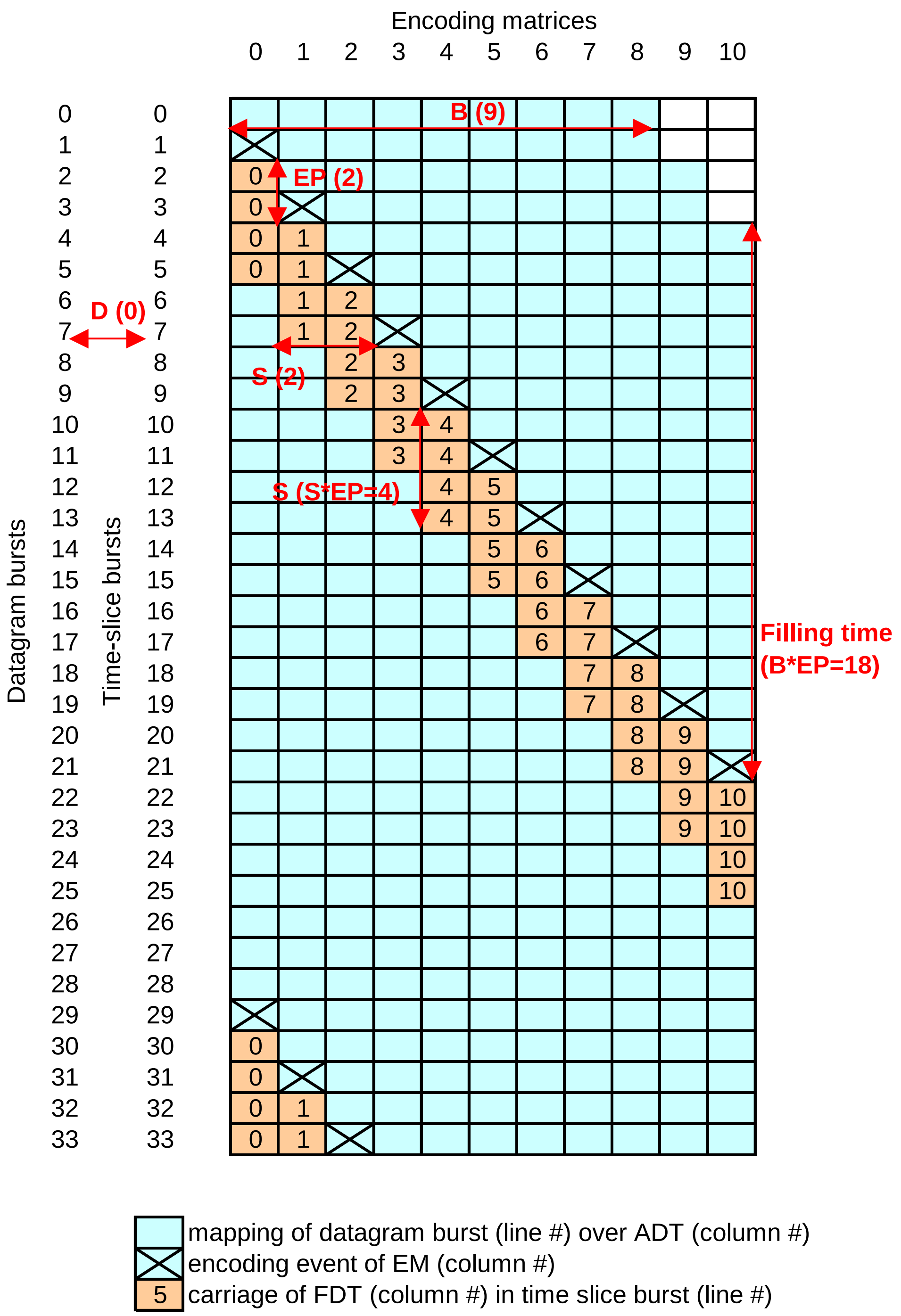


Figure 6.3: One example (EP=2, S=2, B=9)

Figure 6.3 can be read either horizontally or vertically:

- Horizontally:
  - the horizontal reading gives a representation of the sending logic:
    - the first two columns give the impact of the parameter *D* on the datagram burst sending: if *D* is set to 0 (here this is the case), the datagram burst is sent at the same time as the MPE-IFEC time slice burst. For instance, datagram 7 is sent with burst 7. Should the *D* parameter be greater than 0, then the datagram burst *k* would be sent in burst *k*+*D*;

        - for a given MPE-IFEC time slice burst number, if there is an orange cell on the corresponding line, it means that some FEC excerpted from the corresponding FDT is sent in the burst. For instance, at burst 7, MPE-IFEC sections from FDT 1 and 2 are sent. The number of "source" FDT interleaved in the current burst is dictated by the *S* parameter (here set to 2). The choice of the FDT matrix number is dictated by `ifdt_function` in MPE-IFEC specifications [attachment];

    - the horizontal view indicates also how the datagram burst is mapped on the ADST:

        - those ADT that receive columns from the current datagram burst are represented by the cyan colour: for example ADT 0 to 8 receive data from datagram burst 0 and 1, ADT 1 to 9 receive data from datagram burst 1 and 2, etc.;

        - one can see the influence of the *EP* parameter: the list of ADT receiving data is "updated" every EP datagram burst received. The number of the ADT used for mapping is given by `adt_function` in MPE-IFEC specifications [attachment].

- Vertically: the vertical reading gives a representation of the encoding logic:

    - every *EP* bursts, there is an "encoding event" indicated by a cross: one particular matrix encodes the data (it generates the iFDT from the ADT). After the completion of the encoding, the ADT can be freed from its data so that this ADT can be used to host new datagram bursts again. For instance, encoding matrix 4 has an encoding event at MPE-IFEC time slice burst 9. Its data can be reset after then; indeed this ADT it is not used during the 6 following bursts, then it is again filled starting at burst 20;

    - the encoding event creates FEC information that is transmitted during a successive number of bursts that is equal to *S*EP*. For instance, FDT 4, computed before burst 9 is sent, is sent during bursts 10 to 13 in 4 pieces;

    - since all other parameters are normalized by *EP*, the change of list of encoding matrices happens every *EP* bursts and the list has completely be renewed after *EP*B* bursts, which is also the time it takes to fill a matrix. For instance ADT 10 starts to be filled at burst 4 and stops at burst 21, 17 bursts later, which makes 18 bursts overall.

6.2.2 Usage in the context of the DVB-SH

In the DVB-SH context of the current implementation guideline release 1, only RS(255;191) (see EN 301 192 [9], clause 9.5.1) is used as encoding scheme. The MPE-IFEC framework is instantiated on this code with the following mandatory parameters: EP = 1, B ≥ 1, S ≥ 1, D ≥ 0.

Again, to provide some insight, we discuss an even simpler example with *B* = 3 and *S* = 1 where a completely lost burst can be recovered with a link layer FEC code rate of only 3/4. Table 6.1 shows, for each burst, what happens at transmitter and receiver sides and represent the event of burst "n" complete loss:

- on transmitter side:

    - the first row represents datagram bursts, the data to be transmitted;

    - second row represents the 4 parallel encoding matrices used by the link layer: ADT are filled by received datagram bursts and FDT are computed from the ADT;

    - third row represents the actually sent MPE-IFEC time slice bursts made of datagram bursts and IFEC bursts derived from one unique FDT. As *S* = 1, the IFEC burst is constructed from a single FDT;

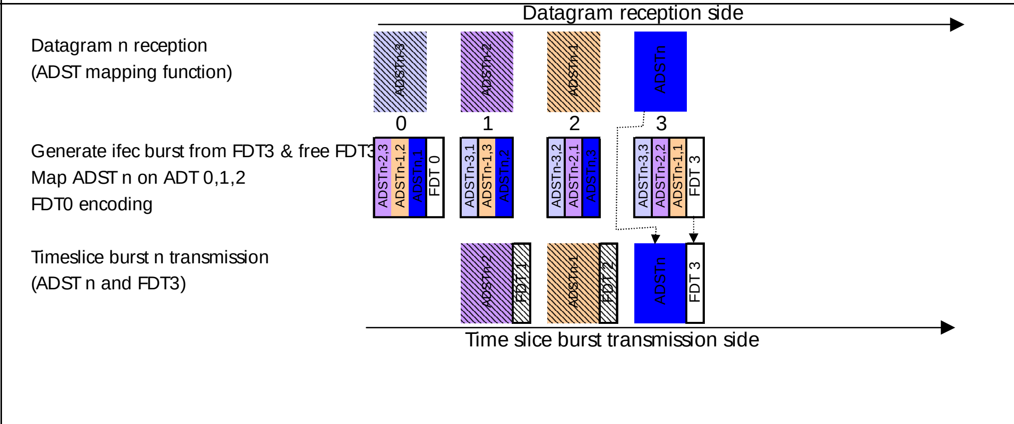
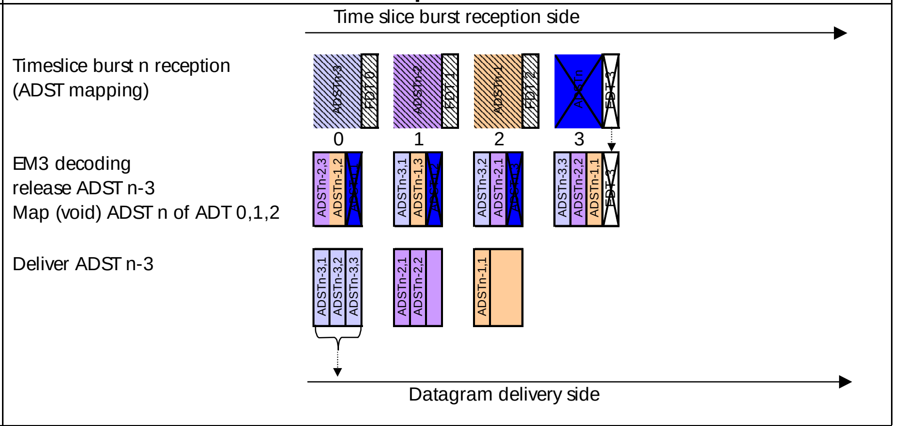
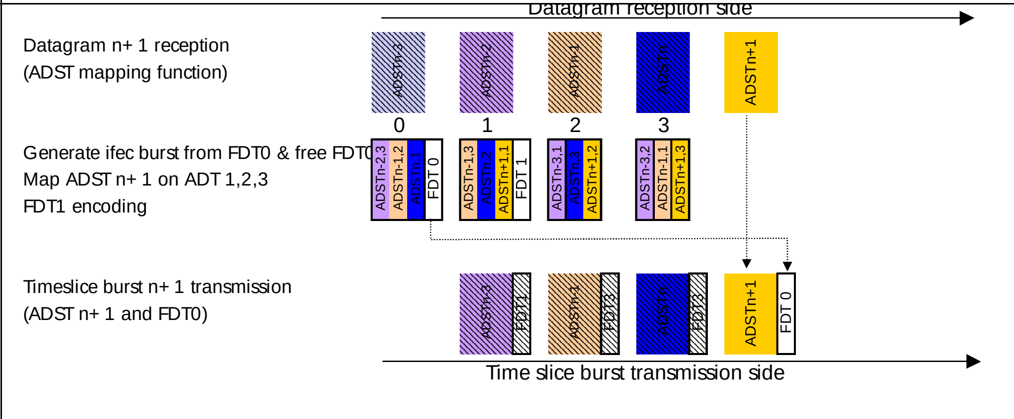
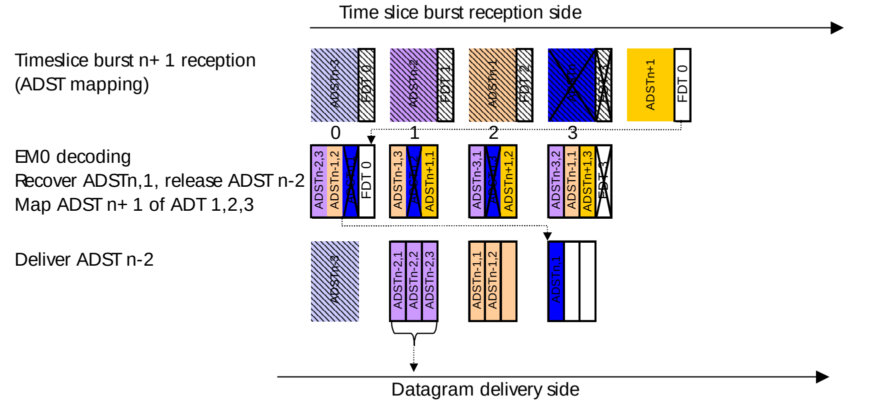
- on receiver side:

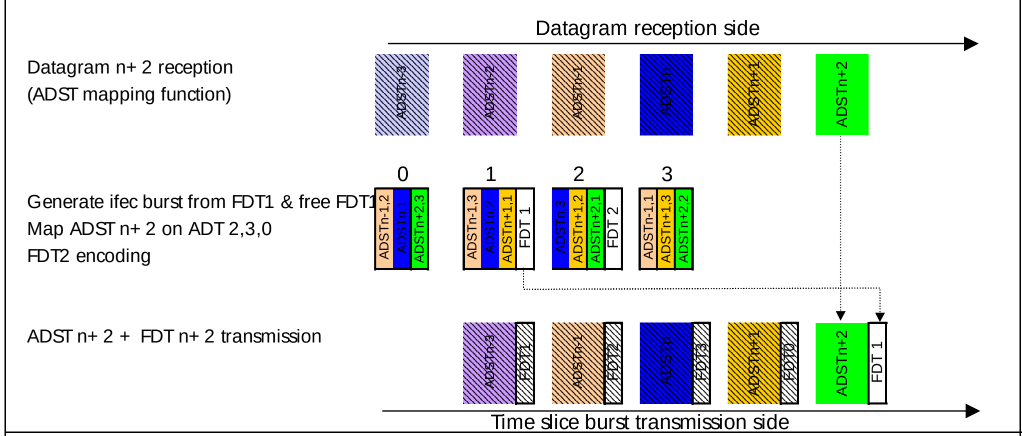
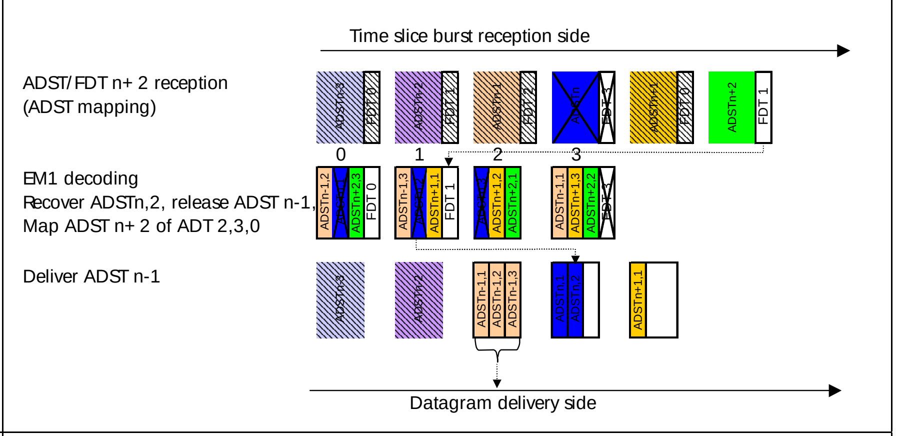
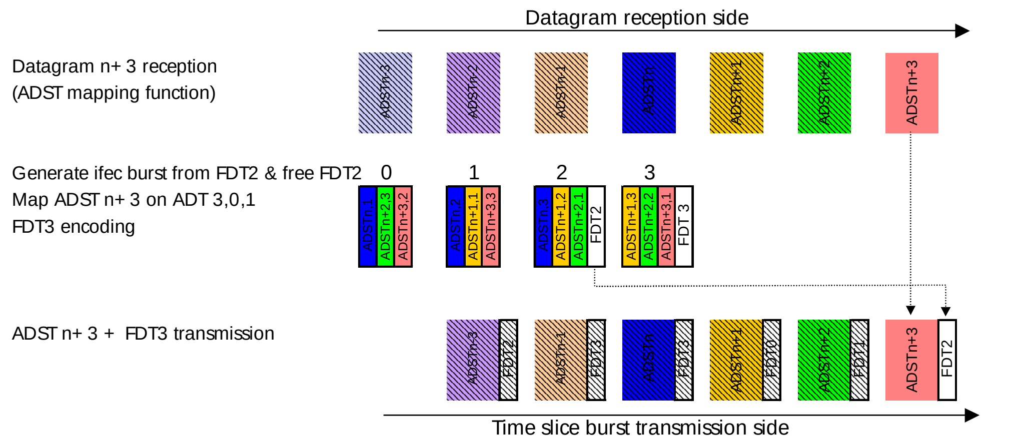
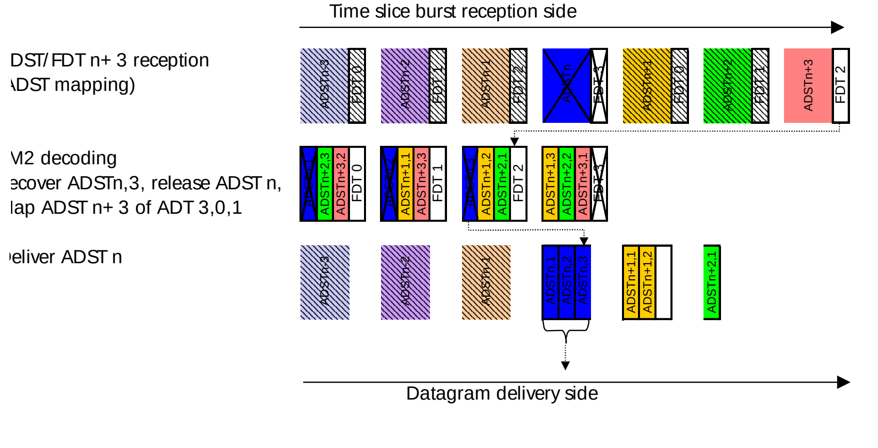
    - first row represents received time slice MPE-IFEC burst, possibly with their losses;

    - second row represents the 4 parallel decoding matrices: ADT are filled by received datagram burst and FDT are filled by received MPE-IFEC burst;

    - third row represents the data to be sent to the higher layer, whether it may be received "lossless" or recovered by FEC decoding.

Table 6.1: example of complete burst recovery

<table>
<tr><th>Step</th><th>Transmission side</th><th>Reception side</th></tr>
<tr><td rowspan="2">n</td><td>
</td><td>
</td></tr>
<tr><td>Transmission of datagram burst n; previously computed FDT 3 is sent in MPE-IFEC time slice burst n; datagram burst n is being mapped over ADT 0, 1, 2; FDT 0 is computed</td><td>We assume previous bursts k<n have been well received but burst n is completely lost (data and FEC). We need to map a void ADST to avoid glitches inside ADT.</td></tr>
<tr><td rowspan="2">n+1</td><td>
</td><td>
</td></tr>
<tr><td>Transmission of datagram burst n+1; previously computed FDT 0 is sent in MPE-IFEC time slice burst n+1; datagram burst n+1 is mapped over ADT 1,2,3; compute FDT 1</td><td>Reception of MPE-IFEC time slice burst n+1; decode FDT 0 and recover ADT(n,1); map ADST n+1 into ADT 1,2 and 3.</td></tr>
</table>

Step	Transmission side	Reception side
n+2		
	Transmission of datagram burst n+2; previously computed FDT 1 is sent in MPE-IFEC time slice burst n+2; datagram ADST n+2 is mapped over ADT 2,3,0; compute FDT 2	Reception of MPE-IFEC time slice burst n+2; decode FDT 1 to recover ADT(n,2); map data burst n+2 into ADT 2,3 and 4.
n+3		
	Transmission of datagram burst n+3; previously computed FDT 2 is send in MPE-IFEC time slice burst n+3; datagram burst n+3 is mapped over ADT 3,0,1; compute FDT 3	Reception of burst n+3; decode FDT 2 to recover ADT(n,3) ⇒ ADT n has been completely recovered; map ADST n+3 in ADT 3, 0 and 1.

6.2.3 A practical example

6.2.3.1 Introduction

In the rest of the document, we will continuously refer to the following typical example: EP=1, B=6, S=4, D=0, MPE-IFEC code rate = 2/3. Assume for further simplicity that the service to be protected has a constant bit rate of 300 kbps and IP packets of 1 000 bytes are sent.

The ADST is determined by its T=1 024 rows and has on average 37 data columns (a data column is a column with at least 1 non padding byte). To accommodate with local variations, we fix C to 40 so that 4 more additional data columns could be absorbed in case of small traffic variations.

The number of ADT over which the ADST are padded is equal to B+S (9).

Each time an encoding event happens, an FDT of N columns is created. Based on the code rate 2/3 and the 37 non padded columns, 19 columns on average are excerpted from this FDT for transmission.

6.2.3.2 Time_slice_fec_identifier

The time_slice_fec_identifier provides the following information about the stream inside the INT.

Table 6.2: Time_slice_fec_identifier

Syntax	Number of bits	Value	Comment
time_slice_fec_identifier_descriptor () {			
descriptor_tag	8	01110111	0x77
descriptor_length	8	00001011	11
time_slicing	1	1	Time slicing is used
mpe_fec	2	00	Mpe fec is not used
reserved_for_future_use	2	11	N/A
frame_size	3	11	1024 rows
max_burst_duration	8	00001001	200 ms
max_average_rate	4	0101	512 kbps
time_slice_fec_id	4	0001	MPE-IFEC is used
T_code	2	00	RS(255,191) is used
G_code	3	000	G=1
Reserved for future use	3	111	N/A
R	8	01000000	R=64
C	13	00101000	C=40
Reserved for future use	3	111	N/A
B	8	00000110	B=6
S	8	00000100	S=4
D	8	000000	D=0
EP	8	00000001	EP=1
}			

This information is firstly retrieved by the receiver to determine:

- what type of link layer protection is active on the elementary stream of interest (none, MPE-FEC, MPE-IFEC, exclusive choices);
- if one link layer is active, which are its main parameters.

Note that the time_slice_fec_identifier parameters of MPE-FEC have not been modified at all: only additional parameters have been added inside the id_selector_bytes (not used by DVB-H) and 1 reserved for further study bit has been used in `mpe_fec` and `time_slice_fec_id` fields. Therefore, when new values are not needed, parameters used by MPE-FEC are reused by the MPE-IFEC with exactly the same definition (for instance `frame_size`, `max_burst_duration`, `max_average_rate`).

6.2.3.3 MPE-IFEC parameter derivation

The receiver can now derive from the time_slice_fec_identifier the required parameters for performing the MPE-IFEC decoding. Applying the parameters selection as specified in MPE-IFEC specifications [attachment], clause 5.3 we have the following list of MPE-IFEC parameters useful for decoding.

Table 6.3: Example of DVB-SH MPE-IFEC parameters selection

Parameter	Unit	Value	Description	Signalling	Scoping
EP	Datagram burst	1	IFEC Encoding Period	Direct via Time_slice_fec_identifier	Time_slice_fec_identifier
D	Datagram burst	0	Datagram burst sending delay	Direct via Time_slice_fec_identifier	Time_slice_fec_identifier
T	rows	1 024	Number of ADST, ADT, iFDT rows: T= MPE-FEC Frame rows /G	Indirect via Time_slice_fec_identifier	Time_slice_fec_identifier
C	columns	40	Number of ADST columns	Direct via Time_slice_fec_identifier	Time_slice_fec_identifier
R	sections	64	Maximum number of MPE IFEC sections per MPE-IFEC Time-Slice Burst	Direct via Time_slice_fec_identifier	Time_slice_fec_identifier
K	columns	40	Number of ADT columns = EP*C	Indirect via Time_slice_fec_identifier	Time_slice_fec_identifier
N	columns	64	Number of iFDT columns = EP*R*G	Indirect via Time_slice_fec_identifier	Time_slice_fec_identifier
G	columns	1	Maximum number of iFDT columns per IFEC section	Direct	Time_slice_fec_identifier
M	ADT	M = B + max(0,S - D) + max(0,D - B) = 10	Number of concurrent encoding matrices M	Indirect (formula dependent on T_code and given in the parameter definition of clause 5)	Time_slice_fec_identifier
k_{max}	N/A	256 - 256 [i.24] = 250	Modulo operator for MPE-IFEC time slice burst counter	Indirect (formula dependent on T_code and given in the parameter definition of clause 5)	Time_slice_fec_identifier
j_{max}	N/A	10	Maximum backward pointing for datagram burst size used in PREV_BURST_SIZE parameter in clause3.5	Indirect (formula dependent on T_code and given in the parameter definition of clause 5)	Time_slice_fec_identifier

This parameter derivation, in particular, enables the receiver to know if it has the capacity to process the MPE-IFEC decoding extensively. The memory is derived from encoding matrix size given by *T*, *C*, *K*, *N* and the number of encoding matrices (for more information on memory sizing, please refer to clause 6.2.6).

6.2.3.4 ADST mapping function

ADST mapping and padding: the ADST function maps the received IP packets on a matrix of size 1 024 rows by 40 columns. This results in the figure below.

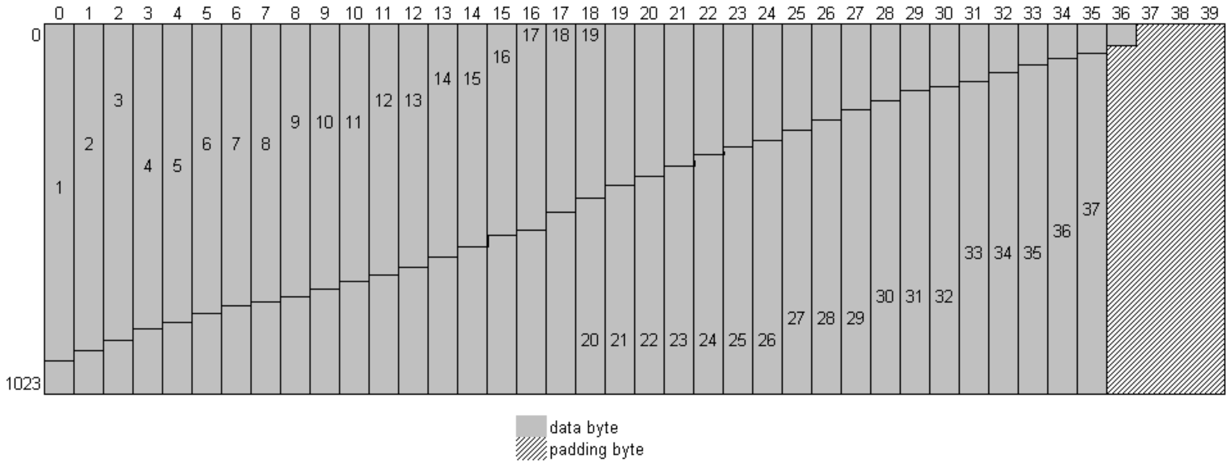


Figure 6.4: ADST and padding

In this figure, one can see that 37 columns (from number 0 to 36) have at least 1 data byte and are so called data columns (even if one, the N°36 has very few data bytes) and 3 are padding columns (from 37 to 38) because they all their bytes are padding. Another remark is that, generally, the loss of 1 IP packet during transmission will occur losses on two columns, with the exception of the first IP packet.

6.2.3.5 ADT mapping

To illustrate the ADST to ADT mapping, we take two views, the ADST and then the ADT.

6.2.3.6 ADST view

From the ADST point, it is observed that the ADST columns are inserted ("shifted" as said in MPE-IFEC specifications [attachment]) into different ADTs. An example of such ADST columns distribution is given in figure 6.5, for the case of datagram burst 0.

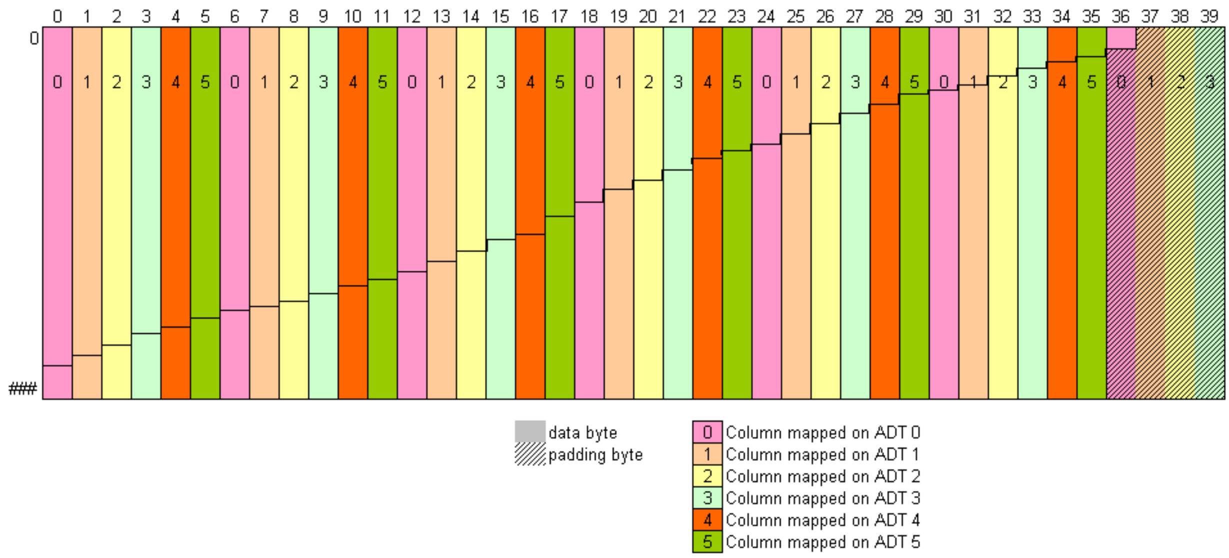


Figure 6.5: ADST to ADT mapping (ADST view)

It can be seen in figure 6.5 that the columns are interleaved between the different ADTs (they are not taken "in block"). Additionally, they are distributed over B ADTs (here B=6 so ADT 0 to ADT 5). The last columns, that are padding columns, are (and must be) also distributed over the B ADTs.

6.2.3.7 ADT view

Structure of an ADT

It is observed that an ADT within a single encoding matrix hosts columns from *B* different datagram bursts. The mapping is defined by the ADST profile (its number of data and padding columns) and the `adt_index` function which determines in which ADT the ADST columns are shifted (inserted).

Due to the shifting done at each burst (at each received burst, new ADST columns are introduced in *B* ADTs), the "population" of one ADT is progressively done between two successive encoding events on this ADT (once the encoding event has occurred, the ADT is reset and ready to accept new columns). Due to the fact that *M* can be larger than *B*, all ADTs do not include columns from all ADSTs: only when *B*=*M* do we have the situation when all ADTs receive columns from any ADST. In the general case when *B*>*M,* during a number of burst equal to *M*-*B*, the recently encoded and reset ADT does not receive any data. When data are again inserted from burst k and ADT m, the following relationship is valid:

adt_index(k+B-1,0)=m, or (k+B-1)[M]=m, or there exist an "i" such that k = M*i + m–B+1

The ADT will then be filled by next *B* bursts, so bursts k, k+1,…k+*B*-1

We have presented as an example in figure 6.6 the ADT 0 after first 20 datagram bursts have been received. In this figure, we can see that the last columns to have been "shifted" in the ADT 0 (the ones that are on the rightmost part), are coming from burst 20, and the first ones to have been shifted in are on the leftmost part (due to the "push" from right to left of newer columns) and are coming from burst 15, which respects the above relationship.

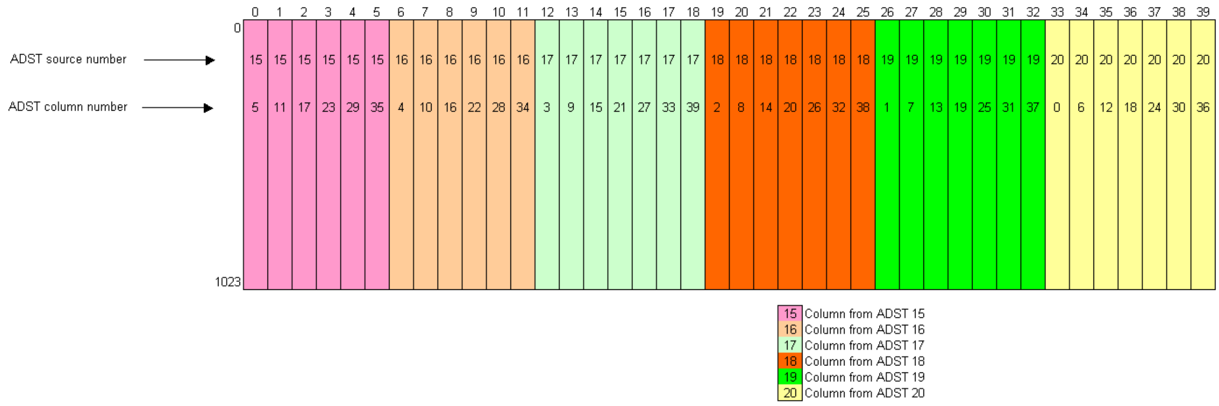


Figure 6.6: ADST to ADT mapping (ADT view)

It can be observed in this figure that the source ADST columns are grouped by "sub-block". This is due to the "shifting" function that inserts a certain number of columns from the same burst. Note also that the source ADST columns indexes are not continuous but separated by *B* (for example see column indexes coming from ADST 15: 5, 11, 17, 23, 29).

The `adt_column` function

More generally, a complete ADT ready for encoding has the following structure presented in figure 6.7.

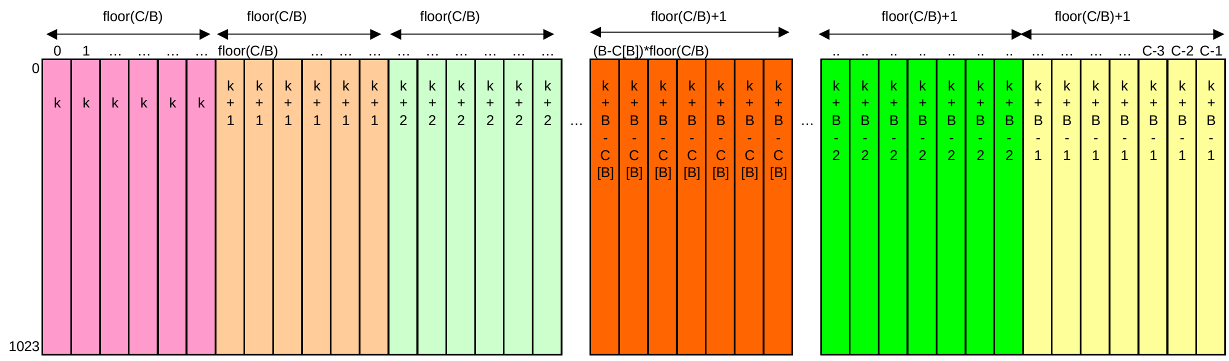


Figure 6.7: Generic ADT structure

The ADT is comprised of B sub-blocks, each sub-block being consisting of a number of columns coming from the same datagram burst and corresponding ADST:

- Each of the first (*B*-*C*[*B*]) sub-blocks hosts floor(*C/B*) columns from same ADST; the floor(*C/B*) ADST columns of the k[th] block, k∈[1;*B*-*C*[*B*]], are the columns j, j∈[0;*C*-1], such that k = *B*-j[*B*], so the formulation of the column numbers is the following:

For k = 1:1:B-C[B], for l = 0:1:floor(C/B)-1, j(k,l)=l*B+(k-1)

- Each of the last C[B] sub-blocks host floor(C/B)+1 columns of the same ADST; the floor(C/B)+1 ADST columns of the k[th] block, k∈[B-C[B]+1;B], are the columns j, j∈[0;C-1], such that k = B-j[B], so the formulation of the column numbers is the following:

For k= B-C[B]:1:B, for l = 0:1:floor(C/B)-1, j(k,l)= l*B+(k-1) and j(k, floor(C/B))= floor(C/B)*B + (B-k)

Hence, the shifting operation in the MPE IFEC specification for each ADST to ADT mapping in case of *EP*=1 may be realized as ***deterministic positions*** inside the ADT when the time to encode the matrix has arrived. These positions are given by the following formula:

adt_column(j) = (B–(j[B]+1))*floor(C/B) + max(0;(C[B])-(j[B]+1)) + floor(j/B)

The use of such `adt_column` formula is easier than implementing the "shifting" method since it establishes a bijection between ADST and ADT columns. An ADST can be then considered as a collection of *C* "pointers" to the ADT columns. This approach is presented in figure 6.7 where the column number of the original ADST are presented below the ADST number (this is the pointer origin) and the column number of the destination ADT are presented at the top of the column (this is the pointer destination). Mapping ADST results in setting the pointers destination to the correct ADT column using the `adt_column` function and this update occurs only once instead of *B* times with the original shifting mechanism. The advantages of this approach are twofold:

- This helps storage optimization since storage is done inside the ADTs rather than on the ADSTs, each ADST being a list of *C* pointers to *C* columns in *B* ADTs. More memory implementation discussion that uses this pointer approach can be found in clause 6.2.6.

- This helps the decoding since when an ADT has been decoded and some of its columns been corrected, the ADST is automatically "refreshed" because the pointer points now to a correct column. More decoder implementation discussion using this concept can be found in clause A.6.2.

Padding:

One ADT columns may not necessarily host data, but also padding bytes and padding columns. Due to the construction of the ADST function that positions padding columns at the end of the ADST, these padding columns always occur at the end of the "sub-blocks" and therefore padding columns are inserted inside the ADT at different positions, but certainly not at the end of the ADT as it is usually the case in MPE-FEC encoding. Assuming in our example that the 4 last columns of each ADST are concerned by padding (the first one is a data column with some data at the top of the column, whereas the 3 last ADST columns are complete padding columns), the ADT will have the following aspect whereby the shaded area represents padding.

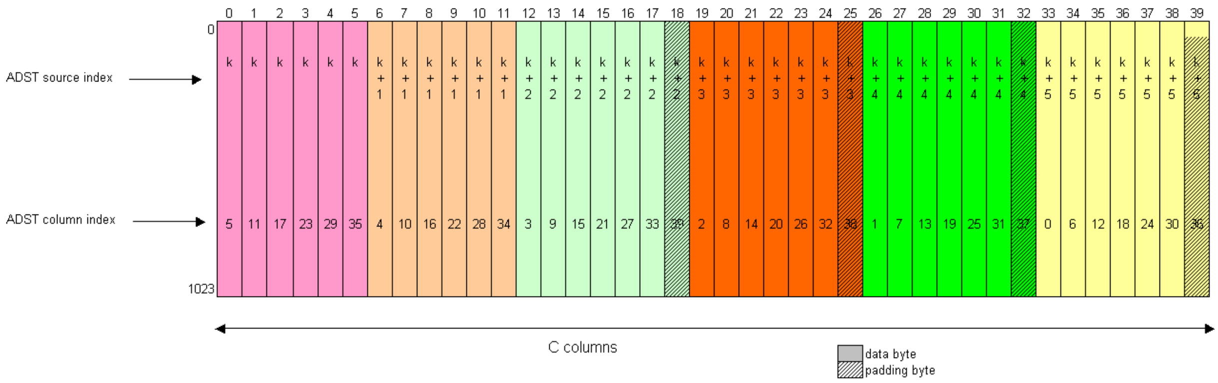


Figure 6.8: ADT aspect with interleaved padding columns

The generation of the Reed Solomon operates on a matrix of 191 data columns, so there is an additional padding of 191-C columns so that resulting ADT is as shown in figure 6.9.

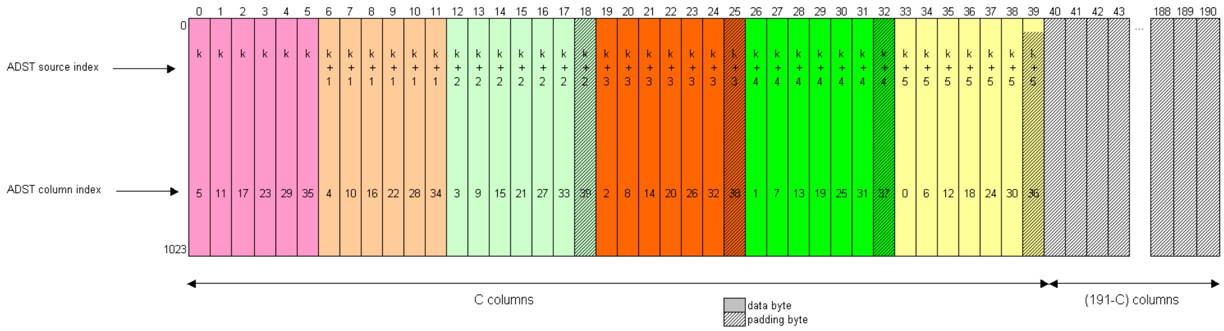


Figure 6.9: Final ADT aspect

The padding is equivalent to a code shortening as defined in EN 301 192 [9], clause 9.3.3.1. Code shortening is required to obtained the signalled ADT size *K*.

6.2.3.8 FDT generation and code rate computation

As soon as the ADT has been constituted, the FEC encoding function is applied to generate the iFDT using the Reed-Solomon code (255, 191). The FEC columns can be as high as 64 but usually only a fraction of these FEC columns is actually transmitted over the air (the first ones), which is equivalent to code puncturing as defined in EN 301 192 [9], clause 9.3.3.2. Puncturing is required to obtain the signalled iFDT size *N*.

With *EP*=1, for each processed datagram burst, exactly one encoding matrix is processed, i.e. the iFDT is generated.

Definition: the actual code rate is computed on a per encoding matrix basis based on the structure of the ADT and the iFDT. So for each encoding matrix m, we have:

$$code\_rate_{actual}(m) = \frac{nof\_data\_columns(m)}{(nof\_data\_columns(m) + nof\_fec\_columns(m))}$$

The actual code rate is usually taken as near as possible to a code rate target, and expected to be the same for all encoding matrices: $\forall m \in [0; M-1], code\_rate_{actual}(m) \approx code\_rate_{target}$. This target code rate is used to compute the number of FEC columns within each iFDT:

$$nof\_fec\_columns = ceil\left(nof\_data\_columns * \frac{1 - code\_rate_{target}}{code\_rate_{target}} \right)$$

.There are at least two ways for computing this number of columns:

Method 1: a fixed number of FEC columns is transmitted in each IFEC slice burst by assuming, for all m, nof_data_column(m)=C' where C'≤C and C' is fixed. Whatever the real completion of the ADT, the transmitted number of RS columns will remain the same. In CBR situation (nof_data_columns~C), this will generate a $code\_rate_{actual}$ almost equal to $code\_rate_{target}$. But in case of VBR, it will make the code rate variable, possibly by large means, depending on variations of nof_data_columns datagram burst by datagram burst and encoding matrix by encoding matrix: when nof_data_columns is equal to C, $code\_rate_{actual}$ is almost equal to the target value $code\_rate_{target}$, but when nof_data_colums is small compared to C, then $code\_rate_{actual}$ may become quite low compared to $code\_rate_{target}$, nearing 0).

Method 2: another way is to compute the nof_fec_columns based on the actual nof_data_column(m) and target code rate. This ensures a quasi constant $code\_rate_{actual}$, whatever the actual nof_data_columns. The precision of the difference between actual and target is a function of the nof_data_column and is better when this latter increases.

These two options for the selection of the RS columns and their impact on $code\_rate_{actual}$ are presented in figure 6.10.

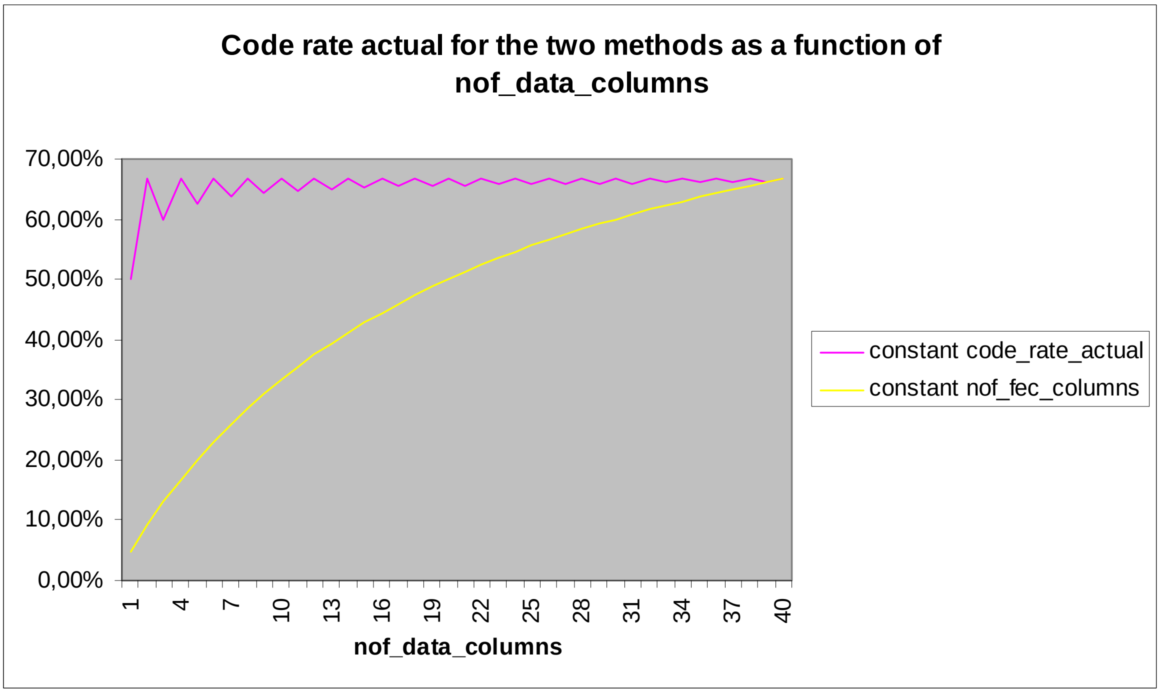


Figure 6.10: Different strategies for RS columns generation

In general for correct support of VBR and statistical multiplexing, the second option is recommended. The first options is recommended when a pure CBR traffic is protected. In our typical case, we assume a $code\_rate_{target}$ of 2/3. Based on the first option, we will assume that the number of FEC columns is equal to $ceil\left(37 * \frac{1-2/3}{2/3}\right) = 19$.

6.2.3.9 IFEC burst generation

Once the RS columns are generated inside one FDT, these columns are stored and interleaved before being inserted inside a time slice MPE-IFEC burst. This is where the S parameter is used: the MPE-IFEC sections inside one IFEC burst are coming from S successive FDT and the index position of one section inside the IFEC burst enables to derive the original FDT. This spreading effect is shown in figure 6.11. It can be seen that, in every IFEC burst i+4, there are:

- 5 IFEC sections coming from FDT i+3.
- 5 IFEC sections coming from FDT i+2.
- 5 IFEC sections coming from FDT i+1.
- 4 IFEC sections coming from FDT i.

More generally, given the MPE-IFEC time slice burst number k', $k' \in [0;k_{max}-1]$, the IFEC section index i, $i \in [0;18]$, the functions `ifd_index` and `ifdt_column` enable to derive the original iFDT index and iFDT column:

ifdt_index(k',j)=(k'-j[S]+M)[M]

ifdt_column(k',j)=j

If nof_fec_columns(i) gives the number of FEC columns present in FDT i, the sending will be as follows:

- in the nof_fec_columns(i)[S] first IFEC bursts, we send floor(nof_fec_columns(i)/S)+1 IFEC sections;
- in the remaining S-nof_fec_columns(i)[S] IFEC bursts, we send floor(nof_fec_columns(i)/S) IFEC sections.

One representation of IFEC burst i+4 is presented in figure 6.11.

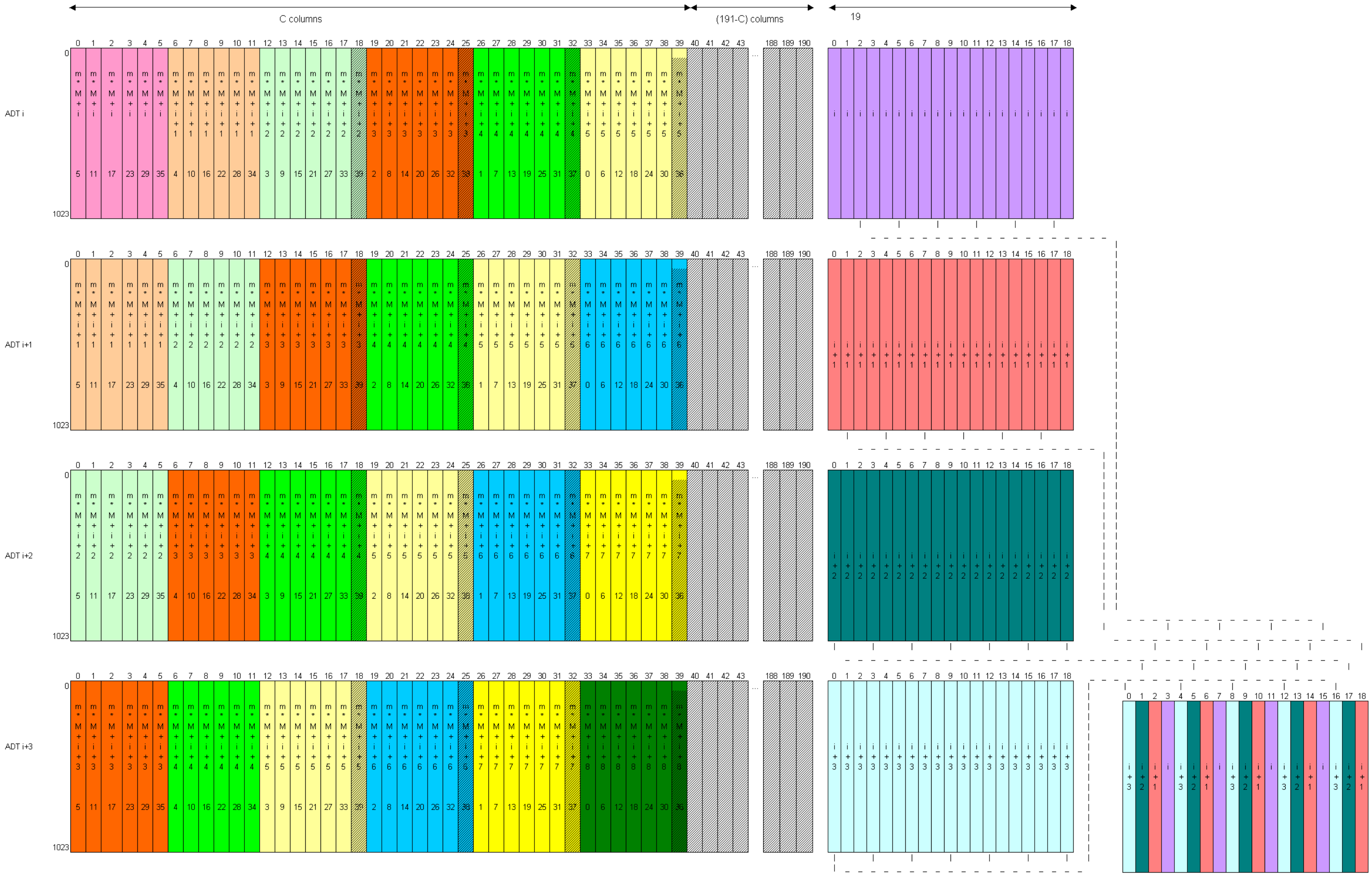


Figure 6.11: Sending arrangement of IFEC burst i+4

Please note that the section index may not be consecutive and there could be index discontinuities when an iFDT has "run out of available columns". Such situations may occur in VBR situations with a fixed code rate: one ADT i has fewer data columns than the following ADT i+1 so that their corresponding iFDT have also different FEC columns available for transmission. Take the current example where the number of column is 19 on all iFDT except iFDT i where only 18 columns are available. The IFEC bursts composition are represented hereafter:

Table 6.4: Source of MPE-IFEC sections in different IFEC bursts

Source iFDT IFEC burst	i-3 (19)	i-2 (19)	i-1 (19)	i (18)	i+1 (19)	i+2 (19)	i+3 (19)	i+4 (19)
i+1	4	5	5	5	-	-	-	-
i+2	-	4	5	5	5	-	-	-
i+3	-	-	4	5	5	5	-	-
i+4	-	-	-	3	5	5	5	-
i+5	-	-	-	-	4	5	5	5
…								

The MPE-IFEC time slice burst i+4 will have only 18 IFEC sections with the following indices.

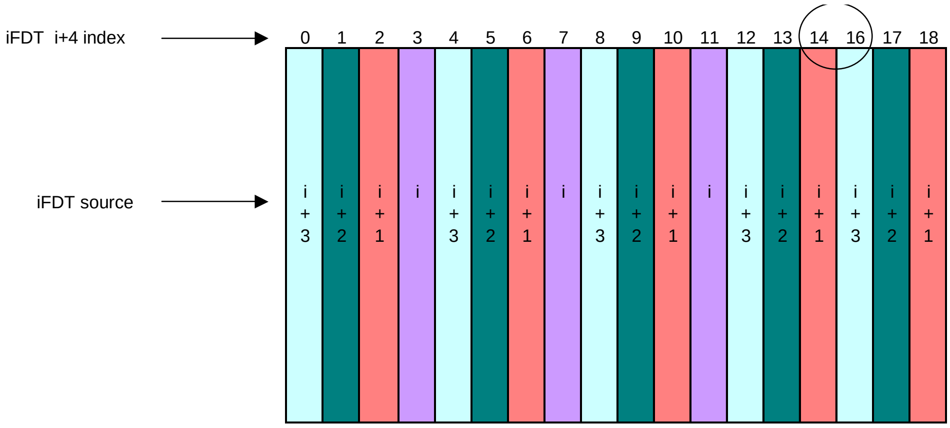


Figure 6.12: IFEC burst I+4 in case of section discontinuity

It is also to be noted that the interleaving over the S IFEC bursts will smooth IFEC burst variations. More scenarios on usage of the section indices are discussed in clause 6.2.7.

6.2.3.10 MPE-IFEC section header

In addition to the static configuration of the time_slice_fec_identifier, real-time signalling is conveyed in the MPE-IFEC headers. Figure 6.13 provides an overview on the syntax of the MPE-IFEC section and its header.

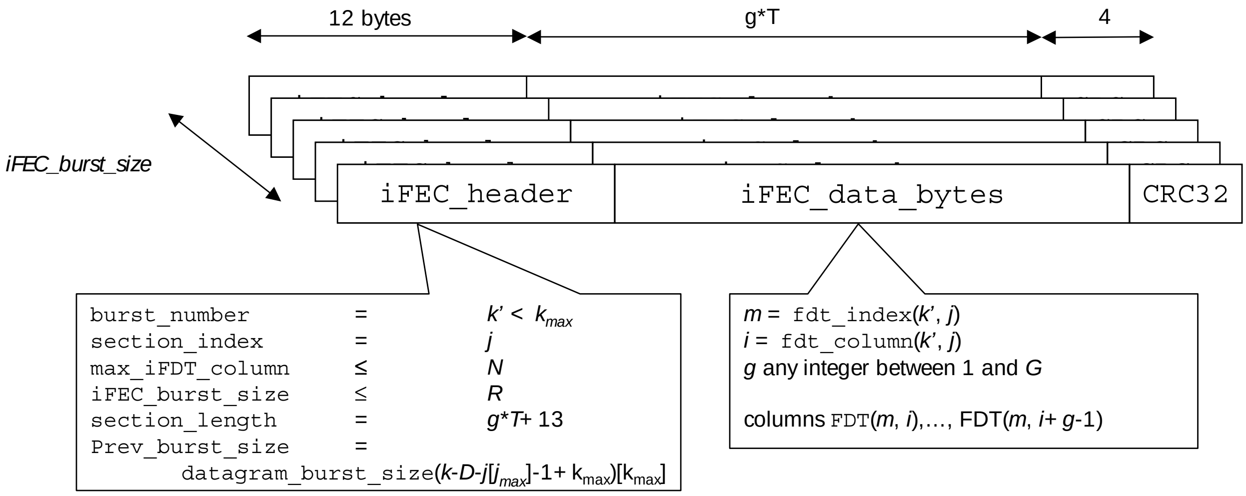


Figure 6.13: MPE-IFEC header and payload structure

For our particular example, the indices have the following information:

- `burst_number` = *k'*, where $k' \in [0;249]$.
- `section_index` = *j*, where $j' \in [0;18]$.
- The iFDT index is obtained as *m* = `ifdt_index`(*k'*, *j*).
- IFEC_data_bytes are obtained from `iFDT`(*m*) and correspond to iFDT columns `ifdt_column`(*k'*, *j*).
- `max_iFDT_column` = floor(255**max_iFDT_column*(*m*)/*N*)=floor(255*19/64)=75.
- `IFEC_burst_size` = 19 (total number of IFEC sections included in this IFEC burst).
- `section_length` = 1 037.
- `prev_burst_size` = *datagram_burst_size*$((k\text{-}D\text{-}j[j_{max}]\text{-}1+k_{max})[k_{max}])$.

Usage of these fields is multiple and several examples are provided in clause 6.2.7.

Real-time parameters according to table 6.5 are inserted. These real-time parameters usage are detailed in clause 6.2.3.11 as they relate to the sending arrangement within an MPE-IFEC time slice burst.

Table 6.5: MPE-IFEC real-time parameters

Syntax	Number of bits	Identifier
real_time_parameters () {		
delta_t	12	uimsbf
MPE_boundary	1	bslbf
frame_boundary	1	bslbf
prev_burst_size	18	uimsbf
}		

6.2.3.11 Burst sending arrangement

Considerations on the burst, Time-slicing and Delta-t parameter

In the MPE-IFEC specifications [MPE-IFEC], there is a reference to an (MPE-IFEC) time slice burst ([MPE-IFEC], clause 2.3.7). This MPE-IFEC time slice burst has the same definition as the burst in DVB-H, as specified in EN 301 192 [9], clause 9.1]: *"a burst is a set of sections delivered on an elementary stream. Between two consecutive bursts there is a period of time when no sections are transmitted on the particular elementary stream. Each burst indicates the start time of the next burst within the elementary stream"*. This definition of burst includes potentially sections of different kinds, like MPE, MPE-FEC and MPE-IFEC, so that the burst is by nature an open concept. In [IFEC], the IFEC time-slice burst is therefore used to designate all the sections, including MPE, MPE-FEC and MPE-IFEC, that age timely grouped. For simplification and clarity reasons, we will in the following refer to the IFEC time-slice burst by the simple term *burst*, not to confuse it with the DVB-H time-slice burst that is referred by the term time-slice burst. Another kind of burst has been defined in MPE-IFEC specifications [MPE-IFEC] called the IFEC burst. This includes the only MPE-IFEC sections. The terminology is presented in the table below.

Table 6.6: Terminology

Name in the present document	Concerned section kinds	Name in DVB-H	Name in IFEC
Burst	All	burst	Burst or MPE-IFEC time slice burst
Time slice burst	MPE and MPE-FEC	Time slice burst	N/A
IFEC burst or MPE-IFEC burst	MPE-IFEC	N/A	IFEC burst or MPE-IFEC burst

Contrarily to the burst, the definition of Time-slicing given in the same document (*"Time Slicing: Method to deliver MPE sections and MPE-FEC sections in bursts"*) refers to the only MPE and MPE-FEC sections, excluding any additional types of sections (like MPE-IFEC). Additionally, the definition of Delta-t given in the clause 9.10 makes a clear reference to a *time-slice burst* (*"The field indicates the time (Delta-t) to the next Time Slice burst within the elementary stream"*).

Note that Delta-t information provided by MPE-IFEC sections **must** be exactly the same as Delta-t information provided by MPE and MPE-FEC sections since, otherwise, the Delta-t provided by different sections within same burst-MPE/MPE-FEC on one side, MPE-IFEC on the other side could point to different positions within the next burst and hence create inconsistent signalling. This is why the definition of the Delta-t field in MPE-IFEC sections is exactly the same as the definition of the Delta-t field in MPE and MPE-FEC sections: *"Delta-t is a time offset indicating the time from the start of the transport packet carrying the first byte of the current IFEC section to the start of the transport packet carrying the first byte of next* ***time-slice burs****"*.

Therefore, the burst (union of the time-slice burst and the IFEC burst) and the time-slice burst **may** not coincide in time due to the exclusion of IFEC sections from time-slice burst definition while, at the same time, their Delta-t signalling points to the same temporal structure, the time-slice burst.

Some examples are given below to clarify, in the case **when no MPE-FEC is used in the elementary stream** (please not that simultaneous used of MPE-FEC and MPE-IFEC is not allowed and is for further study).

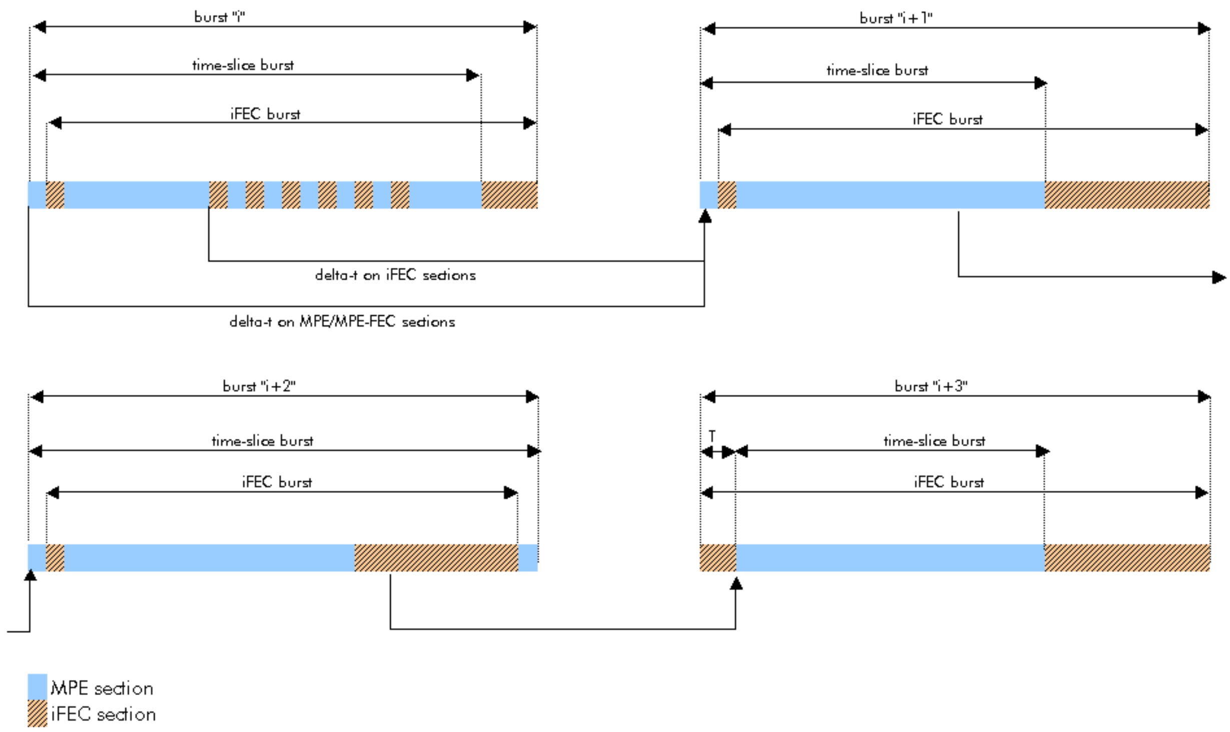


Figure 6.14: Burst sending arrangement (w/o MPE-FEC)

In figure 6.14, we can see three successive bursts of the same elementary stream (bursts i to i+2):

- Delta-t signalling throughout the same burst always points to the first MPE section of the following burst, and not the first MPE-IFEC section, because Delta-t has the same definition, whatever the section, and always refers to the time-slice burst;

- as a consequence of this signalling, if the burst begins with MPE-IFEC sections, they may be missed, unless the delay T between their sending and the actual sending of the time-slice burst is small compared to the Delta-t unit (10 ms).

- the time-slice burst and the burst do not necessarily coincide in time. Many configurations are possible and, in this example, the only burst i+3 sees the matching between both.

Recommendations:

Since the burst can differ from the time-slice burst but the Delta-t signalling is in any case related to the time-slice burst, the following recommendations are given **for cases when no MPE-FEC decoding applies**:

- At least one MPE section **must** be positioned in the vicinity of each burst beginning so that no MPE-IFEC sections are missed. The vicinity is considered such that the delay between the burst and the time-slice burst beginnings is below 1/10th of the Delta-t unit so 1 ms. Depending on the transmission bit rate, this time may allow to send an MPE-IFEC section or not. For instance with a bit rate of 2,4 Mbits, the time it takes to send 1 TS is 0,623 ms which allows sending 1 full MPE-IFEC section in the only case of 256 rows. For simplification, the first section **should** therefore be an MPE section.

- Since the only MPE-IFEC section conveys burst numbering information, at least one MPE-IFEC section **should** be positioned close to the beginning of the burst, such that the receiver may detect the burst number as early as possible while receiving the burst, and position immediately the received MPE sections in their correct ADT, without waiting for reception of late MPE-IFEC sections. Note that this recommendation is not a hard one since it is always possible to receive MPE sections without previously knowing their burst number: these sections are stored until their burst number can be determined by the receiver, by receiving a later MPE-IFEC section conveying this burst number or other means, for instance based on timing.

- After the beginning of the burst, the MPE and MPE-IFEC **may** be freely interleaved during a time interval not exceeding max_burst_duration signalled in time_slice_fec_identifier. After max_burst_duration, no more sections (MPE, MPE-IFEC) **must** be present. The receiver **must** ignore these sections received lately and can switch off the receiver, hence enabling power saving.

- To detect the end of the burst, the receiver **must** check that the end of the time-slice burst and IFEC burst are reached. For that purpose, the receiver checks the received sections flags and looks whether one of the conditions presented in table 6.7 are respected for all lines. These conditions are the results of time-slice burst sending arrangement rules found in EN 301 192 [9], clause 9.10 (*"All sections carrying Application data datagrams of a given MPE-FEC Frame shall be transmitted prior to the first section carrying RS-data of the MPE-FEC Frame (i.e. sections carrying Application data datagrams shall not be interleaved with sections carrying RS-data within a single MPE-FEC frame).[…] Within an elementary stream, sections delivering data of different MPE-FEC Frames shall not be interleaved. […] Note that for each MPE-FEC Frame, MPE sections are delivered before MPE-FEC sections"*) and MPE-IFEC time-slice burst generation and sending arrangement rules found in MPE-IFEC specifications [MPE-IFEC], clause 3.6.

Table 6.7: Conditions for end of IFEC burst detection

<table>
<tr><th></th><th></th><th>Condition 1</th><th>Condition 2</th><th>Condition 3</th><th>Condition 4</th><th>Condition 5</th></tr>
<tr><td rowspan="2">Time-slice burst</td><td>MPE</td><td>An MPE section with a table_boundary set to 1 has been received.</td><td>An MPE-FEC section has been received.</td><td>An MPE-IFEC section has been received with an MPE_boundary set to 1.</td><td rowspan="3">Current time has exceeded previous Delta-t plus max_burst_duration.</td><td rowspan="3">Another service PID has been found, signalling change in the service reception.</td></tr>
<tr><td>MPE-FEC (this is for information only since MPE-FEC cannot be present at the same time as MPE-IFEC in the ES)</td><td>An MPE or MPE-FEC section has been received with the frame_boundary set to 1.</td><td>-</td><td>An MPE-IFEC section has been received with an MPE_boundary set to 1.</td></tr>
<tr><td>IFEC burst</td><td>MPE-IFEC</td><td>An MPE-IFEC section with frame_boundary set to 1 has been received.</td><td>-</td><td>-</td></tr>
</table>

This table can be used as follows on a particular case where the IFEC burst is made of a first MPE section, then one IFEC section, then all remaining MPE sections, then all remaining IFEC sections.

- In case no section at all is lost:
  - Time-slice burst end will be known with the last MPE section (condition 1).
  - IFEC burst end will be known with the last IFEC section (condition 1)
- In case some MPE sections, including the last, and some IFEC section, but not all, are lost:
  - Time-slice burst end will be known with the first MPE-IFEC section received after the last MPE section (condition 3).
  - IFEC burst end will be known with the last IFEC section (condition 1).
- In case some MPE-IFEC sections, including the last, are lost:
  - Time-slice burst end will be known with the last MPE section (condition 1).
  - IFEC burst end will be known with the time out (condition 4).

- Note that one consequence of these rules is the need for the receiver to potentially stay on after last MPE section has been received (to receive MPE-IFEC sections if any) or after last MPE-IFEC section has been received (to receive MPE and MPE-FEC sections if any). This specific behaviour must be applied on a per stream basis: for those streams where no IFEC protection is applied, the same rules as DVB-H will apply (when the last MPE or MPE-FEC frame has been received, the receiver will not have to wait for other MPE-IFEC sections and so will be able to switch off immediately).

6.2.4 Parameters selection

6.2.4.1 Introduction

This clause provides explanations and recommendations for MPE-IFEC parameter selection. To configure the MPE-IFEC, the operator needs to fix the 3 main parameters of the MPE-IFEC framework, namely B, S, D and the code rate: D is selected based on heuristics and the impact on zapping time, B and S are selected based on latency (B+S) and the code rate is selected on performance criteria. All performance results are measures at interface R1 shown on with a D equal either to 0 or B+S.

The reader will see how these parameters can be easily established depending on the sought performance and the channel behaviour. For that purpose:

- we first discuss the D parameter and its dependence with other B and S in clause 6.2.4.2; we show that D can be selected between only two typical values;
- we then discuss how B and S can be derived in clause 6.2.4.3 from link layer latency and code rate;
- we finally show how link layer latency and code rate can be selected based on simulation abacuses in clause 6.2.4.4;
- we also give a possible global procedure in clause 6.2.4.5.

We introduce the following latency definitions:

- "encoder latency/delay": time between reception, at the encoder side, of an IP packet and actual emission of the last (data or IFEC) section related to this IP packet; the data section is the one carrying the IP packet whereas the IFEC section has correction capability on the ADT column(s) where the IP packet is located; the encoder latency can be computed over a set of IP packets belonging to one burst; we then talk of "burst encoder latency":

$$\text{encoder\_latency}(\text{IP\_packet}) = \text{last}\begin{pmatrix}\text{MPE section sending time}\\ \text{MPE - IFEC section sending time}\end{pmatrix} - \text{IP\_packet arrival time}$$

$$\text{encoder\_latency}(\text{burst}) = \max(\text{encoder\_latency}(\text{IP\_packet})\ \text{for all IP\_packets belonging to the burst})$$

- "receiver latency/delay": this is the time between delivery of an IP packet at the decoder side and reception of the first information on this IP packet, whether it is the MPE data section carrying this IP packet or an IFEC section providing parity for an ADT column that includes data bytes for this IP packet. The receiver latency can be computed on a complete burst, we then talk on the burst receiver latency.

$$\text{receiver\_latency}(\text{IP\_packet}) = \text{IP\_packet delivery time} - \text{first}\begin{pmatrix}\text{MPE section arrival time}\\ \text{MPE - IFEC section related to an ADT column}\\ \text{that includes data bytes for this IP packet arrival time}\end{pmatrix}$$

$$\text{receiver\_latency}(\text{burst}) = \max(\text{receiver\_latency}(\text{IP\_packet})\ \text{for all IP\_packets included in the burst})$$

- "end-to-end latency/delay": time between delivery of an IP packet at the decoder side and reception of the IP packet at the encoder site. The end-to-end delay is equal to the encoder latency, the transmission delay and the receiver latency. So end-to-end delay varies only due to the receiver latency since encoder latency and transmission delays are considered as fixed. In steady state, the end-to-end delay is equal to its maximal value, but during non steady states (like zapping) this end-to-end delay may be shortened due to the possible reduction of receiver latency.

6.2.4.2 Recommendations on D

B,S and D are related through the function giving $M = B + \max(0, S - D) + \max(0, D - B)$. This M, in addition to giving the number of encoding matrices, gives the receiver latency and the memory requirements:

- Assuming the simplification that burst are regularly repeated with a temporal period of burst_repetition_interval, the receiver latency at the jitter free interface is equal to $\mathrm{burst\_repetition\_interval} * M$. The jitter-free interface is the R1 interface when late decoding is used.

- Assuming that ADT and FDT sizes are given by ADT_size and FDT_size, the memory required for complete IFEC decoding is equal to $(\mathrm{ADT\_size} + \mathrm{FDT\_size}) * M$.

These formula show the importance of the parameter D and its impact on user perception (via the latency) and terminal sizing (via memory), all other things being equal. For instance for a B equal to 6 and a S equal to 4, we obtain the curve presented in the figure 6.15.

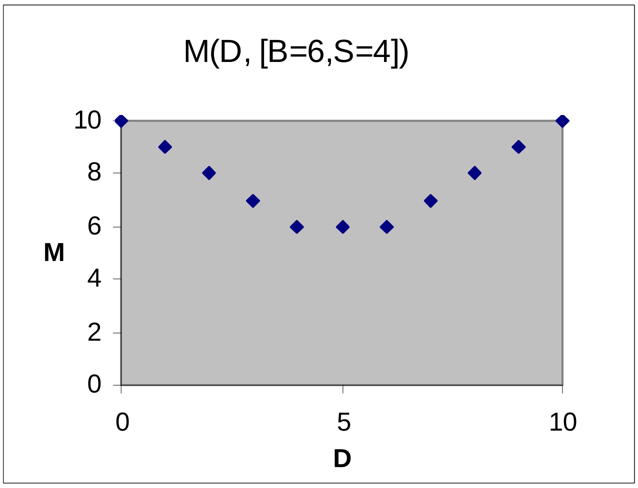


Figure 6.15: M(D, [B=6, S=4])

It can be seen that:

- when D is equal to 0, M is rather large and equal to B+S=10; this configuration corresponds to the situation when the MPE-IFEC sections are always sent after the data they protect, which is good for error correction performance. But the high M value implies long jitter-free receiver interface delay and large memory requirements;

- a minimum value of M is obtained for D taken in the range [4; 6]. Obviously this enables to reduce memory requirements in receiver and latency since the receiver needs to wait less for MPE-IFEC section reception at the jitter-free interface compared to when the sections are not sent with the data they protect. However the MPE-IFEC sections will be mixed with the datagram bursts, which reduces performance because an error affecting a burst likely affects at the same time, the data it conveys, and the FEC that corrects the errors. Increasing D will also increase end-to-end latency;

- after the value 6, M increases again. Data and MPE-IFEC are still mixed, up to the situation when D=B+S (10 in our case). In this case, the data and IFEC sections are again not mixed, which provides good performance results. The receiver jitter-free latency is the same as D=0 but since the IFEC always precedes the data, this enables "early decoding" during zapping (see clause 6.2.5.2).

The influence of the D parameter on different criteria is presented in table 6.8.

Table 6.8: Influence of the selection of D parameter on different criteria

D value	Performance	Receiver latency	Encoder latency	Receiver memory
D=0	+	-	+	0
D=D_{min_sizing}	-	0	0	+
D=B+S	+	+	-	0

NOTE 1: D_{min_sizing} is the lowest D that minimizes the M function. For instance, on our previous example, D_{min_sizing} = 4.
NOTE 2: "+" means the criteria is matched with a good performance, 0 means a neutral performance and "–" means a worse performance.
NOTE 3: Performance is the correction capability at the jitter free interface; receiver latency includes both jitter-free and non jitter-free (during zapping); all other criteria are classical.
NOTE 4: D=B+S provides a good compromise in performance, including receiver latency during zapping, at the detriment of encoder latency. This configuration should be used for contents not delay constrained whereas D=0 should be kept for delay constrained contents, at the detriment of receiver latency and zapping time.

6.2.4.3 Recommendation on B and S

The code rate is defined on a per encoding matrix basis by the formula combining, for a given encoding matrix m, the number of ADT data columns (adt_data_columns(m)) and the number of FDT FEC columns (nof_fdt_fec_columns(m)). The target code rate is defined in clause 6.2.3.8. We also assume, as in this clause, that the same target code rate applies to all encoding matrices.

Code_rate can be selected with complete freedom within the range authorized by C (using padding) and R (using puncturing). If we take our current example, C is fixed to 40 but the number of data columns is on average 37. Depending on the number of FEC columns (from 0 to 64), 64 code rates from 1 to 0,36 are possible as presented in clause 6.2.3.8.

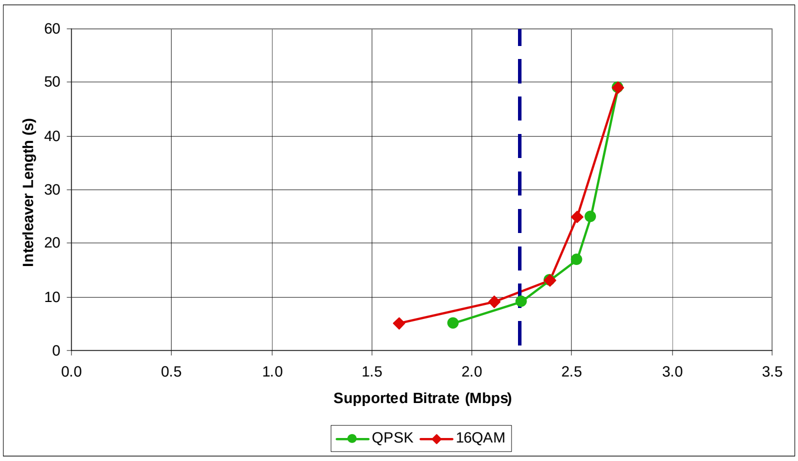
However, an exhaustive simulation campaign determined that the configuration offering the best performance $(B,S,code\_rate)_{actual}$ follows the rule $\left[\frac{B}{B+S}\right]_{actual} \approx code\_rate_{actual}$. Based on this assumption, near optimal parameters B_{opt} and S_{opt} can be derived from target FEC rate and B+S for cases where D=0 by using the following formulas:

$$S_{opt} = \mathrm{ceil}\left(\left(1 - code\_rate\right) * \left(B_{actual} + S_{actual}\right)\right)$$
$$B_{opt} = \left(B + S\right) - \mathrm{ceil}\left(\left(1 - code\_rate\right) * \left(B_{actual} + S_{actual}\right)\right)$$

Since B and S can be easily derived from the couple (B+S), which is equal to M when D=0 and noted M(0,B,S), and the code rate, we will use these two parameters in the following.

6.2.4.4 Selection of M and code rate

Simulation campaigns have enabled to derive the rule between M(0,B,S) and the code rate for a given ESR(5) fulfilment, on a per channel basis. These complete results are presented in the clause A.12. We take the example of a suburban channel with figure 6.16.



NOTE: M(0,B,D)=B+S for D=0 and D=B+S; the curve are valid for these two values of D.

Figure 6.16: SH-A, Class 1 - QPSK 1/2 and 16QAM ¼ - LMS-SU - 50 kmph - 63 dBW EIRP Satellite

Figure 6.16 can be interpreted as follows:

- all points on the curve match an ESR(5) fulfilment of 90 %;
- all points above have a better quality;
- all points below have a lower quality.

The curve is then an iso-ESR(5) one.

The iso-ESR(5) curve has always the same shape with a slope that is increasing as code rate increases, an inflexion point and then an asymptotic point. In the example given, the following points can be found:

- first value is around 1,7 Mbps/5 s (interleaver length directly related to M);
- the inflexion point is at 2,55 Mbps/25 s;
- the asymptotic value is at 2,75 Mbps/50 s.

Two different approaches can be followed for determining which parameters must be selected for a particular system:

- the terminal is memory constrained (compared to the asymptotic value) to a certain M(0,B,S), for instance 25 s: the selected code rate is the code rate found at the intersection of the iso-ESR(5) curve and the horizontal line at the target M(0,B,S); in our case the resulting useful bit rate is 2,5 Mbps, which turns into a code rate of 2/3 because the total bit rate in 5 MHz QPSK1/2 GI=1/4 is 3,357 Mbps. These bit rates are measured at the R2 interface at MPEG2 TS layer;
- the terminal is not memory constrained (compared to the asymptotic value), the objective is to maximize the capacity and so position at the inflexion point of the curve. In our case, the target capacity could be 2,7 Mbps (code rate 4/5); the M(0,B,S) is found at the intersection of the curve and the vertical line at target code rate. In our example the value is 50 s.

A mobile system is also designed to work in a set of different propagation channels. The selection of the link layer parameters must be done for the worst case, typically LMS-ITS. Once selected, the configuration should perform well under less challenging channels. We give the complete example below.

We start with an LMS-ITS dump excerpted from clause A.12.

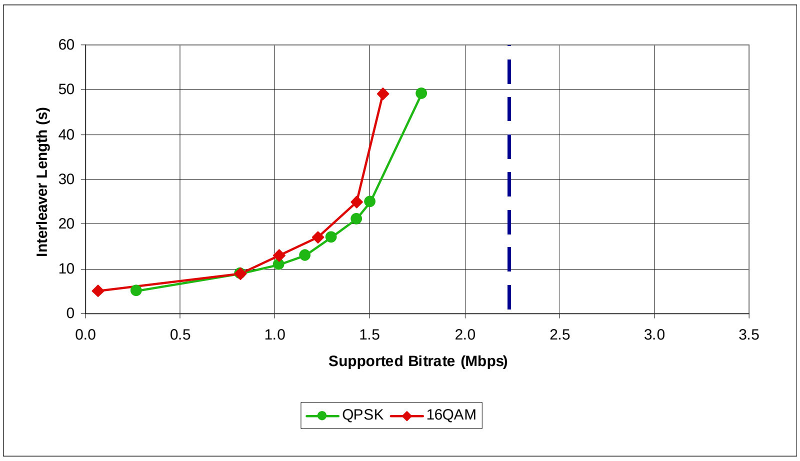


Figure 6.17: SH-A, Class 1: - 16QAM 1/4 and QPSK 1/2 - LMS-ITS - 50 kmph - 63 dBW EIRP Satellite

If we are with a M(0,B,S) equal to 25, this gives a code rate of 0,44. We can see immediately on figure 24 that the performance will be much higher than the target in LMS-SUB (we are above the curve). Finally, since we are in the hybrid frequency in that case, we can also check the performance under TU6 channel: in clause 6.2.5.1.4 we show that a configuration for LMS-ITS or LMS-SU performs perfectly under TU6 channels.

6.2.4.5 Parameter overall selection logic

So the general parameter selection logic is presented in figure 6.18.

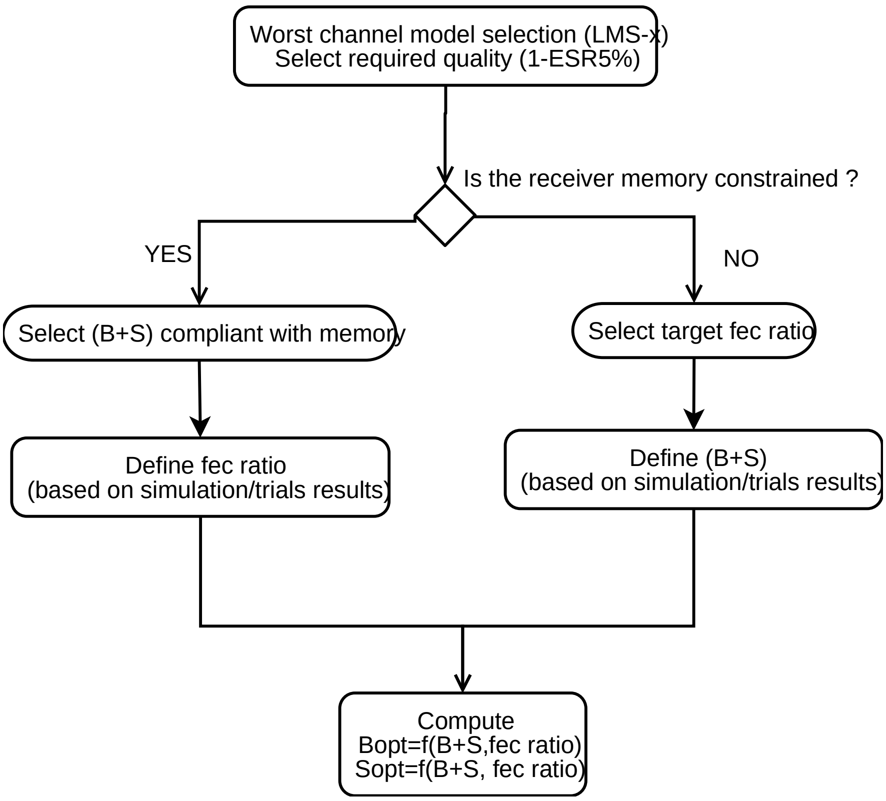


Figure 6.18: link layer parameter selection logic

This selection logic is valid for D values such that MPE-IFEC sections are not transported in the MPE-IFEC same time-slice burst as the MPE they are protecting, so for D=0 and D≥B+S. For intermediate values, the performance will be slightly less requiring higher B+S and/or lower code rate and is for further study.

This procedure is used to recommend parameters in clause 6.2.5.1.3.

6.2.5 Simulated performance

Link layer simulation configuration elements can be found in clause A.6. All detailed simulations results can be found in clause A.12. We give in this clause a summary of the clause A.12 results.

6.2.5.1 ESR(5) performance

This clause excerpts some results from clause A.12. The logic for these results is the following:

- targeted quality:
  - for LMS channels, ESR(5) fulfilment of 90 % is sought;
  - for terrestrial (TU6) channels, ESR(5) fulfilment of 99 %, equivalent to a FER of 1 % is sought;
- memory constraints:
  - depending on terminal category (handheld or vehicular), the memory constraints are different; vehicular terminal can be considered to have "unlimited memory" whereas handheld receiver can be considered to have memory limitations;
- code rate:
  - the target code rate is the one that enables to support similar efficiency as the class 2 receivers (full protection at physical layer); for instance, a QPSK1/2 and 16QAM1/4 will have a target link layer code rate of 2/3;
  - if the quality criteria for a maximum B+S value can not be fulfilled, then the code rate is reduced until the quality is achieved.

The objective of the simulation procedure is to find the ((B+S);code rate) combination that enables to meet expected quality of service under the double constraint of (B+S) stays below maximum memory and code rate does not drop below 30 % of capacity loss. This exercise is performed for different channel environments and receiver categories.

6.2.5.1.1 LMS-ITS 50 kmph

Please refer to annex A.12 for details. There is always an asymptotic value for the interleaving duration, around 25 s to 30 s: increasing the duration beyond this value does not bring any help. The recommended configuration for the link layer is therefore given in table 6.9.

Table 6.9: Recommended link layer configuration in ITS (50 kmph)

Hybrid Frequency modulation	C/N used for simulation	Recommended B+S	Recommended Code rate
SH-A (QPSK1/2 and 16QAM1/4 16QAM2/7 GI=1/4)	11,2 dB	25	½
SH-B (TDM 8PSK1/3, QPSK1/2)	12,3 dB	25	2/3

Additional configurations are for further study, in particular sensitivity of this results as a function of C/N.

6.2.5.1.2 LMS-SU

Please refer to clause A.12 for details. An interleaver of 25 s gives the asymptotic performance. Minimum recommended configuration is given in table 6.10.

Table 6.10: Minimum configuration in SU (50 kmph)

Hybrid Frequency modulation	C/N	B+S	Code rate
All	> 8	25	¾
SH-A (QPSK1/2, 16QAM ¼)	< 8	25	2/3

6.2.5.1.3 Recommended class 1 configuration

Based on previous results and procedure explained in clause 6.2.4, the following class 1 configurations are proposed.

Table 6.11: Recommended class 1 configuration

Hybrid Frequency modulation	C/N	B+S	Code rate
SH-A	< 12	25	1/2
SH-B	> 11	25	2/3

For services which require lower end-to-end latency, lower values of B+S may be used. The appropriate configuration can be derived according to the above procedure (in clause 6.2.4) and using simulation results in clause A.12.

The parameters can be freely chosen provided the required memory, as computed in clause 6.2.6, does not exceed the class 1 memory capability as given in clause 10.

Note that due to incomplete simulation coverage, this table will be amended for the second release.

6.2.5.1.4 TU6 for hybrid frequency

This clause refers only to the SFN case: having selected a class 1 configuration, we know its performance under LMS-ITS and LMS-SU channels. We need now to check also the performance over the terrestrial TU6 channel since the *same* MPE-IFEC configuration is also used over the terrestrial repeaters because of the SFN nature of the network: will the user experience different quality while performing a hand over between the hybrid and non-hybrid channels under the same CGC network (the hybrid frequencies do have an IFEC protection, which is not supposed to be the case of a non-hybrid frequency)?

To evaluate the performance difference, we assume that CGC is working at C/N that enables ESR(5) fulfilment of 99 % without MPE-IFEC and we need to check at this C/N if the hybrid signal, having additional MPE-IFEC protection but lower physical layer code rate, will have lower, same or better quality than non MPE-IFEC protected ones.

As an example, excerpted from annex A12, we use a terrestrial configuration in QPSK1/3 that has an error-free C/N_{NHF} at 3 kmph of 3,5 dB and 1,5 dB at 50 kmph. We now check the performance of the hybrid frequency (still at 3 kmph/50 kmph respectively) but with lower physical code rate (1/2) and the selected link layer configuration (e.g. 2/3), for a 25 s interleaver.

It can be checked in annex A12 that the MPE-IFEC hybrid frequency will need 1 dB more than equivalent physical layer configuration at 3 kmph (the required C/N is 4,5 dB) and 2 dB at 50 kmph. So the channel in hybrid frequency will require a slightly higher C/N than the equivalent one of the non hybrid frequency to ensure same quality as the non hybrid one. The MPE-IFEC compensate partly the higher physical code rate.

Note also that, due to the SFN between satellite and CGC, this added C/N can be compensated in areas where the CGC coverage starts to be less efficiently received. Such SFN gain is introduced in clause 11.

6.2.5.1.5 TU6 for non-hybrid frequency

We check the interest of the MPE-IFEC for non-hybrid channels (content that is not broadcasted over the satellite, only sent over the terrestrial network). For this purpose, we display the useful bit rate as a function of the C/N in clause A.12 and see with MPE-IFEC how dBs can be traded for bandwidth.

This figure shows that a network operator, we can trade capacity with dB:

- in QPSK1/3:
  - 20 % capacity with 1 dB of C/N.
- in QPSK1/2:
  - 15 % capacity with 1 dB of C/N (for comparison, with MPE-FEC we would need 33 % for 1 dB gain);
  - 30 % of capacity with 2 dB.
- in 16QAM1/5:
  - 13 % capacity with 1 dB of C/N;
  - 25 % of capacity with 2 dB of C/N.
- in 16QAM1/4:
  - 25 % capacity with 1 dB of C/N.

Generally speaking, the short physical interleaver performs well in TU6 environment and MPE-IFEC enables to provide intermediate C/N operational points in between the values, in a similar way as, but better than, MPE-FEC.

6.2.5.2 Zapping time performance

6.2.5.2.1 Introduction

This clause deals with zapping performance for only class 1 terminals. For a more general zapping performance discussion, including class 2 please refer to clause 6.3.2. However, some definitions are given or recalled that are also applicable for both classes. After these definitions, we explain the concept of zapping in an MPE-IFEC context and then give a complete example. These are conceptual considerations, zapping simulation results is for further study.

6.2.5.2.2 Definitions

The definitions are either new ones or precision on already given definitions in the light of the newly defined.

Précised definitions:

Late decoding: refers to the techniques used for decoding with maximum protection. The receiver latency is always maximal, the IP packet being delivered to the video decoder only when the maximum protection is achieved (received). Late decoding time implies the end-to-end delay is always maximized whatever the situation because the receiver latency is maximized. Late decoding is explained for class 2 in clause 7.3.3.5.4 and for class 1 in clause A.6. Only class 2 can afford late decoding and fast zapping.

Early decoding: refers to the set of techniques used for decoding with less protection than the maximum one. The receiver latency may be reduced compared to the late decoding case, the IP packet being delivered only when the requested parity has been received and not all the parity. This enable to lower end-to-end delay and accelerate IP packet delivery in case of good reception condition but this induces variable end-to-end delay when more parity is requested during fading events. Early decoding for class 2 is explained in clause 7.3.3.5.5 and for class 1 in clause A.6. Early decoding is requested for class 1 for supporting fast zapping.

New definitions:

Zapping instant: time when the user selects a new program.

Zapping time/delay: delay between the zapping instant and the time this program is actually received by the video decoder (we take into account only the link layer latency); by definition zapping time is mostly equal to the receiver latency. Zapping time can depend on many parameters like air interface configuration (physical and link layer interleaver: do we use a uniform/late profile, is D>0, etc.) but also on local reception condition (is reception quality good?) and receiver strategy (does the receiver supports fast zapping?). In lossless reception situation, zapping time can be as short as a burst reception for terminals applying fast zapping (class 1 with D>0 and class 2 with a uniform late profile), or as long as encoder latency for terminals without any fast zapping support (class 2 with uniform long profile). In bad reception condition, zapping time can exceed these values and last as long as the signal blockage.

Fast zapping: techniques used for reducing zapping time to acceptable delay even in the presence of long interleavers, be they at physical or link layer. They generally involve specific waveform configuration (uniform late for class 2, D>0 for class 1), terminal specific behaviour (late decoding for class 2 and early decoding plus parity recovery for class 1). Not taking into account other delays in the system (video decoding, etc.), the fast zapping time can be as low as the time to receive a single burst.

- In class 1 context, early decoding is usually used for fast zapping: under good reception condition, it is always possible to display content immediately since datagram bursts are transmitted as usual. The receiver can watch the video without waiting for late decoding latency, but may not have all the parity information to sustain a potential loss. Air interface configuration (use of D parameter) and parity recovery techniques can help the receiver to recover the parity progressively so that after a certain period, the receiver has recovered same full protection as in late decoding.
- In class 2 context, fast zapping refers to the interleaver uniform/late configuration in conjunction with late decoding that enables displaying first image immediately under good reception condition; parity is progressively recovered and after the encoder latency, full protection is ensured. This technique is described in clause 7.3.3.5.

Parity recovery: technique used by a receiver applying early decoding to progressively recover the same level of protection as late decoding one. For the link layer, different techniques are presented in clause 6.2.5.2. Note that parity recovery is not requested for the uniform/late physical interleaver since fast zapping is compatible with the late decoding as explained in clause 7.3.3.5.4.

Principles for zapping time analysis in class 1

The typical following scenario happens:

- the end user zaps by selecting one PID on a list, the PID being chosen via the ESG;
- the receiver decodes the MPEG2 TS and looks for corresponding PID;
- once it has found the corresponding PID, it starts receiving MPE and MPE-IFEC sections;
- with the included Delta-t, it can delineates the time-slice bursts and achieve power saving;
- it stores the received sections in their respective bursts, maps them on their respective ADTs and iFDTs so that the Encoding Matrices are progressively "populated";
- after a certain delay, the receiver can decode and deliver IP packets to the video player.

The zapping delay depends on this delay and different receiver behaviours are possible:

- late decoding:
  - the receiver will wait for all the parity before decoding;
  - the waiting delay will depend on the sending arrangement:
    - it will be equal to M when D=0;
    - it will lower to a minimum value when D equals D_{min_sizing};
    - for larger D, the delay will be kept to this value;

- for instance in the example case (B=6, S=4), the initial delay is 10 when D=0 and 6 when D=10. Thanks to this delay, all the parity coming from the data part can be processed and, due to the fact that the FEC parity precedes the data, the FEC is also available;
- however, with late decoding, fast zapping is then not possible (we need to wait for D_{min_sizing}): we need to resort to early decoding;

- early decoding:
  - the receiver does not wait for all parity before decoding; the receiver can start decoding before parity has been received, for instance as soon as it has received a burst. Obviously, this has a cost since the protection may not be enough to protect against normal impairments;
  - as bursts are received, if D is large enough, since FEC is sent before data, FEC will be accumulated to provide some protection in conjunction with previously received burst (for FEC computation, current burst is interleaved with B-1 previous bursts). However, full protection is not achievable since the protection of the currently received burst depends also on following B-1 bursts (the interleaving is a convolutional one);
  - for achieving normal protection, the receiver must perform a late decoding and, for this, has to wait for same delay as in late decoding, for instance 5 burst in addition of the current one;
  - this additional waiting time can be "spread" over time thanks to recovery techniques and/or split with an initial small additional buffering. During this "spreading time", the end-to-end delay will not be constant and will increase from the initial value to the maximum one. So the interface is not jitter-free at the beginning. One technique for achieving this spreading is presented in [i.33]: the data from the link layer is delivered with a reduced rate to the IP layer. A media decoder may be informed to slow down the media playout. More precisely, after zapping, the data signal partially corrected by the MPE-IFEC with a slightly slowed rate than the transmitted rate is played out; the receiver corrects the signal by the MPE-IFEC, and, after a certain period, inversely proportional to the slowing rate, switches to the fully protected signal, at the transmitted rate. [i.34] suggests that 20 % of speed reduction is hardly perceivable by the user, allowing full protection and display at the transmission rate after 25 s, for a B equal to 6 (we need to recover B-1 bursts in addition of the initial one, so 5 bursts and, with a repetition interval of 1 burst per second, this leads to 5 s recovered in 5/0,2 = 25 s). Such techniques may require modification to existing media decoders, and especially audio ones;
  - it is also important to mention what happens if the receiver detects that during the slow down, the FEC is not sufficient. Then an immediate rebuffering may be reasonable.

To summarize, correctly decoding with a long interleaving imposes a buffering delay that is a function of D and can be lowered to values typical of B-1 in addition of the current burst reception. However in many cases, the burst received at zapping instant is correct and can be forwarded to the media player for immediate decoding and display. This action is referred to a "fast zapping". If the terminals plays the signal at normal speed, the IFEC decoding will never benefit from the B-1 datagram bursts that are also used to create the parity. So techniques for spreading the delay are considered.

EXAMPLE: We give hereunder a first analysis of these techniques, assuming only MPE-IFEC is used and excluding therefore MPE-FEC. If we consider an encoding depth of B, a FEC spreading of S and a given code rate, we have the following theoretical characteristics during the zapping time on MPE-IFEC time slice burst k:

- at zapping instant, if D=0 or D>B+S, the MPE-IFEC time slice burst k is received without any protection; typical probability of having an erroneous burst before MPE-IFEC decoding is in the order of 20 % to 30 % in worst LMS-intermediate-tree-shadowed cases but only 8 % in LMS-suburban;
- for subsequent MPE-IFEC time slice bursts, more protection is received that enables better recovery as, in average, more data and FEC for the corresponding encoding matrices is received through early decoding (decoding without having received full parity, either FEC or DATA):
  - if D≥B+S, between zapping instant (MPE-IFEC time slice burst k) and MPE-IFEC time slice burst k+B+S-1, useful MPE-IFEC sections are progressively received up to full reception when MPE-IFEC time slice burst k+(B+S-1) is received);

- between zapping instant (burst k) and burst k+B-1, MPE sections are received that progressively fill the ADTs up to a state of partial completion at burst k+B-1 that then stagnates at this level: even if we wait indefinitely, the ADTs will never be completely filled, which could be considered as a loss, because the missing columns are coming from datagram burst to be received after currently received datagram burst. The level of achievable completion in ADT depends on the initial buffering time D_{Buf}: the more the initial time, the better;

- after a number of burst, the received FEC columns can compensate for the missing ADT columns and can even provide some additional protection for "real" losses. In a typical case (B=6, S=4), during 4 bursts, no error can be corrected but starting with the 6^{th} burst, the correction capacity starts to be positive and a complete burst loss is can be sustained at the 7^{th} position;

- the number of artificially missing columns due to non completion of the data part remains but decreases quickly when the initial buffering increases. So it can be interesting to increase slightly the initial buffering time to ensure that a good level of protection is ensured. In any case, a complementary buffering time is needed to reach full protection (40 columns in each ADT in average).

So during B+S bursts, FEC protection increases regularly but does not reach maximal capacity unless some additional buffering of B-1 is activated.

- the initial buffering before starting to deliver IP packets, as counted in number of bursts, Buf_{init}, can be increased (Buf_{init}>1), this has a good impact on protection of the stream during zapping time;

- complementary buffering can be done by slowing the playout rate by a slowdown_rate factor: the slowdown_rate is the percentage of speed reduction at the output of the IFEC decoder (TR 102 377 [i.21] suggests that 20 % can be supported);

- total duration to recover the data parity (missing datagram burst to fill completely the ADTs) is then ($B-Buf_{init}$)/slow_down_rate.

All these delays have been represented in the following figure as a function of (B+S) and taking for initial buffering delay Dbuf a value of 4 (systematic delay of 4 burst). For instance for (B+S)=10, we recover the initial data after 6 bursts, then the FEC and the final data are completely recovered after 10 bursts. So we can display first image 4 bursts after zapping and be completely protected after 10 burst. Another example for (B+S)=20: after 12 bursts, we have, thanks to the initial delay of 4 bursts, already a good protection over the data. We must receive 20 bursts to benefit from full FEC parity and the final protection is achieved after 45 bursts. Please note the these delays are not additive: they are time thresholds.

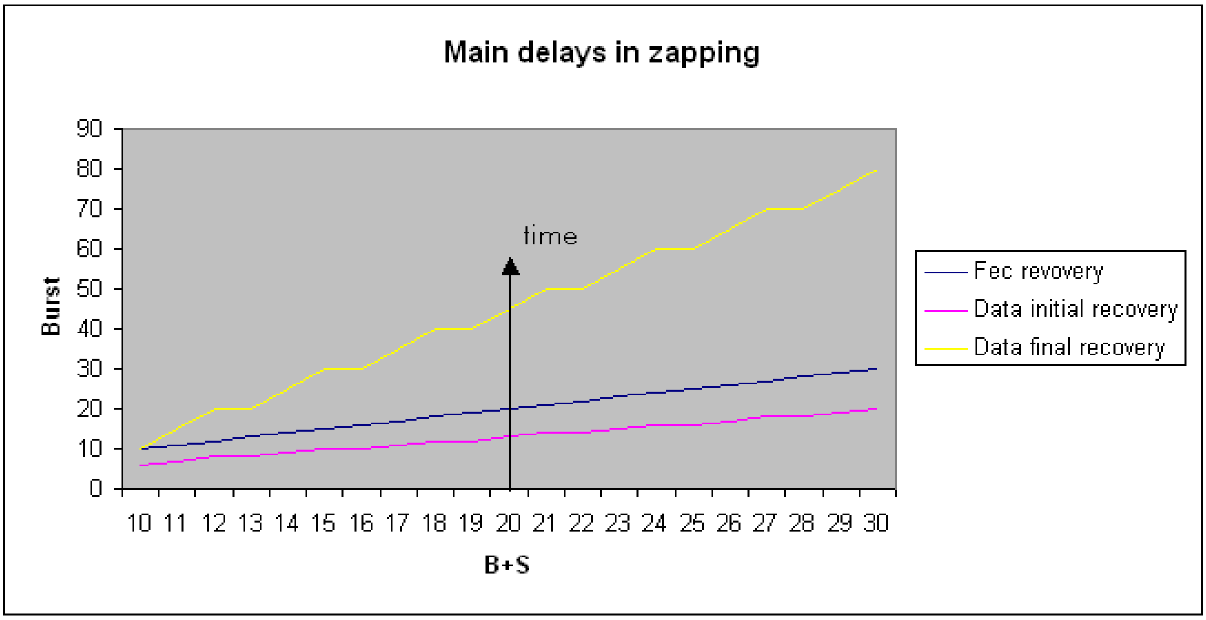


Figure 6.19: main zapping delays

Zapping performance with measured quality during the zapping period is for further study.

6.2.6 Memory requirements

6.2.6.1 Introduction

This clause gives the memory requirements for the receiver. We first explain in clause 6.2.6.2 how the MPE-IFEC memory can be sized in a class 1 terminal by defining a "memory sizing function". We then introduce how this memory can be used in a typical implementation using the bijection between the ADST and ADT introduced in clause 6.2.3.5, distinguishing between constant bit rate and variable bit rate IP flows. We describe an implementation applying for CBR. VBR can always be supported with this implementation but the memory may be significantly higher. Memory management optimization are possible that enable keeping the memory as low as for CBR traffic but this is for further study.

6.2.6.2 Memory sizing function

Hypothesis 1: we consider that ADST columns are stored in ADT via the adt_index and adt_colums functions (see clause 6.2.3.5). As a consequence, there is no need for budgeting memory for ADST and memory is essentially budgeted for Encoding Matrices, ADTs and FDTs.

NOTE 1: One could say that at least 1 ADST buffer could be envisaged for absorbing current burst for which burst number is unknown and which cannot be mapped on ADT/FDT. This is true but, as the reader will see in the rest of the clause, optimizations could probably be possible to store this burst in ADT/FDT columns not already used without affecting quality of the decoding. So the conclusion remains the same, the budget is considered for the only ADT and FDT.

Since we budget only ADT and FDT, memory requirements are directly a function of the number of encoding matrices, which is a function of parameters B,S and D via the M function, and of ADT and FDT individual sizes:

$$\text{memory} = (\text{ADT\_size} + \text{FDT\_size}) * \text{M}(\text{D},\text{B},\text{S}) \text{ where } \text{M}(\text{D},\text{B},\text{S}) = \text{B} + \max(0,\text{S}-\text{D}) + \max(0,\text{D}-\text{B}).$$

Hypothesis 2: we assume that memory is the sum of all ADT and FDT sizes counted by their matrix weight in bytes and zeroing any pointer structure (that will be used in the document below) or ancillary information bits. This approach will give precise order of magnitude but actual implementation may require more memory capacity.

ADT and FDT memory is sized by their number of rows, T, and their number of columns, C for the ADST and ADT and R for the FDT. Since different optimizations are possible, for the parameter C we need to precise between ADT and ADST (C_{adst} and C_{adt}).

Parameters list: the list of parameters required for sizing the memory is given below:

- T is usually selected by the user but minimum values are fixed by the data volume per datagram burst since the maximum number of columns C is fixed to 255. T must be selected amongst {256, 512, 768 and 1 024} but T must be greater than ceil(datagram_burst_size/255). A default working value for T is 1 024.

- C is a function of the traffic profile and its instantaneous variations datagram burst by datagram burst: variable datagram burst will occupy an ADST that must be sized to accommodate its peak size:

  $\forall k \geq 0, \text{T} * \text{C} \geq \text{datagram\_burst\_size}(k) + 1$ => this sizes the C_{adst} parameter.

- Normally C_{adst} should be used for ADT sizing so that $C_{adt} = C_{adst}$. This is the natural option suggested by the specification. But terminal memory management optimizations can enable differentiation between both. These implementations are for further study. So in the following $C=C_{adt}=C_{adst}$.

- When the code rate is fixed, it is a direct function of C_{adt} => this sizes the R parameter such that:

  $$\text{R} = C_{adt} * \frac{1-\text{code\_rate}}{\text{code\_rate}}$$ according to clause 6.2.3.8.

Memory sizing equation: These C_{adt}, T and R parameters enable to derive memory requirements:

$$Memory = \frac{T * C_{adt} * M(D,B,S)}{code\_rate}.$$

NOTE 2: C_{adst} is not considered in the equation; this is extremely important since it enables to de-correlate the IFEC memory sizing from the actual instantaneous variations of the traffic which is measured through the C_{adst}. More considerations on the case where C_{adt} and C_{adst} are different is for further study.

NOTE 3: Memory will need to be sized for a minimum code rate since, the lower the code rate, the higher the memory requirement.

6.2.6.3 Implementation aspects

Some implementation aspects of the memory are described in the following, in particular the memory requirements. Basically two opposite scenarios can be described:

- a first scenario will size the memory based on peak traffic ($C_{adt} = C_{adst}$);
- a second scenario will size the memory on averaged traffic (C_{adt} may differ from C_{adst}). This scenario is for further study.

In the first approach based on peak traffic, the memory requirement is to use M(D,B,S) times the memory required for a maximum burst size as given by the product $T*C_{adst}$. Implementation-wise, each ADST is constituted of a list of C_{adst} pointers, each pointer addressing one of the C columns of one of the M ADTs via the adt_index and adt_column functions. If one ADST column is padded, the memory is still reserved inside the corresponding ADT, leading to potential memory waste. In addition, there must be R columns for the FDT, R being equal at least to $C_{adst} * \frac{1 - code\_rate}{code\_rate}$, leading also to wastes on the FDT part. This is illustrated in figure 6.20.

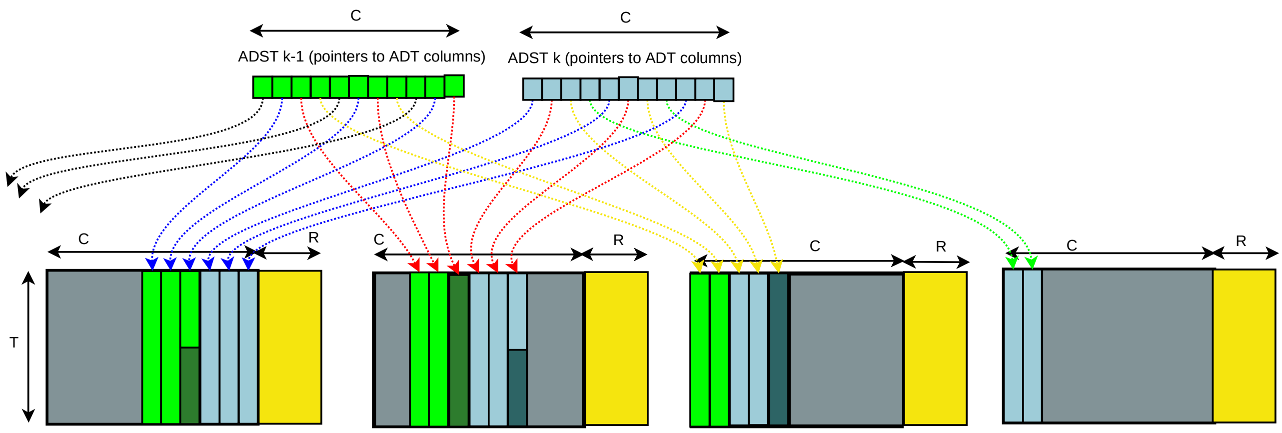


Figure 6.20: Implementation 1 (single pointer)

On this figure, we can see that the padding column is greyed but its memory is "physically" reserved.

The main drawback of this implementation is particularly apparent in case of VBR where it can lead to excessive memory since, for a lot of datagram bursts, the ADST will be made of many padded columns. If we consider the typical example of clause 6.2.3.1, we have on average 37 columns. If we reserve the full 191 columns, this means we will have a waste factor of 5!

Memory evaluation: we provide some memory quantitative evaluations.

For the average MPE traffic, we use the parameter max_average_rate, given by the time_slice_fec_indentifier. In order to represent the burst traffic fluctuations, we introduce a parameter called traffic_variations_ratio which represents the burst traffic variations around this average value. $\forall k \geq 0, C_b(k) \leq C_{b\_ifec} * (1 + traffic\_variation\_ratio)$. This parameter can be important for VBR traffic (more than 100 %). Even in the case of CBR, there are small variations of traffic around a medium value and the ratio is not null.

Hypothesis 4: we assume also that a repetition_interval (average time distance between two successive bursts) can be given. This can be compute by averaging over a sufficient number of bursts.

The number of rows nof_rows is given by T found in the time_slice_fec_identifier (we take 1 024 in the following).

C_{adst} is then sized by the average traffic plus the traffic_variations_ratio and given by following formula:

$$C_{adst} = \text{ceil}\left(\frac{\text{max\_average\_rate} * \text{repetition\_interval} * (1 + \text{traffic\_variations\_ratio})}{\text{nof\_rows} * 8}\right)$$

a) Implementation based on peak traffic: C_{adt}=C_{adst}

b) Implementation based on averaged traffic: for further study

The final C_{ADT} must be taken as the minimum of the two values - $\min(C_{adt};C_{adst})$ -.

In any cases, $R = \text{ceil}\left(\frac{\text{fec\_ratio}}{1-\text{fec\_ratio}} * C_{adt}\right)$

For both cases, assuming for B and S the B_{opt} and S_{opt}, we can derive memory as a function of (B+S), code_rate and max_average_rate:

$$\left.\begin{array}{r}\left.\begin{array}{r}(B+S)\\ \text{code\_rate}\end{array}\right\} \Rightarrow (B_{opt};S_{opt}) \Rightarrow M(D;B;S)\\ \left.\begin{array}{r}\text{max\_average\_rate}\\ \text{repetition\_interval}\end{array}\right\}\qquad\qquad\end{array}\right\} \Rightarrow (C_{adst};C_{adt};R)$$

The resulting bit rates are given in figure 6.21 for some typical maximum average bit rates values (128, 256, 512) and with a maximum value of code rate of 50 %.

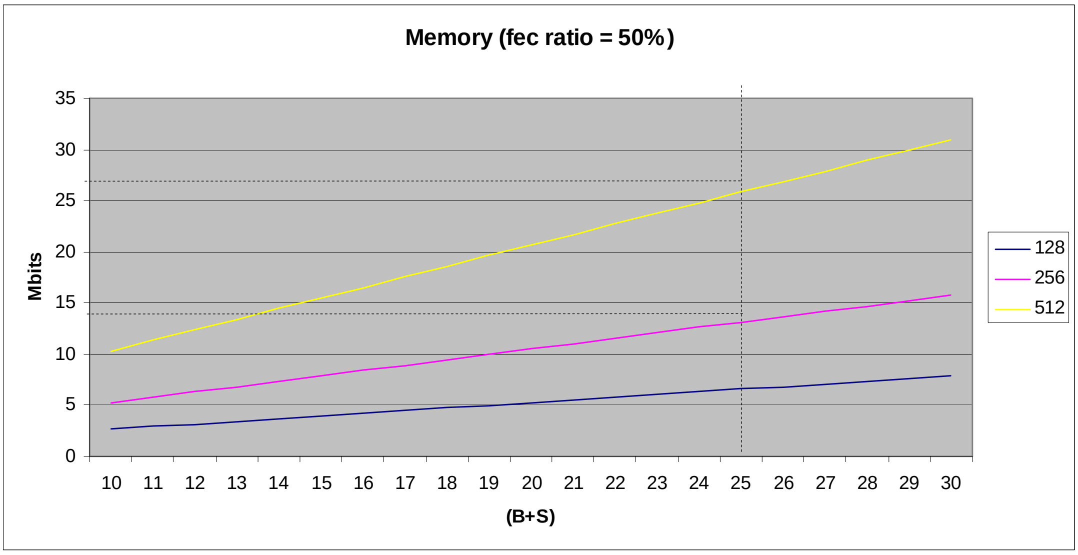


Figure 6.21: Memory requirements in CBR

For a 250 kbps CBR stream, based on simulated performance given in clause 6.2.5, we recommend values presented in table 6.12.

Table 6.12: Typical memory requirements for a CBR stream

	Modulation	B+S (bursts)	Code rate (%)	Memory (Mbits)
LMS-ITS	16QAM	25	40	13
	O-QPSK	20	55	10
	TDM	15	65	8
LMS-SU	all	10	66	5
NOTE 1: These values are excluding any memory for the MPE-FEC decoding. NOTE 2: These values are applicable for CBR. For VBR, one must ensure that the burst size does not exceed C*T. Memory optimizations for supporting efficiently VBR are for further study.				

6.2.7 MPE-IFEC usage scenarios

6.2.7.1 Definition

MPE-IFEC sources: these are sets of elementary streams transporting MPE-IFEC sections coming from the same encoding matrices and having coherent signalling. These elementary streams may be located on same or different TS, same or different frequencies, same or different radio technologies.

6.2.7.2 Introduction

This clause describes typical scenarios of MPE-IFEC usage in a DVB-SH environment, leveraging on the signalling flexibility. We distinguish two basic scenarios, one using a single source of MPE-IFEC and another one using two sources, the latter being for further study.

Single MPE-IFEC source

This is the basic scenario that has been used throughout the document. The system is delivering an IP flow sent in the format of MPE sections, protected by MPE-IFEC sections sent together with the original MPE data sections over the same frequency. This scenario applies for instance to SFN cases where the satellite and CGC are operating on the same modulation and frequency, forcing exactly the same content at the section level to be transmitted under the different areas (satellite only, hybrid satellite and CGC, CGC only).

In a constant bit rate situation, the number of MPE and MPE-IFEC sections is constant from an IFEC time-slice burst to another IFEC time-slice burst. However, when a variable bit rate traffic must be supported, the number of sections present in each burst can vary dramatically. This clause describe how these variations are handled by the sender and how the signalling helps the receiver coping with them.

Assuming the general case of variable bit rate traffic, the amount of MPE-IFEC section is limited by the following factors:

- the used code rate: as explained in clause 6.2.3, maintaining a target code rate will enable a maximum number of MPE-IFEC sections in iFDT is equal to:

$$\text{nof\_fec\_columns\_code\_rate}(i) = \text{ceil}\left(\text{nof\_data\_columns}(i) * \frac{1 - \text{code\_rate}_{\text{target}}}{\text{code\_rate}_{\text{target}}}\right);$$

- the capacity of the IFEC time-slice burst, usually computed in number of MPEG2 TP that turns into a number of MPE-IFEC sections once the capacity to send the MPE sections has been used. We assume this capacity to be nof_fec_columns_capacity(i).

It is assumed that the two variables will not differ significantly, in particular because the IP encapsulator will change nof_fec_columns_capacity(i) to follow the variations of nof_fec_columns_code_rate(i). However some adjustments are possible, especially if the nof_fec_columns_capacity(i) is slightly superior to nof_fec_columns_code_rate(i). It would then be useful to use the resource to send an additional MPE-IFEC section. For that purpose, the `max_iFDT_column` must be sent to a value superior or equal to the largest nof_fec_columns_capacity(i) plus additional margin multiple of S.

The procedure to constitute the IFEC burst can be the following:

- for each IFEC burst, list sections eligible for inclusion inside the burst according to the interleaving mechanism explained in clause 6.2.3.9;
- for that purpose, pick in the S previous iFDT the required number of sections until one of the following conditions is reached:
    - no more IFEC sections are available (in that situation an MPE-IFEC section index discontinuity may happen as explained in clause 6.2.3.9);
    - no more capacity is available in the MPE-IFEC time slice burst;
- once the list of MPE-IFEC sections has been established, insert them in the MPE-IFEC time slice burst and set their real-time information in the header:
    - set the burst number to current value;
    - set the section index to the current value;
    - set the iFDT index equal to the source iFDT;
    - set max_iFDT_column equal to the value computed from target_code_rate as explained in clause 6.2.3.8;
    - set IFEC_burst_size equal to total MPE-IFEC size;
    - compute real-time parameters, including Delta-t (the latter in exactly the same way as in MPE as explained in clause 6.2.3.11).

Typical algorithm supporting such procedure are for further study.

Multiple MPE-IFEC source

Scenarios allowing multiple sources of MPE-IFEC sections are for further study.

6.3 Time-Slicing

Time slicing is one of the key features introduced by DVB-H. It enables three important features:

- to spare battery by powering off the receiver during intervals when no service is listened;
- to support fast zapping;
- to support variable bit rates and so statistical multiplexing in TDMA environment.

The DVB-SH makes full usage of this time slicing information in order to provide same kind of feature support. However the way this is handled depends on the choice of the long interleaver for protection against long impairments (class 1 or 2). In the following we first provide insight on the way Time-slicing signalling is handled in DVB-SH and the impacts of this handling on power saving and VBR. In each impact case, we differentiate between class 1 and class 2 long interleavers.

6.3.1 Signalling

For class 1, time slicing information is important since it enables the terminal to power off. Time slicing signalling is conveyed by data and FEC section headers in real-time parameters (taken from MAC destination bytes in MPE). Since the Time-slicing information is repeated in all sections, the probability to not receive this information is equal to the probability of losing all sections and, if ts_error_indicator is used, to the probability of having, for all section headers, at least one TS among the group conveying those header, erroneous. When such case happens, the receiver will loose Delta-t synchronization and will stay on until it can process a full section header and get a new Delta-t.

Table 6.13: LMS-ITS ratio of lost burst (%)

case	14	17	18	18	21	24
	its	its	its	its	its	its
#burst	3 691	3 691	3 691	3 691	3 595	3 586
#errburst	1 424	1 641	1 251	833	1 069	1 096
#lost burst	978	1 193	788	300	24	27
#lost bursts / #bursts (%)	26	32	21	8	1	1

Table 6.14: LMS-SUB ratio of lost burst (%)

case	24	24	27	30	31	34	37	77	80	80
	sub	sub	sub	sub	sub	sub	sub	sub	sub	sub
#burst	3 667	3 586	3 667	3 667	3 667	3 594	3 590	8 939	3 583	8 984
#errburst	278	1 096	278	300	262	344	342	2 237	139	2 214
#lost burst	190	27	190	223	166	9	12	1 061	2	953
#lost bursts / #bursts (%)	5	1	5	6	5	0	0	12	0	11

Assuming typical case of LMS-ITS, we have (case 18) 21 % of burst that are lost for which no Delta-t can be retrieved. During 20 % of bursts, the terminal, unless some different strategy is applied, will go back to power on, leading to a possible power saving impact. This can be approximated by saying that 80 % of the time, the normal power saving is used while during 20 % of the time no power saving is used. This leads to a degradation of 20 % of power saving.

For class 2 receivers, time slicing information is less important since the long interleaver makes the real-time parameters to be outdated. Other techniques such as the one presented in clause 7.2.3.3.1 enables to rely on other structures like the DVB-SH services signalled by SHIP.

6.3.2 Zapping time impact

This clause addresses how zapping time is impacted by the DVB-SH.

DVB-SH introduces long interleaving, at physical and link layer to counteract long fading experienced in LMS channels. Long interleaving can appear as contradictory with the zapping time since, to ensure protection of the stream, the receiver would need to wait for the full duration of the de-interleaving. Forcing such long zapping times is not acceptable from a user perspective, so different techniques are supported by DVB-SH to provide "fast zapping" while still ensuring good protection levels.

For class 2 receivers, the physical interleaver can be tuned to provide fast zapping using uniform late profiles as described in clause 7.3.3.5. Choosing e.g. a 50/50 uniform/late profile, the zapping time is the time duration of one time interleaver burst which corresponds to roughly 200 ms on the physical layer, plus link layer delays and I-frame searching. The price to pay is that this burst must be received with a few dB more than what is the C/N threshold for this code rate (e.g. for rate 1/4 now with 4,4 dB instead of -0,9 dB as presented in clause 7.3.2.6.4).

For class 1 receiver, selection of adequate link layer parameter (D) will also help reducing the zapping time and increase reception quality during zapping period. Such configurations enable to display the first well received burst immediately after physical interleaver latency assumed to be around 200 ms, plus the delay to receive the burst and search the I frame, so without waiting for the full redundancy to be received. As presented in this clause, the fast zapping provide immediate display in 80 % of the bursts in most stringent channels. The price to pay is that full redundancy has not yet been received and FEC recovery techniques such as the ones presented in clause 6.2.5.2 must be applied to increase protection for following bursts.

6.3.3 Power saving impact

One of DVB-SH key elements is the introduction of a longer time interleaver at the physical level that has a negative impact on the power saving. Two cases are presented below depending on which type of interleaver is used.

6.3.3.1 "Terrestrial" physical interleaver

The linear part of the interleaver introduces a latency within the receiver, because of the convolutional nature, that adds a net time to the power on. The following formula can be used for deriving this information:

$$\text{time\_on} = \text{acquisition\_time} + \text{physical\_interleaver\_duration} + \text{burst\_duration}$$

- burst_duration can be approximated by $\frac{\text{repetition\_interval}}{\text{number\_of\_services}}$

$$\text{power\_saving} = 1 - \frac{\text{time\_on}}{\text{repetition\_interval}}$$

- depending on repetition_interval the power saving will evolve; typical values are:
  - 1 s for repetition interval and 9 programs => burst_duration~111 ms;
  - acquisition_time~50 ms;
  - physical interleaver~200 ms;
  - So time_on~361 ms and power_saving~64 % instead of 84 % in DVB-H.
- impact of loss pattern:
  - as referring to clause 6.3.1 we have power saving degradation of 20 % so 0,8*64 = 51 %.

6.3.3.2 "Long" physical interleaver

In that case, the involved durations are much longer than "terrestrial" case, in the order of multiples of SH-frames. In that conditions, time slicing signalling cannot really be used for providing the relevant off times since the decoding and MPE header processing happens after a time de-interleaving that lasts longer than the Time-slicing signalling itself. However, when service synchronization between DVB-H and DVB-SH is activated, it is possible to recover partially the power saving gain. In service synchronization, bursts are grouped in DVB-SH services signalled by SHIP packet. DVB-SH services are fixed-sized and their repetition interval does not vary. The terminal is then able to pre-determine the off periods without any knowledge on the MPE Delta-t: Time-slicing is actually managed at physical level. This approach is explained in clause 7.2.3.3.1.

- if we take the previous example and consider 3 DVB-SH services, we find:

$$\text{time\_on} = \text{acquisition\_time} + \text{physical\_interleaver\_duration} + \text{burst\_duration}$$

$$\text{time\_on} = \text{acquisition\_time} + \text{late\_tap\_interleaver} + \frac{\text{repetition\_interval}}{3} \qquad \text{power\_saving} = 1 - \frac{\text{time\_on}}{\text{repetition\_interval}}$$

- assuming same interleaver configuration as the one used for the class 2 terminal fast zapping in clause 6.3.2 and same DVB-H service structure as in the short interleaver case in clause 6.3.3.1, we have following values:
  - 1 s for repetition interval;
  - acquisition_time~50 ms;
  - physical interleaver (late taps only)~200 ms;
  - So time_on~583 ms and power_saving~42 % instead of 84 % in DVB-H.
- impact of loss pattern:
  - this degradation of power saving is compensated by the "constant" service structure: there is no need to switch on and search for the next burst and, in case of lossy channels, the power saving is not impacted by the low pattern contrarily to the class 1 (power saving maintained at 42 %).

6.3.3.3 Summary

For a typical configuration with 9 services of 111 ms duration each and a mapping of 3 services within 1 DVB-SH service, the achieved power savings are given in table 6.15. Please not that these values are given as examples, actual values will vary between configurations.

Table 6.15: DVB-SH impact on power saving

	DVB-H (reference)	DVB-SH class 1	DVB-SH class 2
Power saving clear	84 %	64 %	42 %
Power saving lossy (ITS)	N/A	51 %	42 %

6.3.4 VBR/statmux impact

6.3.4.1 Interest of statistical multiplexing

Experiments indicate that when video is coded at a fixed quantizer step size (almost constant quality), the peak bit rate of a difficult picture frame may become more than 10 times the mean bit rate. However, when sufficient number of videos (e.g. around 10) are statistically multiplexed, the overall rate is typically only 1,2 times the long term mean bit rate. Higher bandwidth per channel is required for a fewer number of multiplexed videos." This suggests that for a approaching the average bit rate allocation, one should have a large enough number of channels, typically 10.

Although traffic variations of a statistically multiplexed stream may be shaped and turned into constant bit rate traffic, this would be done at the penalty of the receiver buffering and so zapping time. In order to avoid this buffering time, it is important to support "real-time" variations of the traffic bit rate. DVB-H support such a variation by letting burst vary in size from frame to frame. DVB-SH also supports these variations in a similar way as DVB-H although exact way of supporting depends on physical layer interleaver choice (class 1 and class 2) because different classes do not "mix" DVB-H services in the same way.

6.3.4.2 Terrestrial (short) physical interleaver

With class 1 receives, the long interleaver is performed at the link layer. Since link layer also provides Delta-t information, the burst can freely vary in size and so VBR is fully supported: the addition of MPE-IFEC does not change or delay the reception and processing of the Delta-t signalling. VBR implies that useful traffic will vary from burst to burst, possibly in large proportions. For instance, the volume could multiply (or divide) by a factor of 2 between two successive bursts. Without any IFEC protection, a statistical multiplexing control would allocate dynamically the bandwidth to different streams so that compound of all streams stays within radio fixed capacity. This means that starting dates for burst (and Delta-t) will vary from burst to burst. The variable DATA burst sizes will make ADT more or less filled with DATA columns. Then different strategies can be applied to benefit from this gain of bandwidth:

- *FEC complement protection*: this strategy consists in varying the FEC so that, in each transmitted burst, the volume of DATA plus FEC is always the same. By this means, the freed volume by statmux is used to increase FEC protection and link quality. This leads also to varying code rates, with values lower than the minimum specified. Of course the capacity will not be increased. In the sender operation of MPE-IFEC specifications [MPE-IFEC], clause 3.5, the limitating factor becomes the fec_burst_size computed so that burst size is fixed, the number of FDT FEC columns being as large as required;

- *fixed FEC code rate*: this strategy consists in keeping FEC code rate fixed while increasing the capacity. Depending on the actual ADT size in data columns, a varying number of FEC columns is created so that code_rate_FDT = nof_ADT_data_column / (nof_FDT_fec_column + nof_ADT_data_column). In the sender operation of EN 302 583 [1], clause 3.5, the limitating factor becomes the number of FDT columns dynamically set by the code rate, the fec_burst_size being as large as the maximum size.

Usual implementations maintain the bit rate of a group of individually services within a fixed setting. It is clear in that situation that the *D* parameter MUST be the same for all services, otherwise the different delay applied to services may lead to variable bit rate of the group of services and, possibly, exceed the setting as presented in figure 6.22. This does not prevent having, for instance, one group of statistically multiplexed services with a fixed setting and same D, and another group with another D.

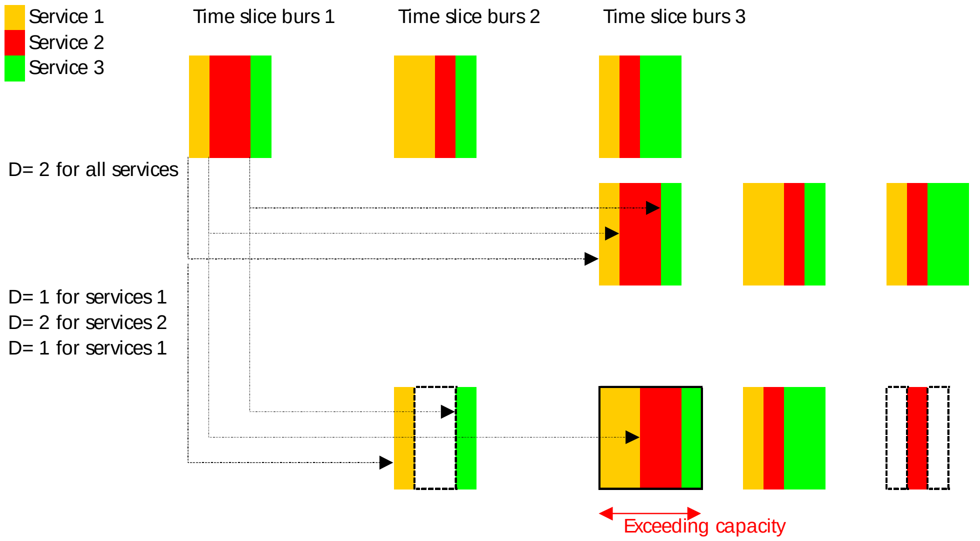


Figure 6.22: influence of D in a group of statmuxed services

To summarize, the terrestrial physical interleaver combined with a link layer interleaver provides sufficient flexibility for supporting VBR and statmux in possibly different ways, some similar to the DVB-H way.

6.3.4.3 Uniform long interleaver

As expressed in clause 6.3.3.2, the uniform long interleaver "breaks" the time slicing signalling. Unless power saving is not a key criteria of the system, in order to support VBR, DVB-H services must be grouped in "DVB-SH services" as presented in clause 7.2.3.3.1. Then it is possible to perform VBR among the services of a same DVB-SH service. The impact on the statistical multiplexing will depend on the size of the DVB-SH service and the number of DVB-H services grouped within, it will grow as the DVB-SH service becomes smaller in size. Actual performance of such a mapping is out of scope of this release.

In case power saving is not a key criteria, there is no need to group DVB-H services in DVB-SH services and each DVB-H service can freely vary in size: the system is then completely compatible with DVB-H statistical multiplexing.

6.3.5 Conclusion

Table 6.16 summarizes the impact of the DVB-SH system on the different features of the time slicing and related link layer features.

Table 6.16: Impact on link layer

	Class 1	**Class 2**
MPE Signalling	Ok	MPE Signalling is obsolete, rely rather on SHIP
Power saving	ok but depends on repetition_interval Degradation in erasure channel according to % of completely_lost_burst	Depends on mapping between DVB-H/SH services (better when 1 to 1 m mapping) rather independent of channel burst loss statistics due to fixed structure
Statistical multiplexing with time slicing	full support	Depends on mapping between DVB-H/SH services (worse when 1 to m 1 m mapping)

Class 1 has less impact / better support of the time slicing and related features: all key features of the link layer can be satisfied at the same time. The class 1 receiver can fully exploit the time slicing and its related features but bandwidth efficiency and power saving may be degraded in certain LMS channels.

Class 2 can compensate the "obsolescence" of the time slicing via the mapping of DVB-H over DVB-SH services, but the mapping optimization goes in two different directions depending if power saving or statistical multiplexing is sought. Actual positioning will then depend on system constraints and in particular on terminal capacity to sustain more memory and more battery requirements: a class 2 terminal having no battery restriction and being always on will be as efficient with statistical multiplexing as a class 1. For other class 2 receiver, a proper balance has to be found between power saving and the gain of statistical multiplexing, dependent on the application scenario.

6.4 Mobility

Mobility is supported in a similar way as DVB-H and subject of a specific implementation guideline. Specificities of DVB-SH mobility is the usage of hybrid_delivery_descriptor that is similar to a terrestrial and/or satellite delivery descriptor and usage of SDT service_availability. Such specification is for further study.

7 Physical Layer elements

This clause summarizes the different building blocks used in DVB-SH. Most of the blocks have already been used within other DVB standards EN 302 304 [3] and EN 302 307 [6] and are considered to be sufficiently well described in the references. Those elements which have been introduced to DVB within the DVB-SH waveform are described with a high degree of detail.

7.1 Overview to the physical layer elements

An overview on the physical layer technologies of the DVB-SH system is given in tables 7.1 and 7.2.

The different technology submodules are grouped as follows:

(A) encapsulation, forward error protection, interleaving and frame adaptation;

(B) OFDM modulation including TPS and reference signal insertion as well as Fourier Transform processing;

(C) TDM modulation including Pilot field insertion and roll-off filtering.

Table 7.1: Technology sub-modules and their descriptions (part 1)

Category	Technology sub-module	Description	Related features
(A)	Turbo code with block length of 12 282 bits	Subset of 3GPP2 turbo code has been selected as FEC scheme.	High power efficiency. Block length of 12 282 bits offers also high flexibility for time interleaver design.
(A)	Turbo code with code rates between 1/5 and 2/3	Wide range of code rates with stepping of approx. 1…1,5 dB in terms of required energy per code bit over N_0.	High receiver sensitivity for low code rates. Low code rates allow powerful time interleaver design and outage protection at physical layer. Higher code rates are selected when link layer protection is used.
(A)	block code structure suitable for MPEG-TS encapsulation	Block structure allows to encode 8 MPEG-TS packets in one turbo encoded word. Block code framing is aligned to the SH framing.	Diversity combining is simplified. Synchronization information for hand-over and combining can be derived from the framing.
(A)	CRC for each MPEG-TS packet	Additional error detection mechanism.	Allows support of error mitigation techniques or link layer protection.
(A)	Flexible time interleaving	Time interleaving of different length is applied at physical layer.	Flexible exchange of fading protection between physical layer and link layer.
(A)	Short time interleaving (approx. 300 ms)	Interleaver length is selected accordingly to available memory. Class 1 receivers with reduced physical layer memory size are supported.	Additional link layer protection is selected to cope with fading channels, especially for the satellite.

Table 7.2: Technology sub-modules and their descriptions (part 2)

Category	Technology sub-module	Description	Related features
(A)	Long uniform time interleaving (approximately 10 to 15 s)	Best "channel averaging" for fading channels and short random signal blockages. Class 2 receivers with extended physical layer memory size are addressed.	Combined with low code rates, receiver provides high sensitivity. All signal blockages are handled by the physical layer. Profile may have impacts on the access time after switch-on or recovery after blockage.
(A)	Long uniform/late time interleaving (approximately 10 to 15 s)	Interleaver profile optimized for short zapping time and fast recovery after temporary longer signal blockages. Reduced "channel averaging" for fading channels. Class 2 receivers with extended physical layer memory size are addressed.	Combined with low code rates, receiver provides high sensitivity. All signal blockages are handled by the physical layer. Profile allows fast access time after switch-on or recovery after blockage due to the "late" burst contribution.
(B)	Pilot symbol aided OFDM	Waveform identical to DVB-T with changes in the TPS bit description.	Allows reuse of existing OFDM demodulators for DVB-T or DVB-H, with some adaptations.
(B)	Addition of 1 k mode	Waveform for 1 k FFT length added to support higher speeds and/or smaller bandwidths.	Allows reuse of existing OFDM demodulators for DVB-T or DVB-H, with some adaptations.
(B)	Addition of 1,7 MHz channelization	Waveform for L-Band channelization added.	Allows reuse of existing OFDM demodulators for DVB-T or DVB-H, with some adaptations.
(C)	Pilot symbol aided TDM	Waveform derived from DVB-S2 with fixed framing and preamble distance.	Pilot symbol pattern is designed to support synchronization and tracking also at very low C/N values. Regular pilot scheme allows the prediction of framing.
(C)	Modulations QPSK, 8PSK and 16APSK	Waveform allows flexible choice of modulation independent of the physical layer code rate. Modulations are derived from DVB-S2.	Different modulation orders allow to efficiently use available satellite power, independent selection of parameters for satellite and CGC and support for local content insertion.

7.2 Turbo code and time interleaver

7.2.1 Introduction

This clause gives an overview on features which comprises of the encapsulation and encoding of MPEG-TS packets into turbo encoded words. Additionally, the time interleaving operating on turbo encoded words is introduced.

7.2.2 Turbo code

7.2.2.1 Introduction

This clause will explain the characteristics of this FEC solution from the receiver side. Details are also given for combining received code bits from two reception chains (OFDM/OFDM non-SFN or OFDM/TDM) with identical or different puncturing pattern and/or code rates.

The transmission side is described in the waveform document [1], therefore the focus of this clause will be (after some short introduction on the encoder) on the implementation of the turbo decoder.

7.2.2.2 Overview of the key elements

7.2.2.2.1 Top level description

The turbo encoder is a block encoder, which works on a block length of $L_{TC\text{-}input}$ bits. The parameter $L_{TC\text{-}input}$ can be either 1 146 bits (signalling field) or 12 282 bits (payload).

The turbo encoder consists of two Recursive Systematic Convolutional (RSC) encoders, each of the encoders producing an output of three bits. The first RSC produces the bits X, Y_0 and Y_1. The output of the second RSC encoder is called X', Y_0' and Y_1', respectively.

The two encoders are connected by an intra-codeword interleaver (called *3GPP2 interleaver* in figure 7.1). The interleaving instruction is given in EN 302 583 [1], clause 5.3.3. For each block length $L_{TC\text{-}input}$, the instructions to shuffle the input bits are different; however this can be described by selection of one parameter and another small lookup table.

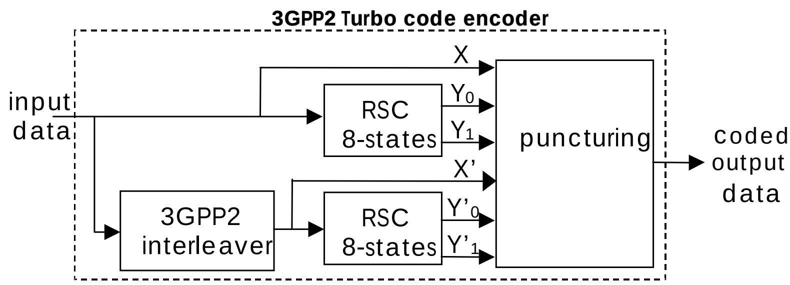


Figure 7.1: Turbo encoder schematic

7.2.2.2.2 Parameters and numbers

The number of output bits at the *internal* interface [X Y_0 Y_1 X' Y_0' Y_1'] is (independently of the selected code rate or puncturing pattern):

$$N_{Outputs}*(L_{TC\text{-}input}+L_{TailBits})$$

This results in the following parameters:

- $N_{Outputs}$ = 6 (size of the internal interface [X Y_0 Y_1 X' Y_0' Y_1'])

- $L_{TC\text{-}input}$ = 1 146 or $L_{TC\text{-}input}$ = 12 282
- $L_{TailBits}$ = 6

Using the above formula, it can be derived that for the payload block length $L_{TC\text{-}input}$ = 12 282, the number of bits at this internal interface is 73 728 bits. From these bits, only a fraction is used for transmission. This fraction is derived by the puncturing pattern for the data and the tail bit periods. An example is given in the next clause.

7.2.2.2.3 Examples for puncturing the [X Y0 Y1 X' Y0' Y1'] vector

The method of puncturing the data and the tail periods of the turbo encoder is described in the waveform document [1] for code rates 1/N with 1<N<5. For other code rates like 2/5, the generation of the data and tail periods is more difficult, therefore an additional example is given here.

- Puncturing pattern ID = 6: code rate R=2/5, standard
- Data puncturing pattern:
  1;0;0;0;0;0; 1;0;1;0;0;1; 0;0;1;0;0;1;
  1;0;1;0;0;1; 1;0;1;0;0;1; 0;0;1;0;0;1;
  1;0;1;0;0;1; 1;0;1;0;0;1; 0;0;1;0;0;1;
  1;0;1;0;0;1; 1;0;1;0;0;1; 0;0;1;0;0;1
- Tail puncturing pattern:
  1;1;1;0;0;0; 1;1;1;0;0;0; 1;0;1;0;0;0;
  0;0;0;1;1;1; 0;0;0;1;1;1; 0;0;0;1;0;1

The puncturing patterns are organized in groups of 6 elements, each such group specifying the puncturing of the 6 internal output bit vector [X Y_0 Y_1 X' Y_0' Y_1'] for one data or tail bit.

The repetition period of the data puncturing pattern is 12 input bits. For the turbo input word length of 12 282 bits, the first 12 276 input bits are processed by repeating the data puncturing pattern 1 023 times (`floor(12 282/12)`). For the remaining 6 input bits (the 12 277th to the 12 282th input bit), only the first 6 groups of the data puncturing pattern are used, before the tail puncturing pattern is applied. The number of generated bits using the data puncturing pattern is 30 704 bit (selecting punct_Pat_ID = 6).

The tail puncturing pattern is used for the 6 tail bits only, and it is not repeated. The number of generated bits using the tail puncturing pattern is 16 bit (selecting punct_Pat_ID = 6).

In total, the turbo encoder generates 12 288/R = 30 720 bit.

Please note that the number of bits generated using the data puncturing pattern is *not always* 12 282/R and the number of bits generated using the tail puncturing pattern is *not always* 6/R for all selections of the puncturing pattern ID, however the overall number of generated bits using the data and the tail puncturing pattern is *always* 12 288/R.

The short block length of 1 146 input bit is only used together with puncturing pattern ID = 0, such that the statements above only apply to the block length of 12 282 bit.

7.2.2.2.4 Bit-wise interleaver

The bit-wise interleaver is an intra-codeword interleaver which operates on the punctured output of the turbo encoder. The interleaver increment is selected according to the code rate and therefore according the turbo code word length at the output of the puncturing. Its task is to prepare the code word for the use together with the convolutional time interleaver and the transmission over channels with burst erasure behaviour.

Taking into account the interleaver profiles, it can be seen that - what concerns one code word - the distribution of the encoded bits in time is usually not equally spaced but somehow grouped:

- first of all, the minimum unit to be processed by the time interleaver is one interleaving unit (IU) of 126 bit each;
- additionally, the use of time slicing groups several of these IU onto one relatively small burst.

If one or more IUs or time slices are lost, the performance of the turbo decoder may suffer from such bursty data losses. The approach of the bit-wise interleaver is to transform bursty losses on the transmission channel into "approximately evenly distributed" bitwise erasures at the input of the turbo decoder in the receiver. This loss pattern can be recovered with higher probabilities by the turbo decoder, dependent on the reception condition of those bits which have not been erased or exhibited deep fades.

The bit-wise interleaver is designed such that the distance between erased bits of a burst is maximized; this is done by choosing an interleaver increment which is relatively prime to the codeword length after puncturing.

One example for code rate 1/5 is given in the figure 7.2. The parameters are:

- block length N_{TCB} after turbo encoder and puncturing: 61 440;
- bit-wise interleaver increment: 247.

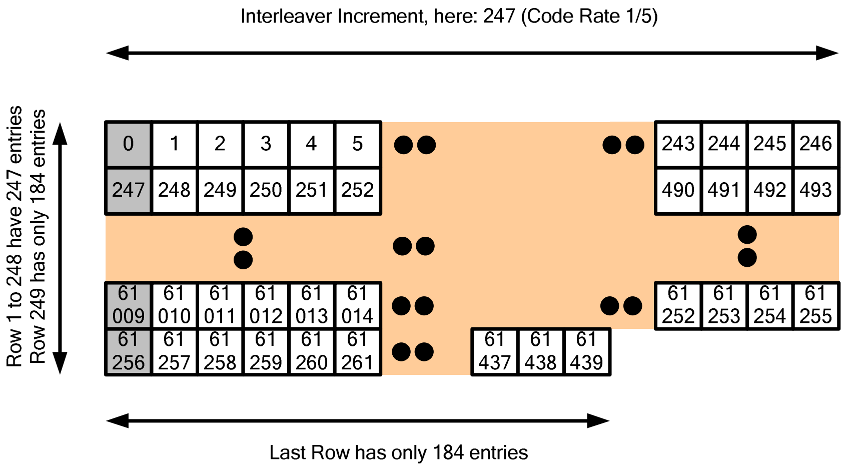


Figure 7.2: Bit-wise interleaver

The first 300 index positions are given in the following table:

$b_0 \ldots b_{299} = a_{H(w)}$ with H(w) = {

0	247	494	741	988	1235	1482	1729	1976	2223	2470	2717	2964	3211	3458	3705
3952	4199	4446	4693	4940	5187	5434	5681	5928	6175	6422	6669	6916	7163	7410	7657
7904	8151	8398	8645	8892	9139	9386	9633	9880	10127	10374	10621	10868	11115	11362	11609
11856	12103	12350	12597	12844	13091	13338	13585	13832	14079	14326	14573	14820	15067	15314	15561
15808	16055	16302	16549	16796	17043	17290	17537	17784	18031	18278	18525	18772	19019	19266	19513
19760	20007	20254	20501	20748	20995	21242	21489	21736	21983	22230	22477	22724	22971	23218	23465
23712	23959	24206	24453	24700	24947	25194	25441	25688	25935	26182	26429	26676	26923	27170	27417
27664	27911	28158	28405	28652	28899	29146	29393	29640	29887	30134	30381	30628	30875	31122	31369
31616	31863	32110	32357	32604	32851	33098	33345	33592	33839	34086	34333	34580	34827	35074	35321
35568	35815	36062	36309	36556	36803	37050	37297	37544	37791	38038	38285	38532	38779	39026	39273
39520	39767	40014	40261	40508	40755	41002	41249	41496	41743	41990	42237	42484	42731	42978	43225
43472	43719	43966	44213	44460	44707	44954	45201	45448	45695	45942	46189	46436	46683	46930	47177
47424	47671	47918	48165	48412	48659	48906	49153	49400	49647	49894	50141	50388	50635	50882	51129
51376	51623	51870	52117	52364	52611	52858	53105	53352	53599	53846	54093	54340	54587	54834	55081
55328	55575	55822	56069	56316	56563	56810	57057	57304	57551	57798	58045	58292	58539	58786	59033
59280	59527	59774	60021	60268	60515	60762	61009	61256	63	310	557	804	1051	1298	1545
1792	2039	2286	2533	2780	3027	3274	3521	3768	4015	4262	4509	4756	5003	5250	5497
5744	5991	6238	6485	6732	6979	7226	7473	7720	7967	8214	8461	8708	8955	9202	9449
9696	9943	10190	10437	10684	10931	11178	11425	11672	11919	12166	12413}				

7.2.2.3 Combining at the input of the turbo decoder

This clause explains decoding strategies in the presence of more than one demodulator and received signal.

7.2.2.3.1 Overview

Figure 7.3 gives a first overview on the combining possibilities. The architecture shown is a placeholder for real implementations but tries to sketch the principles behind combining at the turbo decoder input.

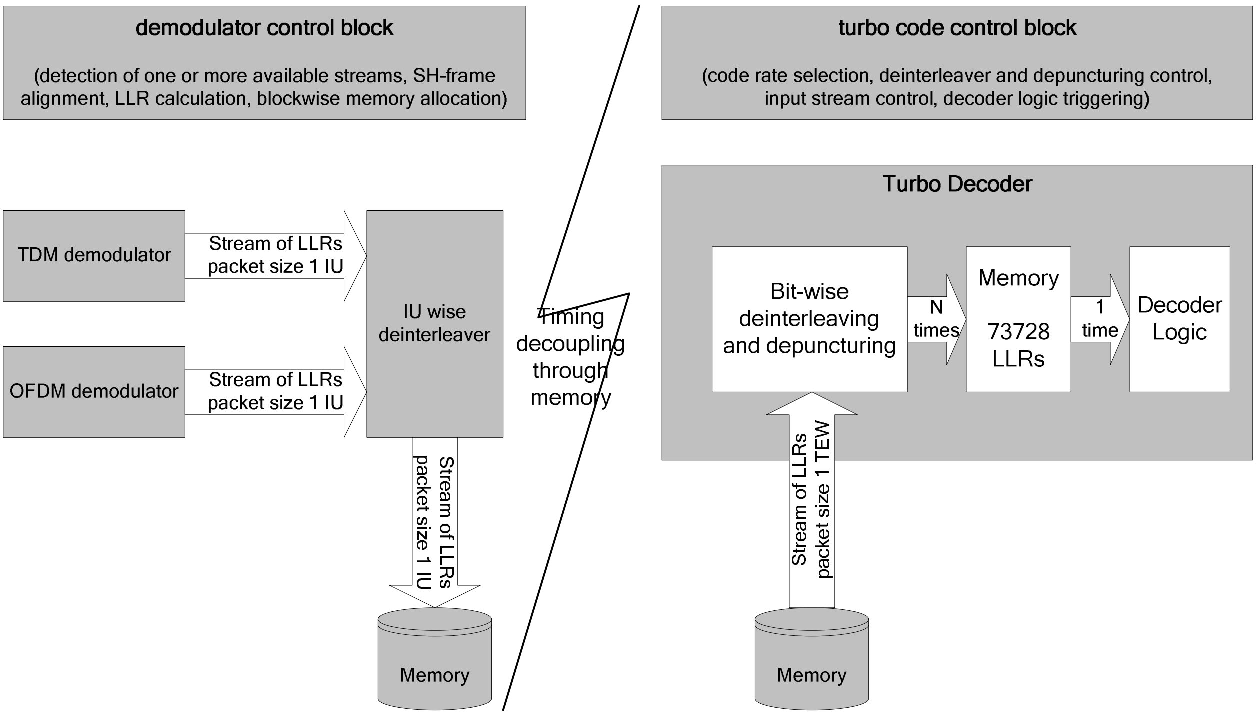


Figure 7.3: Possible turbo decoder integration into the receiver

Even in the presence of N>1 demodulated signals, the turbo decoder logic must be run only once after the combining of demodulated streams has taken place. In DVB-SH scenarios, either N = 1 or N = 2 demodulated signals are typically available at the receiver input:

- TDM for satellite transmission, OFDM for terrestrial transmission;
- OFDM for satellite and terrestrial transmission (in MFN mode).

As far as antenna diversity is considered, the combining of more than one antenna may take place before or after the demodulator stage, resulting in one or two demodulators from the turbo decoders' perspective.

7.2.2.3.2 Implementation

The turbo decoder itself can be considered to be made-up by various internal blocks. Their functionalities are:

- **Bit-wise deinterleaving:** the bit-wise deinterleaving reverts the bit-wise interleaving of the transmitter. Dependent on the selected puncturing pattern ID, this stage works on block sizes between 18 432 bits and 61 440 bits which can be directly read from the deinterleaver memory. This bit-wise deinterleaving is invoked several times per turbo decoding process, dependent on the number N of demodulators active for the same code word.

- **Depuncturing:** the depuncturing reverts the puncturing process of the transmitter. Dependent on the selected puncturing pattern ID, this stage works on the same input block size as the bit-wise deinterleaving. It is also invoked several times per turbo decoding process, if more than one demodulator is active.

- **Combining:** the combining reads the LLR entries already present in the decoder input memory (corresponding the codewords from those demodulators that have already been combined) and adds the LLRs just read from the interleaver memory for one codeword received by one of the N demodulators. Hence a codeword from another demodulator is combined with the codewords of those demodulators that have been combined by this block before for the current turbo info word. The combined LLRs are then again stored in the decoder input memory. This process is repeated until the N codewords from the demodulators have been combined that are associated with one turbo info word. Note that this process can even be carried out without performance loss when one of the demodulators does not receive a signal: in this case, the LLRs from this demodulator are zero, hence the LLRs from the other demodulators are not changed by adding the zero-codeword from the inactive demodulator.

- **Decoder Input Memory (either input and working memory or both memories combined):** this memory has the size of the non-punctured turbo encoder code word of 12 288*6 = 73 728 LLR values. Dependent on the architecture and the speed of the decoder logic, this memory has either to be instantiated twice (double buffering) or only once (see figure 7.4). This memory is read and written by the depuncturing stage several times per turbo decoding process, but read only once by the decoder logic when no other input is available and the decoder logic can start.

- **Decoder logic:** the decoder logic is called once all available demodulated streams have been processed and combined on the memory. The output is forwarded to the next stage in the receiver, e.g. the CRC16 check and the decapsulation.

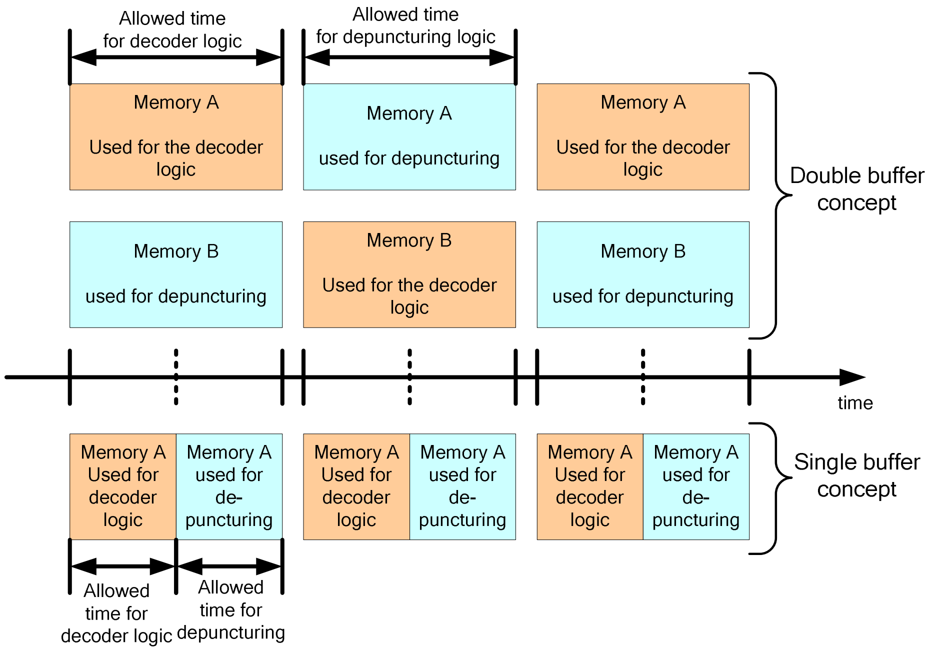


Figure 7.4: Possible turbo decoder memory architecture

7.2.2.3.3 Maximum ratio combining and complementary code combining

Combining of two signals is done by the accumulation of all LLRs available for one bit; it is a simple addition of LLR values. This is performed as depicted in figure 7.5. Please note that the figures do not represent exactly the puncturing patterns nor the code word structures introduced in DVB-SH, but should give an idea on the way combining works.

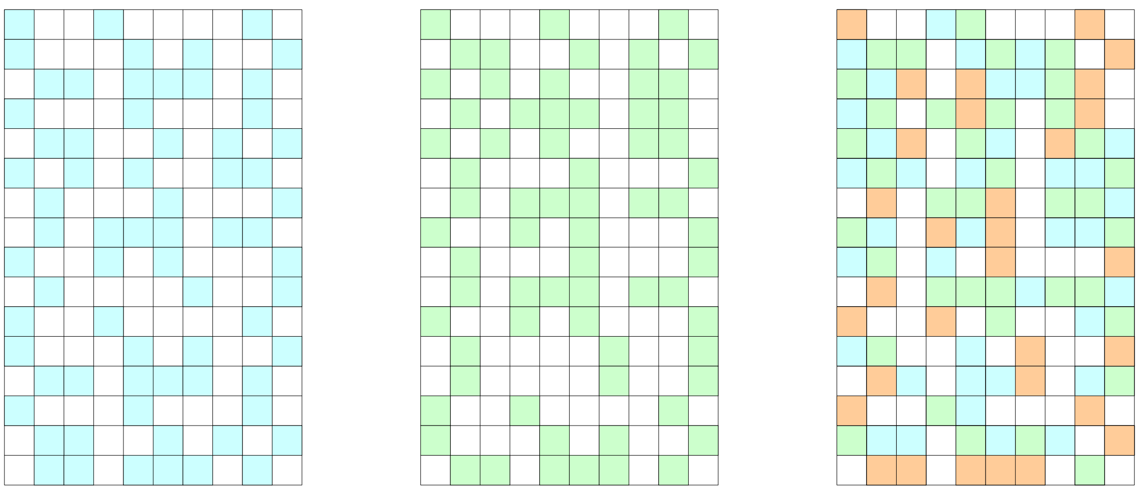


Figure 7.5: Received LLRs from demodulator 1 (blue), received LLRs from demodulator 2 (green) Combined LLRs from both modulators (blue, green, orange)

The different colours represent the LLRs received from different demodulators:

- **figure on the left hand side:** the blue squares represent LLRs received from demodulator 1. This codeword is a valid codeword which can be decoded at the threshold $(C/N)_1$ for the selected code rate (e.g. 1/2);
- **figure in the middle:** the green squares represent LLRs received from demodulator 2. This codeword is a valid codeword which can be decoded at the threshold $(C/N)_2$ for the selected code rate (e.g. 2/5);
- **figure on the right hand side:** all squares in colour represent the combined LLRs available from both demodulators. This codeword is a valid code word which can be decoded at the threshold $(C/N)_3$ for the combined code rate. The colours represent the following:
  - **orange:** these LLRs have been received by **both** demodulators and have been combined by pure addition of the LLRs. Addition of these LLRs is equivalent to **maximum ratio combining** of the demodulated signals;
  - **green/blue:** these LLRs have either been received through demodulator 1 or 2. Each green or blue square complements the other received code word. These LLRs are **complementary code combined**.

Dependent on the choice of puncturing pattern IDs on the transmit side, the percentage of LLRs being maximum ratio combined or complementary code combined can vary. In general, maximum ratio combining of the complete codeword is a special case of combining codewords for the following configuration:

- identical code rates on the different transmission paths;
- identical puncturing pattern IDs on the different transmission paths.

7.2.2.4 Selection of turbo code rate together with link layer protection

The DVB-SH standard offers a wide range of combinations for code rates both on the physical layer and the link layer. The overall spectral efficiency depends on both protection mechanisms and the selection of modulation order. A short computation example for the code rate selection is given here:

- physical layer code rate: $R_{PHY} = 1/2$;
- link layer code rate: $R_{IFEC} = 2/3$;
- overall code rate: $R_{OVA} = 1/2 * 2/3 = 1/3$.

The proper choice of code rates and modulation orders is the key for the solution of the trade-off between:

- high spectral efficiency (by selection of high code rates with lower protection);
- high robustness in typical DVB-SH environments (by selection of low code rates with high protection).

This choice is even complicated by the fact that the redundancy can be assigned seamlessly between physical layer and link layer. Some recommendations (for QPSK modulation order) are given here:

- assignment of **all redundancy** to the physical layer (typical $R_{PHY} = 2/5$ or lower):
  - should be chosen if long physical layer interleavers are used;
  - should be chosen if higher values of Doppler spread in OFDM have to be supported (see clause 7.3.1.3);
- assignment of **the larger part** of the overall redundancy to the physical layer (typical $R_{PHY} = 1/2$):
  - should be chosen if the link layer FEC protection is relatively low (typical R_{IFEC} around 2/3);
  - should be chosen if long physical layer interleavers can not be used, e.g. due to memory constraints;
  - may degrade the C/N thresholds for error-free reception as higher C/N values for static and mobile reception are necessary;
  - impacts the satellite link budgets and terrestrial coverage planning;

- assignment of **the smaller part** of the overall redundancy to the physical layer (typical R_{PHY} = 2/3):
  - should be chosen if the link layer FEC code rates is relatively high (typical R_{IFEC} around 1/2);
  - should be chosen if long physical layer interleavers can not be used, e.g. due to memory constraint say severely degrade the C/N thresholds for error-free reception as higher C/N for static and mobile reception are necessary;
  - impacts the satellite link budgets and terrestrial coverage planning.

According to the results presented in clause A.12, the selection of code rates on the physical layer has to be made according to the following criteria:

- available satellite link budgets (given EIRP) and terrestrial coverage planning (given repeater EIRP and density);
- physical layer margins necessary for certain reception scenarios and receiver classes;
- desired spectral efficiency and desired reception quality.

The choice to select the split of redundancy between the physical layer and the link layer will mainly be driven by the terminal classes addressed. The selection guide between solutions is for further study.

7.2.2.5 Processing at PHY to support Erasure decoding in UL

Being the DVB-SH interface between Physical and Link layer, the MPEG-TS is the only mean to transport reliable error information (synchronized with the data payload) between these two layers.

The DVB-SH frame supports an additional CRC16 error detection mechanism at the level of each individual TS. By checking this CRC16, the physical layer can detect erroneous TP and set accordingly the transport_error_indicator bit. This bit can then be used by following layers, in particular the link layer to optimize decoding process. Therefore, the mechanisms defined in ISO/IEC 13818-1 [8] for MPEG-TS packets should be used in order to signal the integrity of the MPEG-TS payload. Different algorithms are possible and one is proposed hereunder:

The algorithm depicted in figure 7.6 takes both the CRC16 in the EHEADER and the various CRC16 over the User Packets (UP) into account.

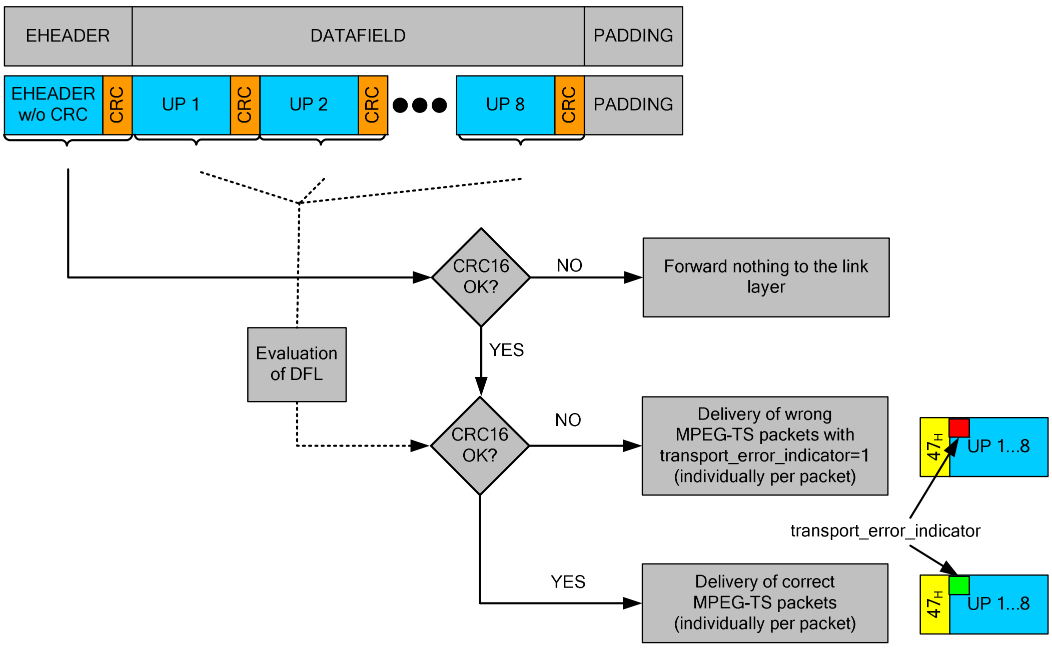


Figure 7.6: Proposed algorithm for decapsulation to support erasure decoding in UL

7.2.2.5.1 Overview on the proposed algorithm

In total, the payload of one turbo code word consists of up to 9 (nine) CRC-checks, each of length 16 bit:

- The first one to be considered is the CRC16 over the first 98 bit of the EHEADER. If this check fails, the EHEADER must be considered to be corrupted. No information on DataField Length (DFL) is available for this turbo code packet, therefore it can not be derived how many MPEG-TS packets had been encapsulated. No information can be forwarded to the link layer.
- Up to 8 following CRCs16 have to be evaluated next, if the CRC16 over the EHEADER is correct and the SYNC byte has the expected value 47_H. The DFL indicates the number of MPEG-TS packets in the DATAFIELD section. Over each packet of 187 bytes (UP), the CRC16 is calculated. The following applies:
  - if the CRC16 is wrong, the transport_error_indicator has to be set to 1, and the packet has to be delivered to the link layer processing;
  - if the CRC16 is correct, the transport_error_indicator remains untouched (it may have been set already by any other TS handler on the transmit side).

7.2.2.5.2 MPEG-TS packet format

The format of the MPEG-TS transport packet is given in table 7.3.

Table 7.3: MPEG-TS transport packet format

Syntax	No. of bits	Mnemonic
transport_packet(){		
sync_byte	**8**	**bslbf**
transport_error_indicator	**1**	**bslbf**
payload_unit_start_indicator	**1**	**bslbf**
transport_priority	**1**	**bslbf**
PID	**13**	**uimsbf**
transport_scrambling_control	**2**	**bslbf**
adaptation_field_control	**2**	**bslbf**
continuity_counter	**4**	**uimsbf**
if(adaptation_field_control == '10' || adaptation_field_control == '11'){		
adaptation_field()		
}		
if(adaptation_field_control == '01' || adaptation_field_control == '11') {		
for (i = 0; i < N; i++){		
data_byte	**8**	**bslbf**
}		
}		
}		

The relevant flag to be set in case of a CRC16 failure on the user packet (UP) is the transport_error_indicator.

7.2.2.5.3 Generation of MPEG-TS null-Packet

In case of a non-recoverable EHEADER, it is proposed to transmit a so-called MPEG-TS null packet. Although indicated by the name, the payload of a null-packet is not the all-zero sequence but defined as follows:

Byte[0]	Byte[1]	Byte[2]	Byte[3]	Byte[4]	Byte[5] …	Byte[187]
47_H	$1F_H$	FF_H	10_H	FF_H	FF_H …	FF_H

Please note that all bytes between Byte [4] and Byte[187] are set to FF_H.

7.2.2.5.4 Signalling of wrong MPEG-TS packet

In case of an erroneous CRC16 over one of the user packets, the transport_error_indicator has to be set to 1. This is done by an OR-combination of the value 80_H with the MSB (most significant bit) of Byte1. This helps to preserve any transport_error_indicator that has been set along the transmission chain. The way to set the transport_error_indicator is depicted in figure 7.7.

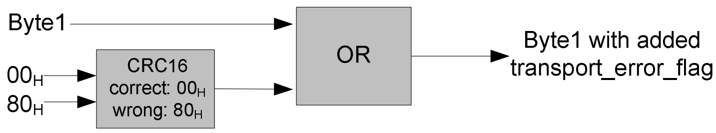


Figure 7.7: Setting the transport_error_indicator flag

7.2.2.6 C/N performance values

7.2.2.6.1 Ideal performance in AWGN channel

An ideal receiver should have the theoretical performance given in table 7.4 for OFDM and table 7.5 for TDM. The values exclude any LL-FEC. An ideal transmitter is assumed, the noise bandwidth is set to 4,8 MHz.

Pilot overhead in C/N is only taken into account what concerns the boosting of pilots in OFDM (0,3 dB) which reduces the Es/N0 of the data carriers with respect to the C/N. Any other overhead (signalling, guard intervals) is addressed in the spectral efficiency curves.

Table 7.4: Theoretical C/N (dB) in AWGN channel for OFDM @ BER = 10^{-5}

Code rate	QPSK	16QAM
1/5	-3,6	0,7
2/9	-3,1	1,3
1/4	-2,5	1,9
2/7	-1,8	2,8
1/3	-0,9	3,7
2/5	0,1	5,0
1/2	1,4	6,8
2/3	3,5	9,7

Table 7.5: Theoretical C/N (dB) in AWGN channel for TDM @ BER = 10^{-5}

Code rate	QPSK	8PSK	16APSK
1/5	-3,9	-1,3	0,4
2/9	-3,4	-0,7	1,0
1/4	-2,8	-0,1	1,6
2/7	-2,1	0,7	2,5
1/3	-1,2	1,6	3,4
2/5	-0,2	2,7	4,7
1/2	1,1	4,4	6,5
2/3	3,2	6,9	9,4

7.2.2.6.2 Ideal performance in Rice channel

Only results for TDM (QPSK) are available and listed in the following table. The simulation assumes perfect interleaving/deinterleaving (i.e. uncorrelated fading), the Rice factor used is 3 dB.

Table 7.6: Theoretical C/N (dB) in Rice channel (K=3 dB) for TDM @ BER = 10^{-5}

Code rate	QPSK
1/5	-3,4
1/4	-2,2
1/3	-0,4
1/2	2,2

7.2.2.6.3 Ideal performance in Rayleigh channel

Only results for TDM (QPSK) are available and listed in the following table. The simulation assumes perfect interleaving, the Rice factor used is minus infinity dB (equivalent to Rayleigh distribution).

Table 7.7: Theoretical C/N (dB) in Rayleigh channel for TDM @ BER = 10^{-5}

Code rate	QPSK
1/5	-3,2
1/4	-2,1
1/3	-0,2
1/2	2,9

7.2.2.6.4 Ideal performance in burst erasure channel

At this point in time, only results for OFDM (QPSK) are available. Table 7.8 gives an example for the possible performance for the burst erasure channel. This channel removes parts of the redundancy directly at the input of the deinterleaver. One modulation (QPSK) and one code rate (1/3) has been selected to demonstrate the performance of the turbo code. Each turbo code word of rate 1/3 consists of 18 capacity units, which equals 288 interleaver units.

The table has to be read as follows:

- the first and second rows show the selected erasure rate;
- the third row calculates the remaining effective code rate, i.e. the ratio of transmitted information bits over the received (non-erased) code bits: code rate / (1-erasure rate), after the applied burst erasure;
- the fourth row displays the minimum C/N necessary to reach a word error rate (WER) of 10^{-3};
- the fifth row displays the additional C/N necessary compared to AWGN performance (= no erasures).

Table 7.8: Performance of turbo codes on burst erasure channel

Erased interleaver units	0/288	24/288	48/288	72/288	96/288	120/288	144/288
Erasure rate in per cent	0 %	8,3 %	16,6 %	25 %	33,3 %	41,6 %	50 %
Effective Code rate	0,333	0,3636	0,4000	0,4444	0,5000	0,5714	0,6667
Minimum C/N req. for WER=10^{-3}	-0,9 dB	0,0 dB	0,6 dB	1,4 dB	1,8 dB	3,0 dB	4,4 dB
Additional C/N req. compared to AWGN	0,0 dB	0,9 dB	1,5 dB	2,3 dB	2,7 dB	3,9 dB	5,3 dB

This table is a good indicator to test receiver performance in AWGN and blockage channel. For AWGN channel, the actual required C/N to achieve a certain WER results in the implementation loss achieved, whereas the block erasure channel allows to derive the implementation loss in burst fading channels.

The latter also helps to evaluate the performance of the time interleaver described in clause 7.2.3. The interleaver profiles are designed such that - depending on the current reception condition and/or receiver memory - not all interleaver units are necessary to successfully decode the turbo code word. This allows to rapidly decode data on the physical layer in case of receiver switch-on or change of service, without waiting for a full interleaver length before starting to decode. The same argument can be used to evaluate the performance degradation of memory-limited receivers that are forced to drop part of the transmitted interleaver units.

The performance of the receiver in this burst erasure channel scenario is crucial for the switch-on time (zapping time) of the receiver; the expected behaviour is explained further in details in clause 7.2.3.5.4.

7.2.3 Time interleaver

7.2.3.1 Introduction

Time interleaving in general helps to counteract the effects of signal fading on the transmission channel. Reception instants with bad signal quality can be counteracted with good signal reception.

Interleaving always has to be aligned with the assigned code rate and modulation scheme used for transmission, as well as with the expected reception field strength:

- if a system is working with relatively low margins (difference between received and required field strength for error-free reception is lower than 4 dB), interleaving over a long time span (e.g. 10 s) may worsen the reception as it spreads short errors over a long time span. See the top drawings of figure 7.8 for details;
- if indeed a system works with higher margins (difference between received and required field strength for error-free reception is higher than 7 dB), interleaving over a long time span (e.g. 10 s) can strongly improve the reception as it is able to recover many bad reception conditions. See the bottom drawings of figure 7.8 for details.

Interleaving can be seen as time-averaging of the received C/N. It averages out good and bad reception conditions within the length of the interleaver. This explanation is of course a bit simplified, as the C/N does typically not follow strict on-off characteristics, but has to be described with more parameters like typical fading durations, their occurrence probability and the varying C/N due to small-scale fading effects. These second order statistics are reflected by the channel models applied in the clause A.7. As shown in the previous clause, the FEC is designed to cope with a certain erasure percentage, dependent on the code rate selected for the physical layer FEC.

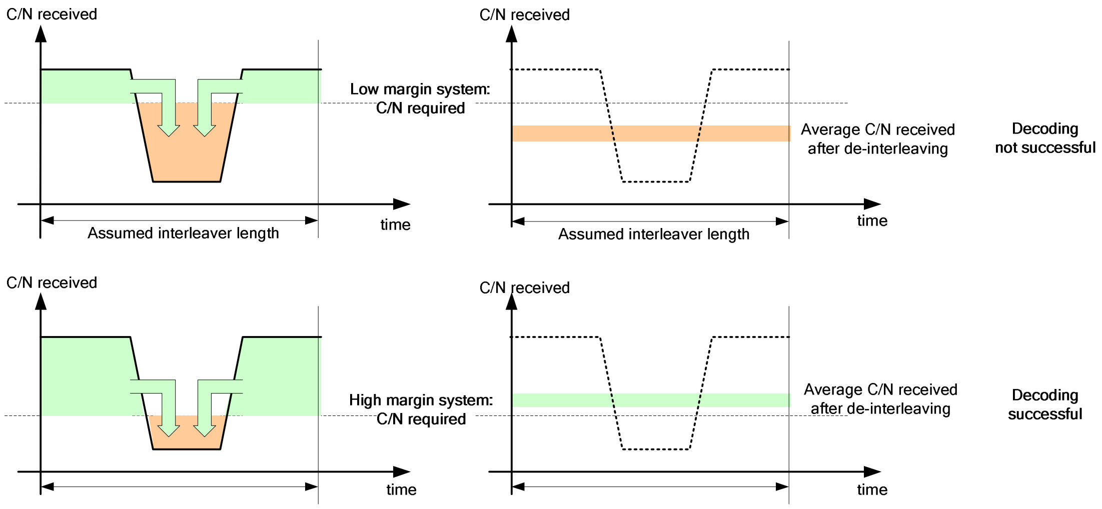


Figure 7.8: Interleaving with different margins; top: not sufficient margin; bottom: sufficient margin

Physical layer interleaving (which is addressed here) makes use of the fact that one turbo code word is cut into various pieces (up to 48, identical to the number of interleaver taps) and spread over many distinct time instants. In the receiver, these pieces are again collected, reordered and forwarded to the turbo decoder.

Dependent on the capabilities of the receiver (refer to the receiver Class definition), a decision has to be taken whether the receiver is able or not to successfully deinterleave the selected SH signal. Details on typical interleaver profiles are given in clause 7.2.3.3.4, deinterleaving strategies are given in clause 7.2.3.4.3, and details on the compatibility of the receiver with the transmitted signal is given in clause 7.2.3.4.

7.2.3.2 Selection of receiver classes

7.2.3.2.1 Overview

Two different types of receivers are addressed in TS 102 585 [2]:

- Class 1 receivers can work with a number of interleaver units up to 6 528, each of 126 bits, which is the equivalent to 1/2 (half) of the SH-frame capacity. Two types of interleaver configurations in the transmit signal are compatible with this type of receiver:
  - short uniform interleaver;
  - the late burst of a long uniform-late interleaver;
  - the late burst of an early-late interleaver.
- Class 2 receivers can work with a number of interleaver units up to 417 792, each of 126 bits, which is the equivalent to the capacity of 32 SH-frames. All types of interleaver configurations in the transmit signal (which are within the range of the memory storage capabilities of the receiver) are compatible with this type of receiver:
  - short uniform interleaver;
  - long uniform-late interleaver;
  - long uniform interleaver;
  - long early-late interleaver.

7.2.3.2.2 Class 1 / Class 2 compatible interleaver profile selection

As discussed within clause 7.2.2.6.4 0, Class 1 receivers may still operate with interleaver profiles intended primarily for Class 2 receivers. Therefore, it was introduced as a mandatory requirement that any receiver not only derives the full list of interleaver taps, but cross-checks this with its own memory and decoding capabilities. The expected behaviour of any terminal (class 1 / class 2) is the following:

- calculate all interleaver taps as signalled in either TPS (OFDM) or signalling field (TDM);
- drop `k` taps which exceed the memory capabilities of this receiver type and keep only `L[0]` to `L[47-k]`;
- check whether the remaining code rate is still decodable (remaining code rate < 1,0);
- decode the remaining fragment of turbo code word if both the memory and the remaining code rate allow this.

One example for the performance penalty of such a configuration has been given in table 7.8. Assigning e.g. a late burst contribution of 50 % of all taps (*nof_late_taps* = 24), which can all be decoded by a class 1 receiver, the performance loss would be 5,3 dB (see clause 7.2.2.6.4) for an effective code rate of 2/3 instead of a total code rate of 1/3. These values may differ for the selection of other code rates and/or assignment of taps to the late part, and depend on the channel conditions.

The benefit of such a network configuration is that both class 1 and class 2 receivers can co-exist in a network primarily intended for class 2 receivers, allowing also a soft-transition between the classes.

7.2.3.3 Interleaver profile description using the waveform parameters

7.2.3.3.1 Overview

For all transmitted signals in a DVB-SH system, the interleaver parameters can be chosen and signalled independently. Certain rules apply when setting the interleaver parameters for signals that are intended to be combined in the receiver prior to FEC decoding, see clause 7.2.2.3.3 for details.

The difference in the signalling also results from the different SH frame lengths in OFDM and TDM mode, which are defined in multiple of CU (capacity units) and IU (interleaver units) and not in units of time. Defining identical interleaver profiles for OFDM and TDM will typically lead to slightly different interleaver length (in time). However, the maximum memory capabilities of the receiver classes have to be respected.

The interleaver parameters are always defined from the receivers perspective, giving clear guidance on the selection of parameters for the de-interleaving process.

Long physical layer interleaver contradicts with variable burst sizes and distances of DVB-H when power saving is addressed. However, mapping techniques of DVB-H over DVB-SH can partially compensate the contradiction. In general, time slicing (as introduced by DVB-H and preserved in DVB-SH) and long time interleaving (as introduced within DVB-SH) seem to be contradictory:

- *time slicing* tries to concentrate the data to reduce the on-time of a receiver, thus lowering significantly the power consumption;
- *time interleaving* tries to spread the data over a large amount of time to benefit from time-variant channels.

The tools provided in DVB-SH provide the necessary flexibility in the choice of time slicing and time interleaving jointly. Therefore, the concept of *SH-service* was introduced, signalled by the *service_synchronization_function*. Each SH-service has a constant repetition period RP_{SH} and may cover a number of time-sliced services. These SH-services are then fed into the time interleaver who's repetition period RP_{TI} should be identical to the repetition period of the SH-service.

From the transmitter's perspective, this categorization of services looks like *n* virtual transmission channels with reduced throughput. By selecting a reasonable number of SH-services, this SH-service category aggregates in the transmitter:

- a small number (2 to 3) of DVB-H services within one SH-service.

By choosing 1 out of *n* present SH-services, this SH-service category allows the receiver:

- to decode only parts of the whole multiplex (reducing significantly the throughput requirements and deinterleaver memory necessary)
- to benefit from time slicing on an SH-service by SH-service basis (less but still considerable power savings when compared to DVB-H time slicing mode)

Figure 7.9 displays the relationship between time slicing services and SH-services, between SH-services and SH frames, and the output of the time interleaver. The mapping and parameter selection is the following:

- n_1 DVB-H services map onto *one* SH-service; n_1 may be different for each SH-service;
- *One* SH-service maps onto n_2 EEFRAMEs; n_2 may be different for each SH-service;
- The repetition period RP_{SH} for each SH-service is **identical** in the number n_3 of SH frames;
- The time interleaver works with the repetition period RP_{TI}. The repetition period is given in SH frames with the following condition: $RP_{TI} = RP_{SH}$ = slice_distance (with slice_distance being defined in the next clause).

The usage of the parameter set as described above ensures that features like VBR support (at the transmitter) and time slicing (at the receiver) can be, at least partially, exploited jointly with (long) physical layer interleaving.

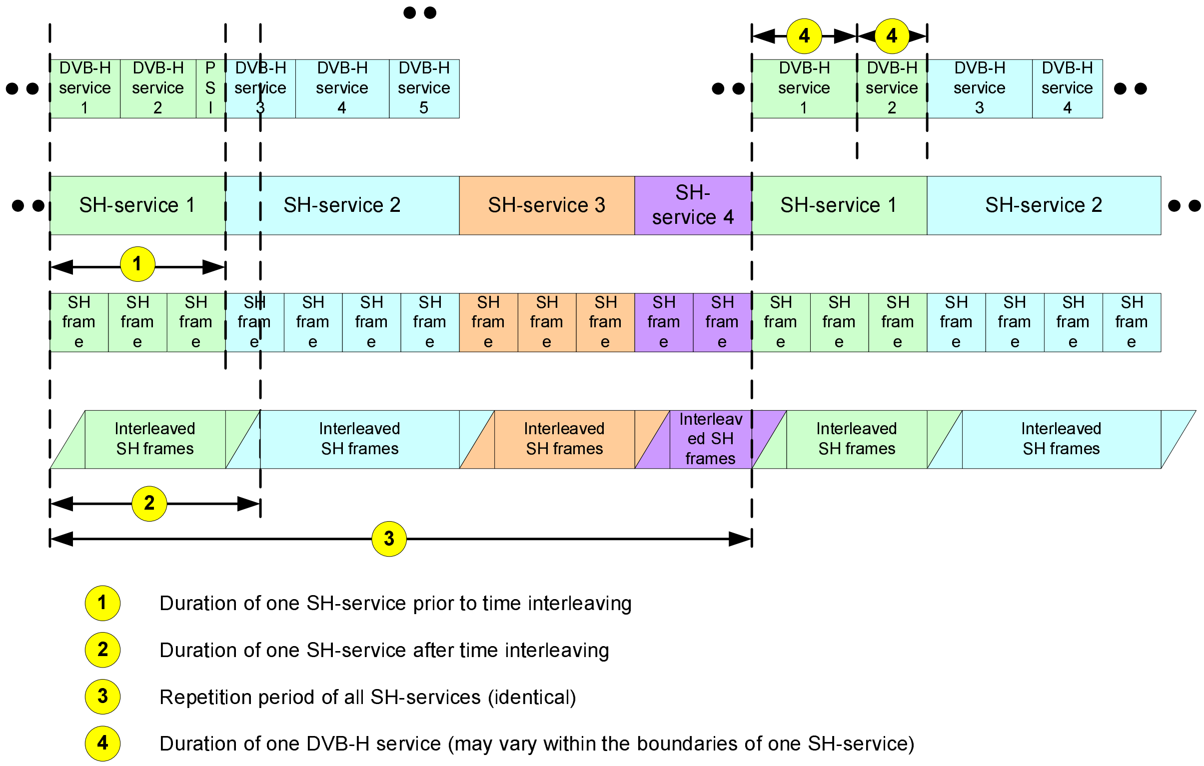


Figure 7.9: Timing relationships between Time-Slice, SH-Service and SH-frame, prior to and after interleaver

7.2.3.3.2 Parameters and description

The waveform document [1] specifies the following table of parameters for the description of the interleaver.

- **Common_multiplier:** this parameter is a multiplier of the number of interleaver units for all interleaver taps *within one slice*. The minimum value is 1, the maximum value is 63.
- **Nof_late_taps:** this parameter partitions the interleaver into a *late* and a *non-late* part. The border between the two different parts is somewhere between 0 (no late part available) and 48 (only late part available).
- **Nof_slices:** this parameter defines the number of interleaver slices over which the interleaver distributes the data. This parameter has to be set together with the next parameter slice_distance to determine the interleaver length in time, and has to be aligned to time slicing as described in clause 7.2.3.3.1. In case of selection of a late part only, this parameter is set to 1, and all subsequent parameters can be set to 0.
- **Slice_distance:** this parameter defines the distance between two interleaver slices in SH-frames. The capacity (in CU) of an SH-frame depends on the modulation type, modulation order (TDM), and roll-off (TDM) selection. To transform this value in an increment in IU, the parameter `nof_Data_CU` has to be calculated according the following clause.
- **Non-late increment:** this parameter is an additional multiplier for all interleaver taps *within all non-late slices*. It has to be multiplied with the common_multiplier to determine the increment in all slices which are *not* the late part of the interleaver.

In order to ease modulator design, it is recommended to have the following constrains applied to the selection of the common_multiplier:

- using OFDM with 16QAM modulation, the common_multiplier must be a multiple of 2 (in 4 k modes) or 4 (in 8 k modes); for all other modes, no constraints apply;
- using OFDM with QPSK modulation, the common_multiplier must be a multiple of 2 (in 8 k modes); for all other modes, no constraints apply.

7.2.3.3.3 Calculation examples

Figure 7.10 gives an overview on how to derive exactly the tap delays for the time interleaver.

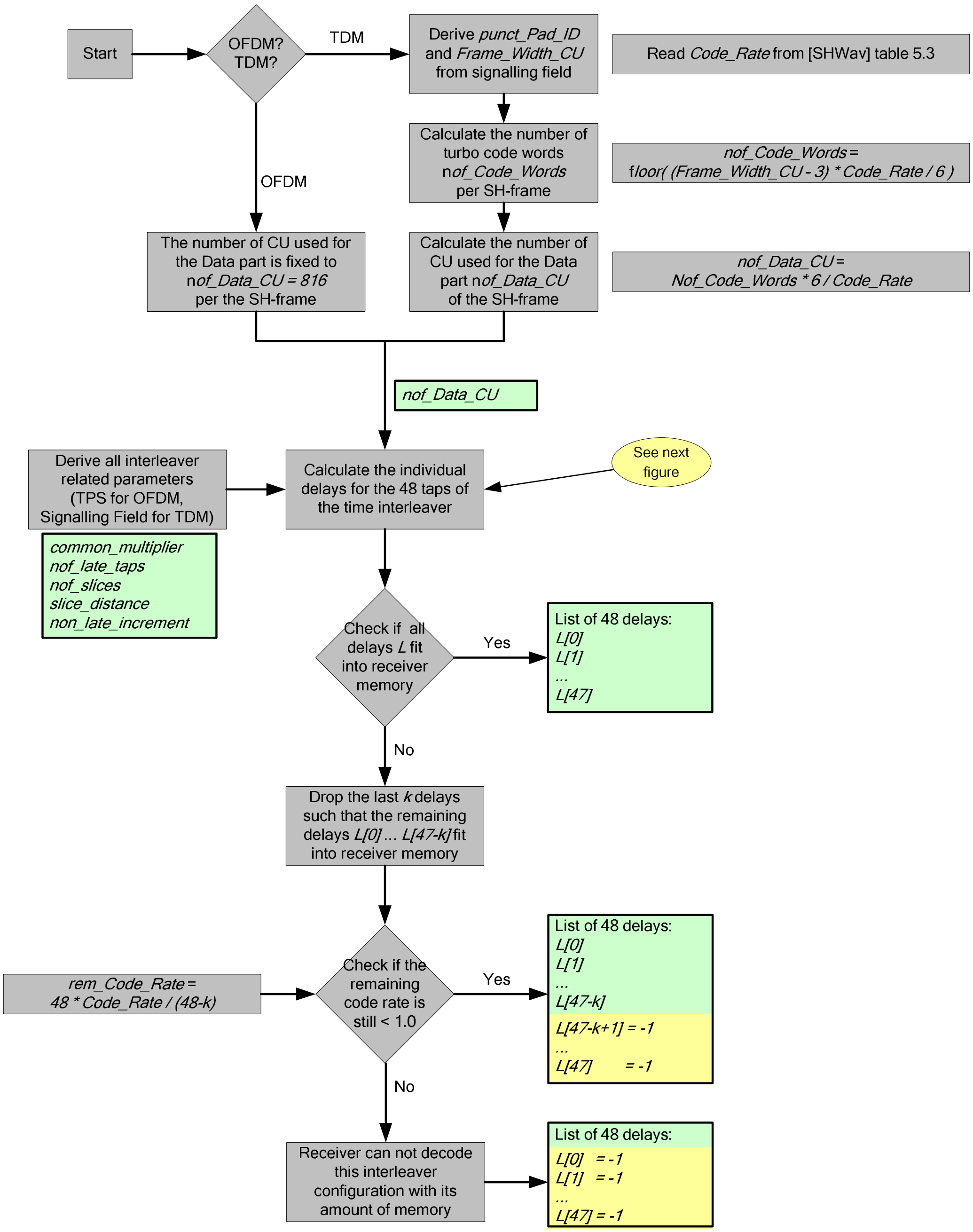


Figure 7.10: Process description for deriving the interleaver parameters

Figure 7.11 explains the derivation of the list of the 48 interleaver delays *L[0]* to *L[47]* from the parameters as derived above.

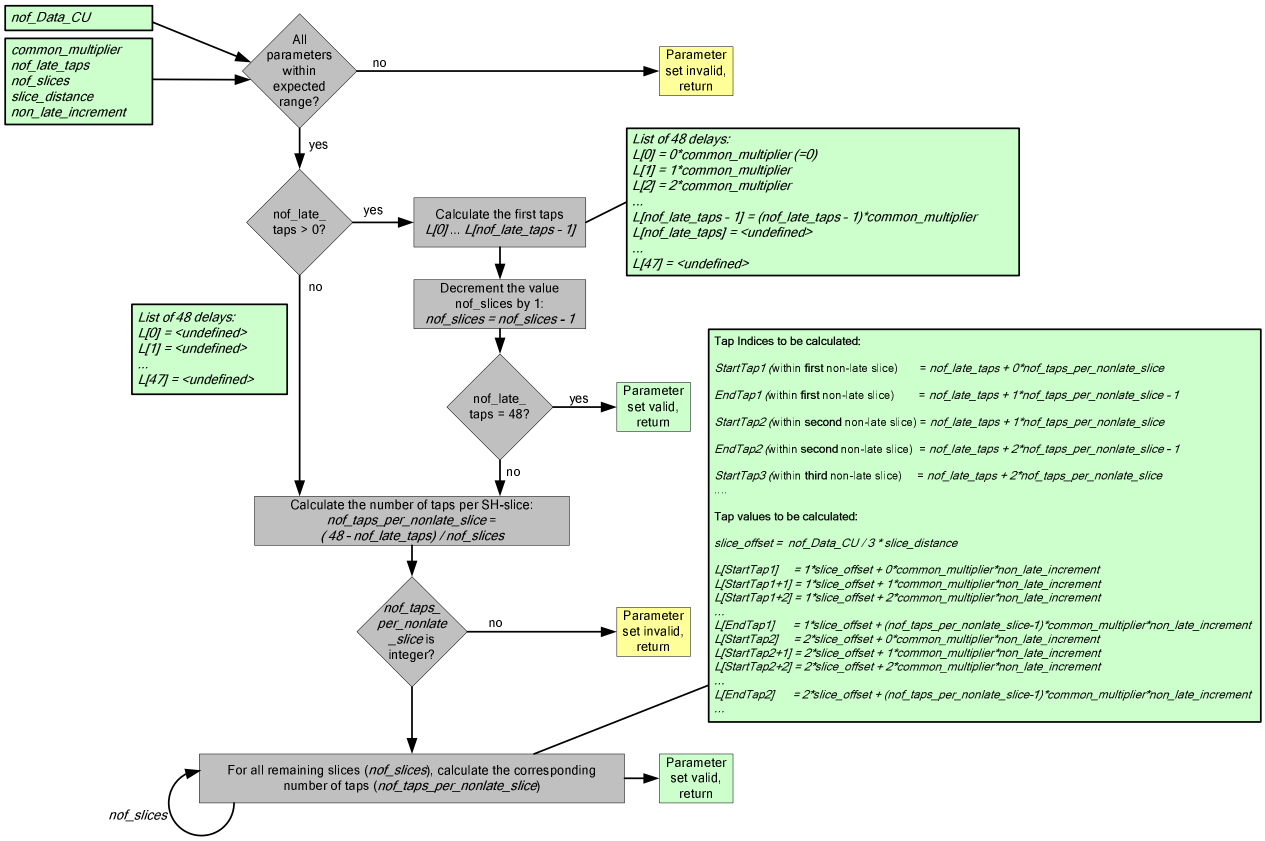


Figure: 7.11: Process description for calculation of interleaver parameters

7.2.3.3.4 Typical interleaver profiles

In this clause, typical interleaver profiles are presented. These interleaver profiles have been evaluated to be compliant to the receiver classes as mentioned above and taken as reference for simulation and evaluation. In clause A.4, the performance characteristics for these profiles are given over the reference channels selected.

The support of time slicing is incorporated into the interleaver profiles, which can be parameterized according to the rules applied for the generation of time slicing on the link layer. To preserve the power-saving effect of time slicing, the parameters for a long interleaver and time slicing need to be selected jointly, otherwise reduction in the gain of power-saving through time-slicing may occur.

Short uniform interleaver profile (reference Terr48)

The *short uniform* interleaver profile is intended for Class 1 receivers capable of saving the equivalent of 1/2 (half) SH-frame in OFDM mode. The resulting maximum interleaver length for this *short uniform* interleaver profile is therefore 1 (one) SH-frame in OFDM mode.

Depending on modulation order, channel bandwidth and guard interval selection, this is equivalent to time spans between 100 ms (bandwidth 5 MHz, guard interval 1/32, modulation 16QAM) and 240 ms (bandwidth 5 MHz, guard interval 1/4, modulation QPSK).

The *short uniform* interleaver profile used to derive simulation results is defined in clause A.4. A view of this interleaver path is given in figure 7.12.

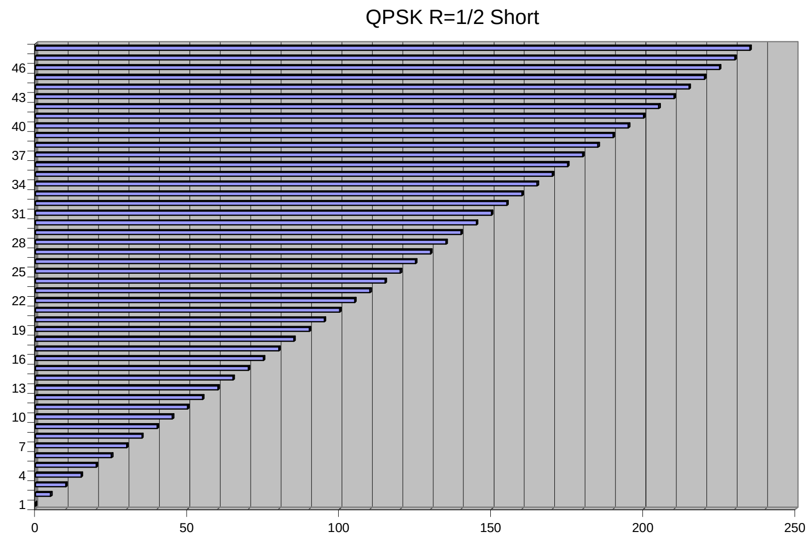


Figure: 7.12: Interleaver delays for taps L[0] to L[47] (short uniform)

Long uniform interleaver profile (reference Uni10)

The *long uniform* interleaver profile is intended for Class 2 receivers capable of saving the equivalent of 32 SH-frames in OFDM mode. The resulting maximum interleaver length for this *long uniform* interleaver profile is therefore in the range of 64 SH-Frames in OFDM mode.

Depending on modulation order, channel bandwidth and guard interval selection, this is equivalent to time spans between 6,4 s (bandwidth 5 MHz, guard interval 1/32, modulation 16QAM) and 15,6 s (bandwidth 5 MHz, guard interval 1/4, modulation QPSK).

The *long uniform* interleaver profile used to derive simulation results is defined in clause A.4. A view of this interleaver path is given in figure 7.13.

The use of MPE-IFEC "on top" of the *long uniform* interleaver profile is not recommended, but not strictly forbidden as both classes of terminals (Class 1 and Class 2) have to support MPE-IFEC.

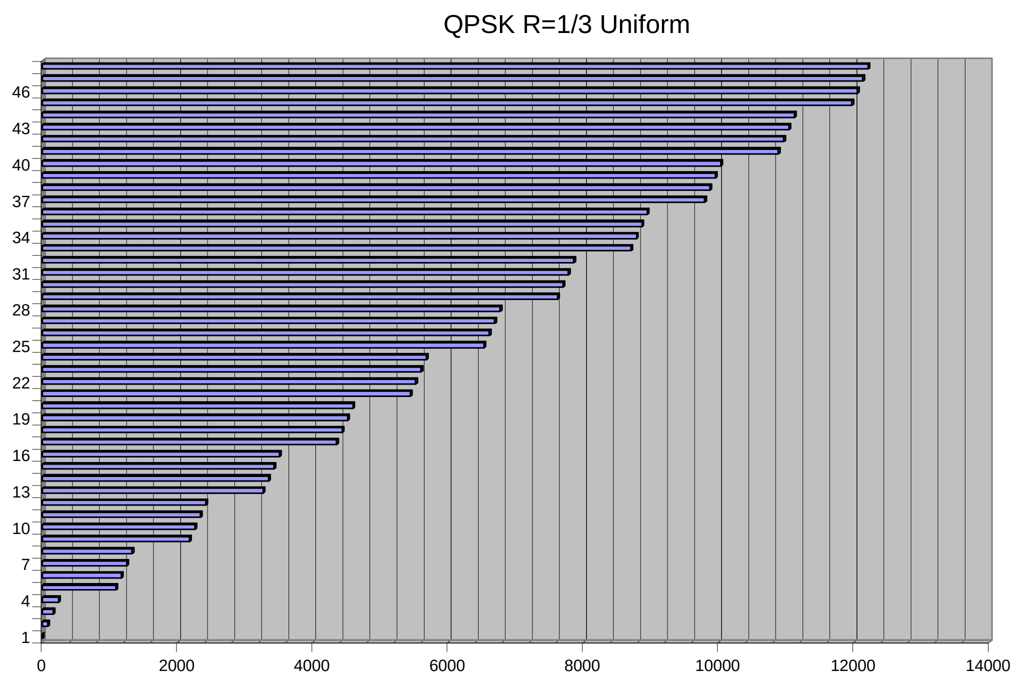


Figure 7.13: Interleaver delays for taps L[0] to L[47] (long uniform)

Long uniform-late interleaver profile (reference ULate10)

The *long uniform-late* interleaver profile is intended for Class 2 receivers capable of saving the equivalent of 32 SH-frames in OFDM mode. The resulting maximum interleaver length for this *long uniform-late* interleaver profile is typically larger than for the *long uniform* interleaver profile, and can be in the range of up to 128 SH-Frames in OFDM mode.

One special case occurs when considering Class 1 terminals in networks transmitting signals intended primarily for Class 2 terminals. Depending on the detailed parameter selection, a Class 1 terminal may still - with penalty in mobile performance - be able to successfully decode the received signal, even when the full diversity is not exploited. The recommended Class 1 receiver behaviour is explained in clause 7.2.3.2.2. A proper parameter selection of this *long uniform-late* interleaver profile is mandatory for this coexistence to work.

Depending on modulation order, channel bandwidth and guard interval selection, this is equivalent to time spans between 12,8 s (bandwidth 5 MHz, guard interval 1/32, modulation 16QAM) and 31,2 s (bandwidth 5 MHz, guard interval 1/4, modulation QPSK).

The *long uniform-late* interleaver profile has been used to derive simulation results is defined in clause A.4. A view of this interleaver path is given in figure 7.14.

The use of MPE-IFEC "on top" of the *long uniform-late* interleaver profile is not recommended, but not strictly forbidden as both classes of terminals (Class 1 and Class 2) have to support MPE-IFEC.

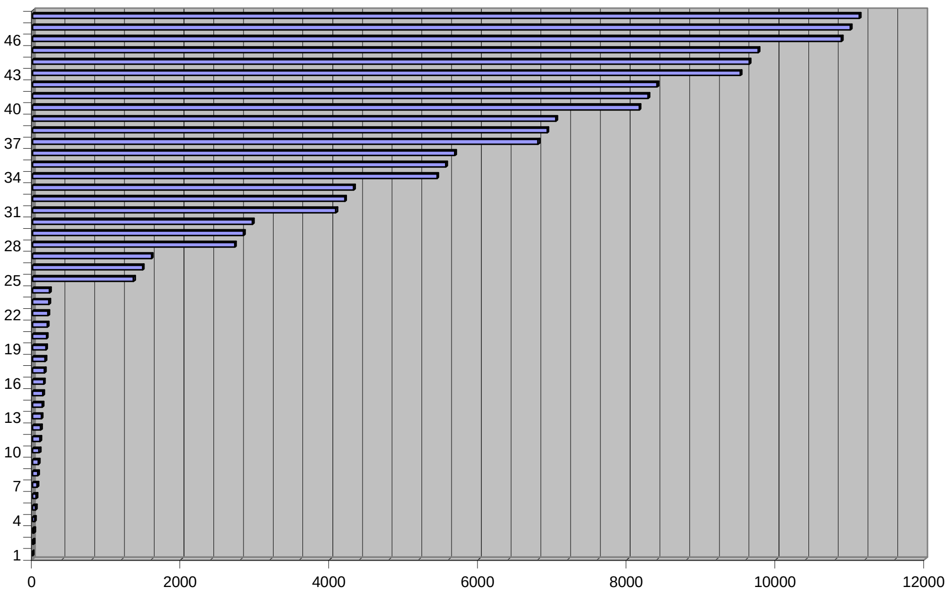


Figure 7.14: Interleaver delays for taps L[0] to L[47] (long uniform-late)

Long early-late interleaver profile

The *long uniform-late* interleaver profile is intended for Class 2 receivers capable of saving the equivalent of 32 SH-frames in OFDM mode. The resulting maximum interleaver length for this *long early-late* interleaver profile is typically larger than for the *long uniform* interleaver profile, and can be in the range of up to 128 SH-Frames in OFDM mode.

One special case occurs when considering Class 1 terminals in networks transmitting signals intended primarily for Class 2. Depending on the detailed parameter selection, a Class 1 terminal may still - with penalty in mobile performance - be able to successfully decode the received signal, even when the full diversity is not exploited. The recommended Class 1 receiver behaviour is explained in clause 7.2.3.2.2. A proper parameter selection of this *long early-late* interleaver profile is mandatory that this coexistence can work.

Depending on modulation order, channel bandwidth and guard interval selection, this is equivalent to time spans between 12,8 s (bandwidth 5 MHz, guard interval 1/32, modulation 16QAM) and 31,2 s (bandwidth 5 MHz, guard interval 1/4, modulation QPSK).

An example of the *long early-late* interleaver is shown in figure 7.15. The use of MPE-IFEC "on top" of the *long early-late* interleaver profile is not recommended, but also not forbidden as both classes of terminals (Class 1 and Class 2) have to support MPE-IFEC.

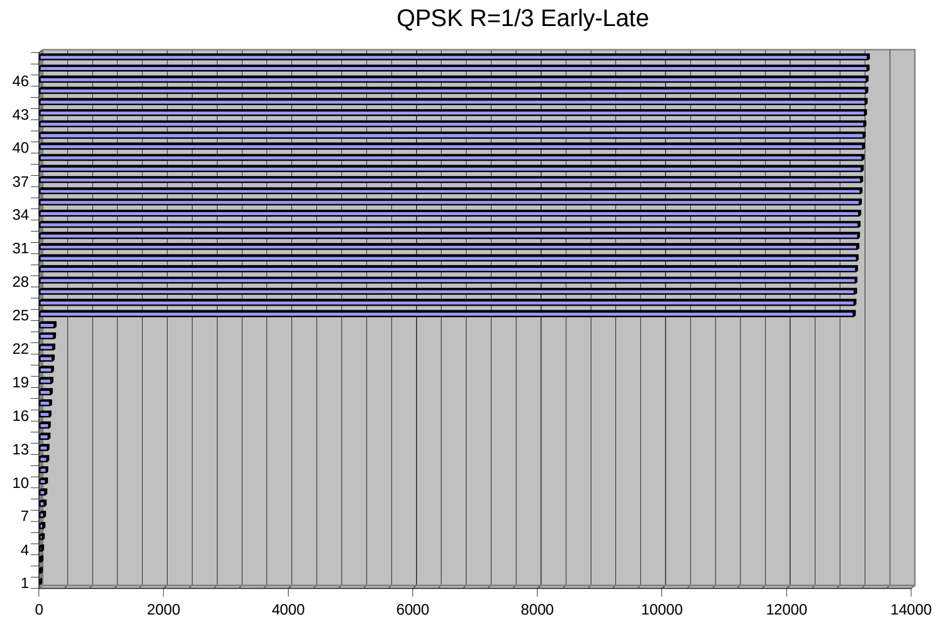


Figure 7.15: Interleaver delays for taps L[0] to L[47] (long early-late)

7.2.3.4 Interleaver profile alignment between satellite and terrestrial / different carriers

This clause is dedicated to give advice on the alignment of interleaver profiles between satellite and terrestrial transmission, if different carrier frequencies are used. This clause is not applicable to SH-A in SFN mode.

The synchronization aspects which are not interleaver-related are given in clause 7.5 for all types of DVB-SH networks. There, the following items are addressed:

- distribution network delays and the use of the SHIP packet for synchronization;
- compensation of the satellite round-trip-delay using the SHIP mechanisms.

7.2.3.4.1 Overview

Certain rules apply when interleaving parameters are selected differently for the satellite and the terrestrial transmission. This is possible in the following network configurations:

- SH-A in MFN mode;
- SH-B.

The interleaver profile selected will in most cases not be identical, due to the possibility to choose the parameters individually for both transmission paths in terms of:

- modulation and modulation order (OFDM/TDM, QPSK/8PSK/16QAM/16APSK);
- code rates (between 1/5 and 2/3);
- interleaver length (between 100 ms and 30 s);
- throughput.

To be able to combine both demodulated signals, and to minimize the penalty on the receiver in terms of memory consumption and de-interleaver memory, the transmit signals are aligned such that the difference in delay is compensated already in the transmitter. Figures 7.16 and 7.17 introduce the underlying concept.

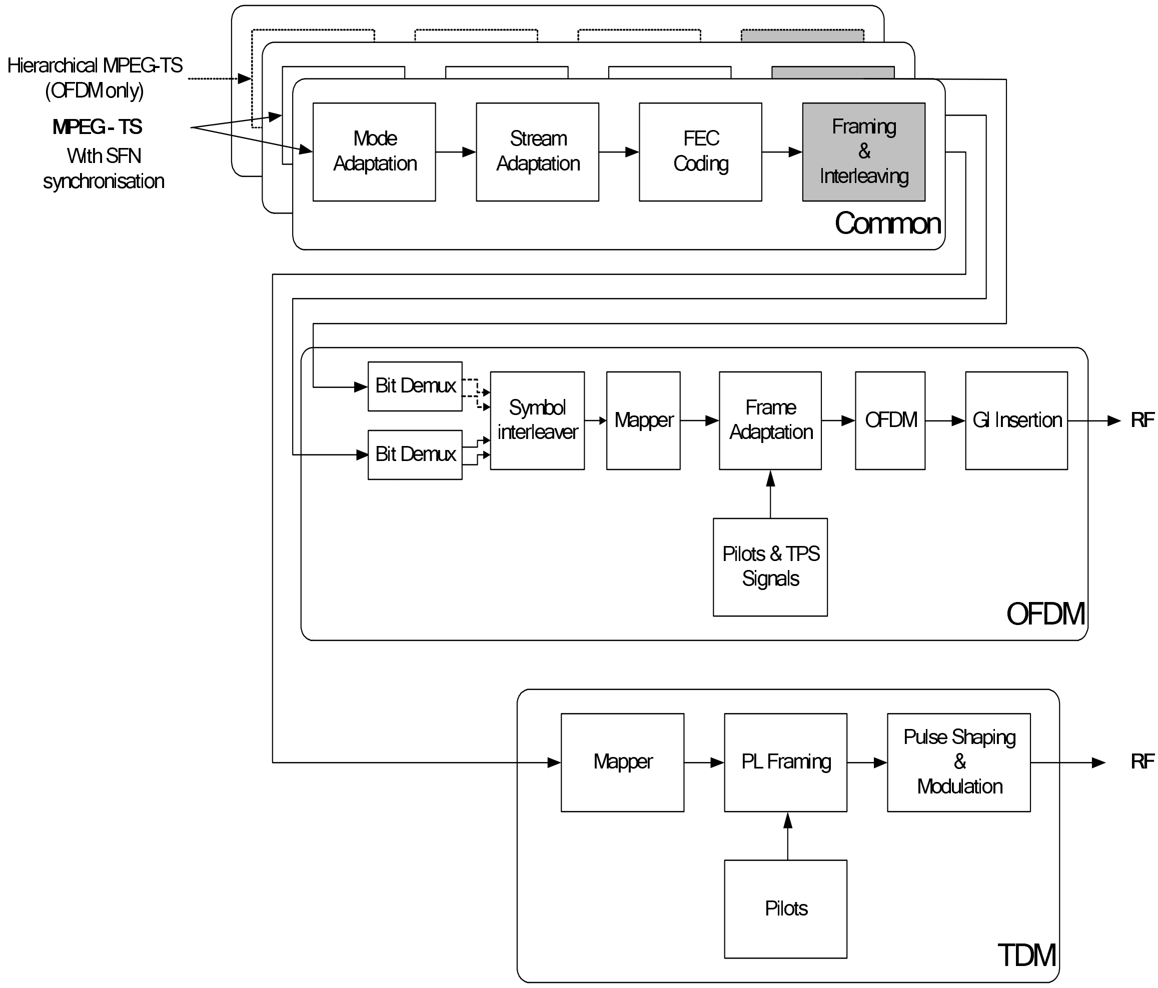


Figure 7.16: Reference system block diagram (physical layer)

The blocks marked in grey (framing and interleaving) have to be considered for each input stream individually.

To minimize the memory impact on the receiver, the transmitters are requested to align the transmission latency between all different time interleavers used in the DVB-SH system. Figure 7.17 shows the concept.

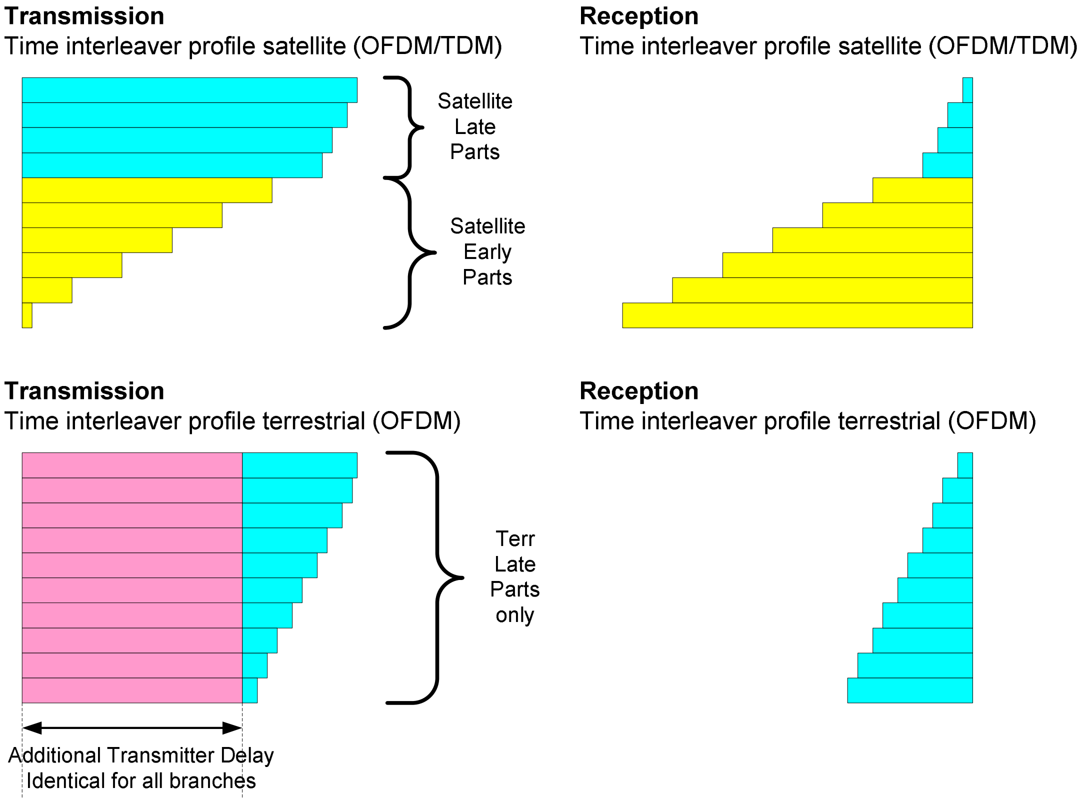


Figure 7.17: Deployment of interleaver profiles with different length

In this example, the "effective" time interleaving is shorter for the terrestrial part than for the satellite part (see both drawings on the right hand side, which displays the de-interleaver required on the receiver). The transmitter of the terrestrial signal therefore has to compensate the different end-to-end delays on the shorter transmission path, in this case the terrestrial one.

7.2.3.4.2 Interleaver synchronization on the transmitter side

If two different parameter sets exist within the hybrid DVB-SH network (as explained before), which work on the same MPEG-TS input stream and may therefore be combined by a receiver capable of signal combining, the end-to-end latency has to be compensated at the transmission side by the following means:

- Check if both the tps_ship and the tdm_function() of the SHIP packet [1] are used and valid.
- Calculate the latency (in SH-frames) of both time interleavers (OFDM/OFDM or OFDM/TDM).
- Calculate the difference of latency (in SH-frames) between both time interleavers.
- Compensate the difference of latency (in SH-frames) between the two time interleavers on the path with the smaller latency; this is performed extending the delay lines of the time interleaver.

To derive the difference in SH-frames, the interleaver configuration has to be evaluated according to the following rules. The granularity for this difference is one SH-frame. Any other possibility of deriving the delay between satellite and terrestrial (e.g. incorporating the fully calculated table of interleaver taps L[0] to L[47]) ***must not be used*** to avoid different interpretations (e.g. rounding effects) by different implementations.

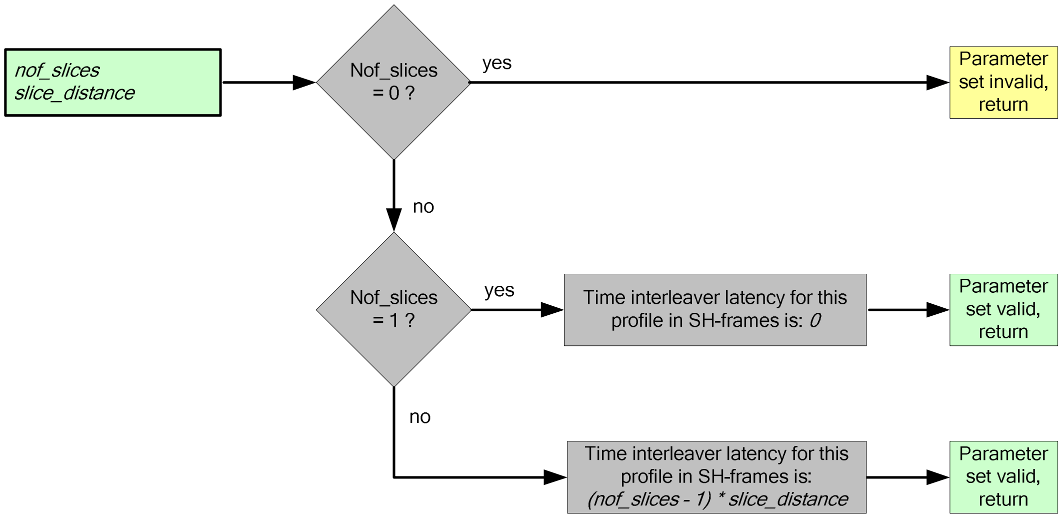


Figure 7.18: Rule to derive the time interleaver latency (in SH-frames)

As soon as the difference of time interleaver latencies is known, the time interleaver with the shorter latency adds to each delay line L[0] to L[47] the corresponding amount of interleaver units to compensate.

The number of interleaver units *per delay line* to add is derived as follows:

- derive the SH frame capacity in unit "capacity units" for the relevant time interleaver (see figure 7.10);
- derive the difference of interleaver latencies in SH frames (see above description);
- multiply these two values and divide by 3 (the latter is the conversion factor between capacity unit, interleaver unit and the 48 delay lines of the time interleaver).

7.2.3.4.3 Interleaver synchronization on the receiver side

This clause will be addressed in release 2. The following items will then be covered:

- frame acquisition for OFDM and TDM;
- code word acquisition;
- hybrid combining recommendations.

7.2.3.5 Fast interleaver synchronization strategies

7.2.3.5.1 Motivation

When long interleaver profiles are used, minimization of the access time to successfully decode a service is crucial. Independently whether it is the first access to a service ("zapping time") or the access to a service after a signal dropout ("recovery time"), it is preferable that the user gets fast feedback on whether the current location is suited for reception.

The choice of a proper decoding strategy is essential for streaming services which are protected by long interleavers. Independently whether the long interleaving is done on the physical or the link layer, different decoding strategies as presented in the following are possible.

7.2.3.5.2 Definition of jitter and delay

The major impact of the choice of decoding strategy is related with jitter and delay. The following definitions apply:

- **jitter:** variation of decoding instant or variation of decoder output rate;
- **delay:** time span between injection (transmitter side) and decoding (receiver side) of payload;

- **reduced diversity decoding:** makes use of the fact that the full diversity (redundancy) is not needed in all reception scenarios. As soon as the receiver has gathered sufficient packets to decode successfully, it starts the decoding process immediately. The decision is typically made on the effective code rate:
    - choose the net code rate (e.g. 1/3);
    - derive the ratio of packets received for one code word (e.g. 40 %);
    - calculate the effective code rate (60 % erased bits from code rate 1/3 correspond to effective code rate of 5/6);
    - check whether the receiver is capable of decoding (effective code rate < 1,0 or any other limit < 1,0).

7.2.3.5.3 Early decoding

Early decoding tries to minimize the end-to-end delay of a time-interleaved signal. Figure 7.19 shows the typical delays which are in many cases smaller than the full time span of packets.

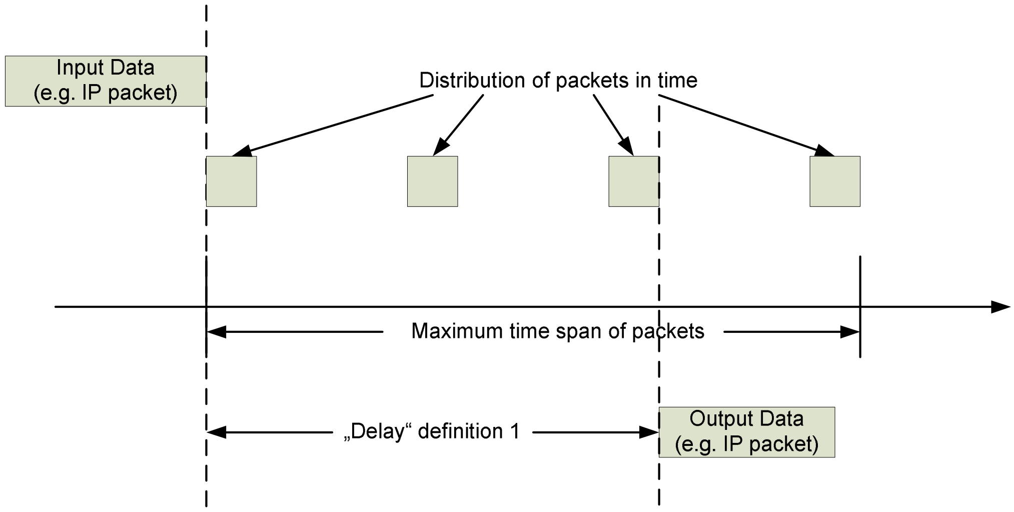


Figure 7.19: Principle of early decoding

However, assuming that not all injected packets are necessary for decoding purposes, the jitter at the output of the physical layer may increase as soon as the receiver runs into a channel situation that is worse than before: it needs now the full redundancy transmitted, such that for a certain amount of time, no or only very few packets will be processed.

The summary is the following:

- early decoding does not always need the full end-to-end delay;
- early decoding can provide short zapping times in good reception conditions;
- video rebuffering is necessary in fading environments.

7.2.3.5.4 Late decoding

Late decoding tries to minimize the jitter at the output of the physical layer decoder. It always waits for the full redundancy transmitted; the decoding instant is almost jitter-free. Figure 7.20 shows the constant delay which is the full time span of redundancy transmitted.

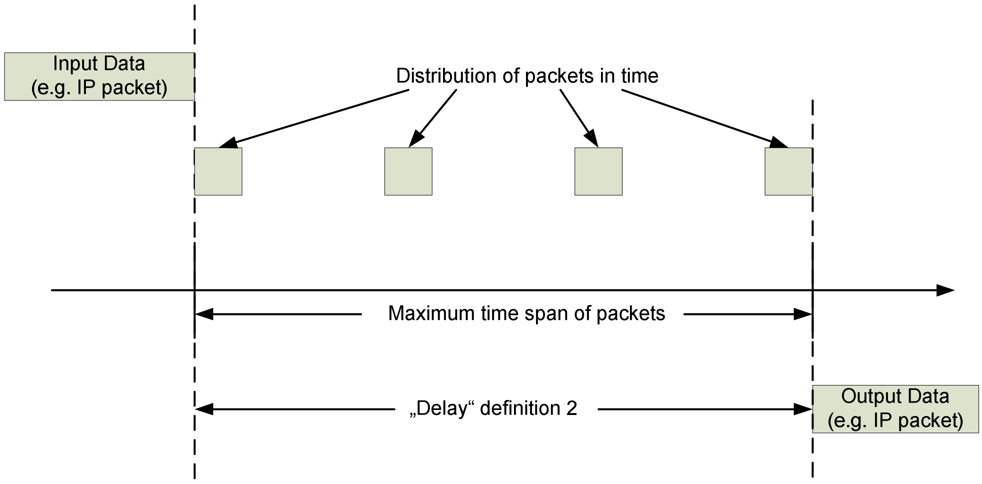


Figure 7.20: Principle of late decoding

The interleaver profiles defined in the waveform document is designed for the late decoding principle: it is possible to transmit a "late" burst which is delayed very long in the transmitter but has only a very short delay in the receiver.

When access is requested to a service which is interleaved over a long time span, this late burst enables the receiver to reconstruct the data transmitted without having received all parts before. Then, as time advances, more "uniform" parts are available, making the reception more robust, until the full interleaver length is reached and full diversity is achieved.

The summary is the following:

- late decoding allows jitter-free output of the physical layer;
- short zapping times can be achieved in good reception conditions;
- the waveform is designed to support late decoding;
- the full end-to-end delay is always experienced even in good reception conditions.

7.2.3.5.5 Zapping time in case of long physical layer interleaver

The use of low code rates provides a high coding gain, a high performance in case of fading and offers short zapping time. In theory only 1/R of the data has to be received to allow a decoding, if the received data are error free. If not yet all data of one code word are received this is equivalent to a higher code rate. If for example for a code rate of R=1/3 40 % of the interleaved data are received this can be considered as a code rate of R=5/6. For this code rate a higher C/N is required. If the available C/N is low a lower code rate may be required for error free decoding. In this case the receiver has to wait for the next burst to progressively lower the required C/N. The interleaver profiles are configurable. The Uniform-Late profile ensures that a big part of the encoded code word is available already with the first burst. In case of good reception conditions this part is already sufficient for successful decoding. The Uniform-Late interleaver profile is recommended, if a short zapping time is important. The late decoding principle ensures a jitter free decoding (jitter introduced by the multiplexing, variable bit rate coding and the time slicing are not considered here). The principle of the zapping behaviour of the physical layer is summarized in figure 7.21:

- an input burst is subdivided in several small bursts by the time interleaver. To simplify the drawing one burst is split in 6 short bursts;
- in case of uniform-late profile 50 % of the data are send as "late burst". In the example this is equivalent to 3 burst are spread over a long time, whereas the remaining 3 bursts are send a one bigger burst;
- from a FEC coding point of view the partitioning of the data in short burst is equivalent sending complementary punctured subsets of the Turbo encoder output. Assuming a code rate of R=1/3 the subset send as late burst is equivalent to a code word encoded with a code rate of R=2/3;

- if the receiver is tuned to the channel at the time t0 for the burst n the late burst is available at the time t1. This defines the zapping time for good reception conditions. The remaining data are considered as "erasures". According to table 7.8, 50 % erasure are allowed if the C/N is 5,3 dB above the threshold. This is equivalent to a nearly instantaneous zapping. In this case the latency caused by time slicing and the AVC encoder is dominating;

- in bad reception conditions a lower code rate is required to decode the data. If for example 4 (3 are transmitted as late burst, 1 out of 3 from the uniform part is received) are received the equivalent code rate is R=1/2 or the not yet received data are considered as "33 % of the data are erasure". According to table 7.8 the required additional C/N is 2,7 dB;

- for very bad reception condition when all of the transmitted redundancy is required to receive the data the zapping time may be up to the full interleaver length.
  Please note: If all redundancy is assigned to the Turbo code the physical layer uses a lower code rate resulting in a higher receiver sensitivity compared to a configuration where the parts of the redundancy are assigned to the IFEC.

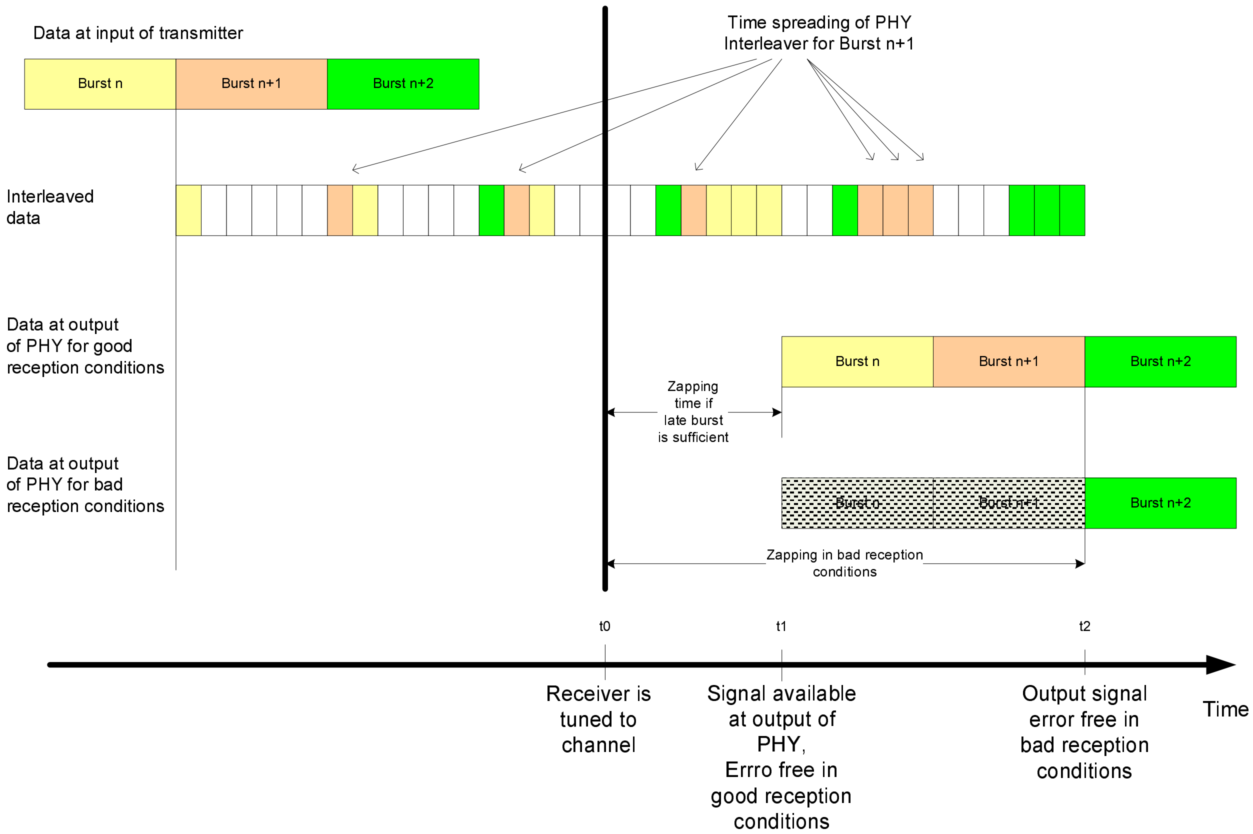


Figure 7.21: Principle of fast zapping

The numbers used for the example should be compared to the link budget. Typically a link margin (difference between available C/N for LOS reception and required C/N for AWGN channel) in the range of 6 dB to 10 dB is recommended (see clause A.12). With the given numbers the following conclusions can be derived

- for LOS reception (very good reception condition) the available link margin is typically sufficient to support instantaneous zapping;

- if the receiver is in a bad reception condition when the channel is selected a longer zapping time may result;

- receiver offering a lower link margin (e.g. receivers with low gain antenna) main have a higher zapping time;

- for terrestrial reception the reception area can be split in three parts;

- area with high C/N: short zapping time is provided;

- area with medium/low C/N: a longer zapping time has to be accepted;
- area with very low C/N: if all redundancy is assigned to the physical layer (no IFEC is used) the receiver offers a higher sensitivity.

7.3 OFDM and TDM elements

7.3.1 OFDM Elements

7.3.1.1 Overview

The OFDM modulation part is a full reuse of the functions of DVB-H with reasonable extensions necessary for the application "DVB-SH". The main new element is the introduction of a 1 k-FFT mode suitable for all channelizations. Furthermore, the TPS signalling has been extended to support the full signalling necessary in DVB-SH.

All design guidelines given in the Implementation Guidelines for DVB-H [3] are still valid and are explicitly referred here.

The main difference in the use of the OFDM part with respect to DVB-H is its use also at lower spectral efficiencies, down to QPSK modulation with code rate 1/5. This imposes higher requirements on the demodulator which should be designed and tested to comply with low C/N, satellite large-scale fading channels and frequency-selective reception scenarios.

In the next clause, one example for the performance of OFDM in a mobile channel is given to demonstrate the interaction between demodulator and decoder. Further OFDM specific simulation results are given in clause A.11.

7.3.1.2 C/N performance in mobile TU6 Channel

The following graphs summarize the TU6 performance for various configurations. The simulation conditions and the results are taken from clause A.12. The plots are given as "capacity over C/N" for different waveform configurations:

- 16 QAM modulation with R={1/5, 1/4, 1/3} and different physical layer interleaver profiles;
- QPSK modulation with R={1/3, 1/2} and different physical layer interleaver profiles.

Figure 7.22 highlights the additional interleaver gain of long physical interleaver. For 50 kmph already a short interleaver is sufficient and a longer interleaver provides only a small additional gain, whereas for lower speed the gain is significant. IFEC

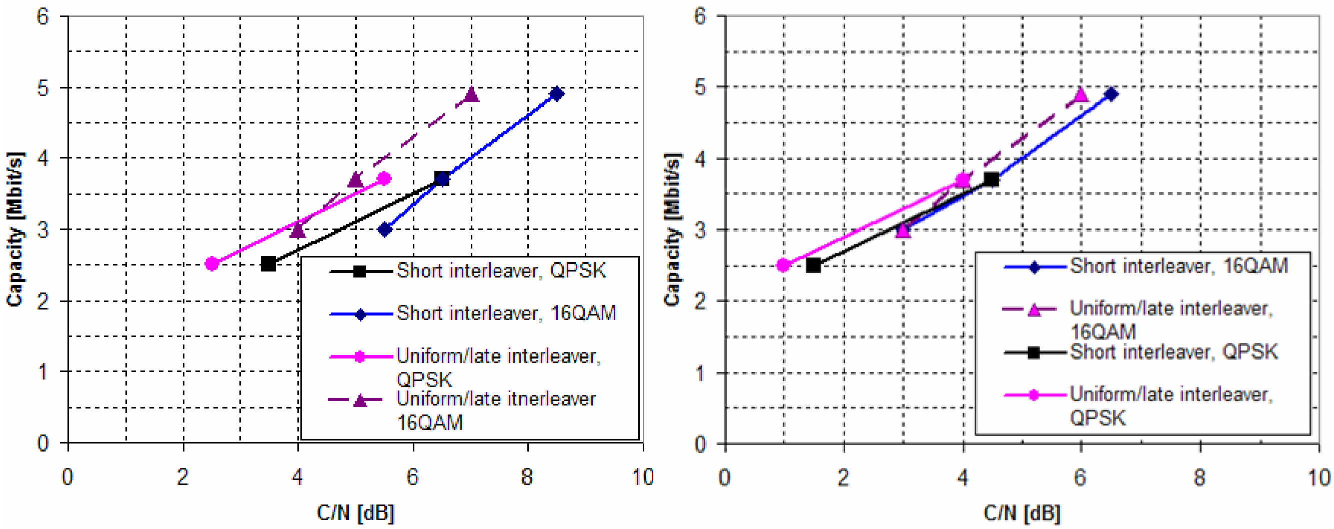


Figure 7.22: Capacity vs. C/N results for OFDM, TU6 channels, 3 kmph (left) and 50 kmph (right)

7.3.1.3 Doppler performance of QPSK modulation

The used reference receiver model characterizes a DVB-SH receiver performance in an ideal way using two numbers, C/N_{ain} and Fd_{3dB}. The C/N_{ain} gives the minimum required C/N for BER = 10^{-5}. The C/N-curve is flat up to high Doppler frequencies. Fd_{3dB} gives the Doppler frequency, where the C/N requirement has raised by 3 dB from the C/N_{ain} value.

Simulation results are given for QPSK and different code rates in the following.

This clause gives a first indication on the performance of the DVB-SH waveform using the reference demodulator of clause A.11.2.1) instead of ideal channel estimation. The values for ideal and real channel estimation are summarized in table 7.9. The interleaver configuration used for this simulation is the short uniform interleaver.

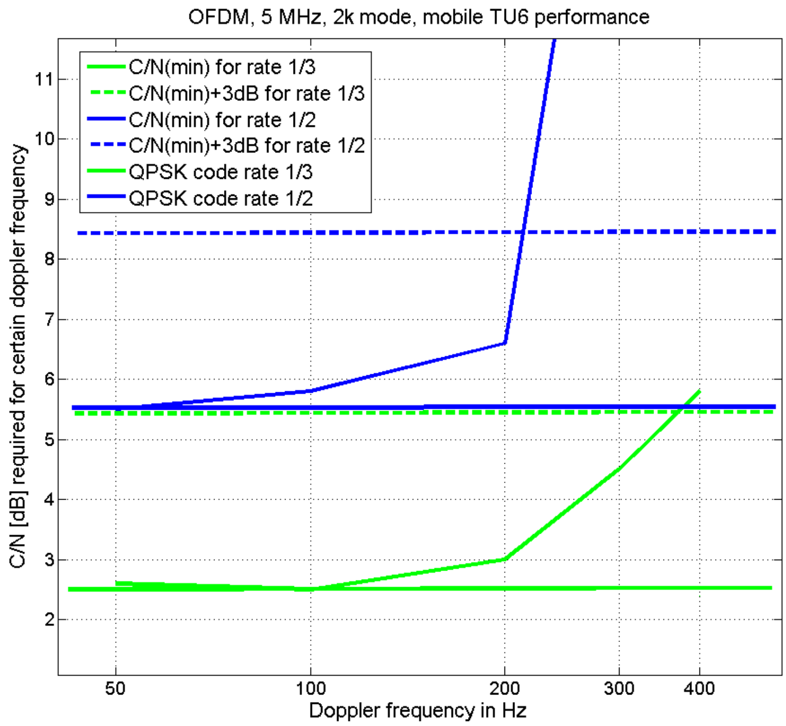


Figure 7.23: Simulated DVB-SH OFDM reference receiver C/N behaviour in mobile TU6 channel

From figure 7.23, the relevant parameters C/N_{ain} and Fd_{3dB} can be extracted for this case.

Table 7.9: Simulated DVB-SH OFDM reference receiver C/N behaviour in mobile TU6 channel

Code rate	Ideal estimation: C/N_{ain}	Real estimation: C/N_{ain}	Fd_{3dB}
1/2	4,0 dB	5,5 dB	230 Hz
1/3	1,0 dB	2,4 dB	370 Hz

Additional results on performance in mobile TU6 channels are for further study.

7.3.2 TDM Overview

The TDM modulation of DVB-SH is derived from DVB-S2 with reasonable adaptations necessary. An important new element is the constant distance of the pilot fields, independent from code rate choice, plus an addition of a signalling field incorporating all relevant parameters for FEC and interleaving.

The guidelines concerning modem algorithms design are given in the clause A.11.

TDM-specific simulation results are presented in clause A.11.

7.4 Combining techniques

Only a short overview of combining techniques is given below. The detailed guidelines on these techniques are for further study.

7.4.1 Overview to combining technology

In general the diversity combining technologies can be grouped in:

- antenna diversity;
- satellite-to-terrestrial hand-over and vice versa;
- diversity combining between different reception path (including hand-over to other frequency bands).

For the satellite-to-terrestrial combining and hand-over different scenarios has to be considered.

- SH-A, SFN
  In this case no special considerations has to be made for the combining. The satellite signal is considered as additional repeater in a SFN. But significantly higher delay spread may result and new characteristics of the channel impulse response have to be taken into account.
- SH-A, MFN
  In this case the same content may be available over different carriers. In this case it may be worthwhile that the receiver scans in the background for alternative sources for the program to support a seamless hand-over.
- SH-B
  In case of SH-B architecture the satellite and terrestrial signal is demodulated by separate demodulators. Three combining points are applicable in this case:
  - the signal is selected after the FEC decoding (selective combining). This method is not recommended;
  - the combining is done after de-interleaving and before FEC decoding. This allows that different interleaver profiles and even different code rates are used for the satellite and terrestrial signal. This method is called code combining. Code combining is very similar to maximum ratio combining. Further details on code combining and maximum ratio combining are given in clause 7.3.2.3;
  - combining before the de-interleaving (maximum ratio combining or selective combining):
    This method will only work if the terrestrial branch and the satellite branch use the same code rate and interleaver parameter. For a fully SH-B compliant receiver this method is therefore not recommended. This combining point is attractive for antenna diversity.

In case of SH-B other strategies may be applicable also. For example it may be possible switch off a receiver branch completely to reduce the power consumption. The evaluation of these techniques requires joint channel models covering the satellite and terrestrial propagation characteristics in one model. These models are currently under development.

7.4.2 Summary of combining techniques

Different methods of combining techniques are defined:

- **selective combining:** the signal, which provides better quality, is chosen;
- **maximum ratio combining:** the signals are combined, weighted accordingly to its specific reception quality;
- **complementary code combining:** different code bits are combined, chosen accordingly to its occurrence in the puncturing pattern.

7.4.3 Coexistence with other Physical layer elements

For further study.

7.5 Synchronization

This clause explains how synchronization is achieved in a DVB-SH network. Distinction is made on the different hybrid radio frequency configurations (SFN and non SFN).

7.5.1 Transmitter configuration in satellite-terrestrial SFN

7.5.1.1 Parameter selection

In the case of satellite-terrestrial SFN operation (SH-A and SFN), full synchronization between the satellite component and the terrestrial network is necessary. The terrestrial network includes all terrestrial repeaters belonging to one SFN cell. Throughout the whole DVB-SH network, different non-overlapping SFN cells may exist which have to be locally synchronized to the satellite but not between themselves.

Fine synchronization of terrestrial repeaters can be achieved, even when inclined GEO satellites are used as space component. This is achieved via:

- coarse time and frequency synchronization performed on the satellite signal inside the hub; with this coarse synchronization, the satellite signal is perfectly synchronized with the hub location;
- fine time and frequency synchronization performed at the terrestrial repeaters to compensate for variations within the satellite coverage, so between hub and SFN cells positions. This synchronization can be performed on a per-repeater granularity.

With these two steps of synchronizations, it is regarded feasible that the satellite signal can be synchronized with a remaining time error in the range between -5 % and +5 % of the guard interval length at each repeater cell centre. The difference of propagation time between the centre and the border of the repeater cell depends mainly on the configuration of the terrestrial transmitters, but also on the size of the terrestrial cell.

The information given below shows the order of magnitude for worst conditions (given a certain cell size). The principle behind the calculation of the delay called δd is given in figure 7.24. The following formula can be applied:

$$\delta d = \text{distance(satellite, repeater)} + \text{radius(repeater)} - \text{distance(satellite, receiver)}$$

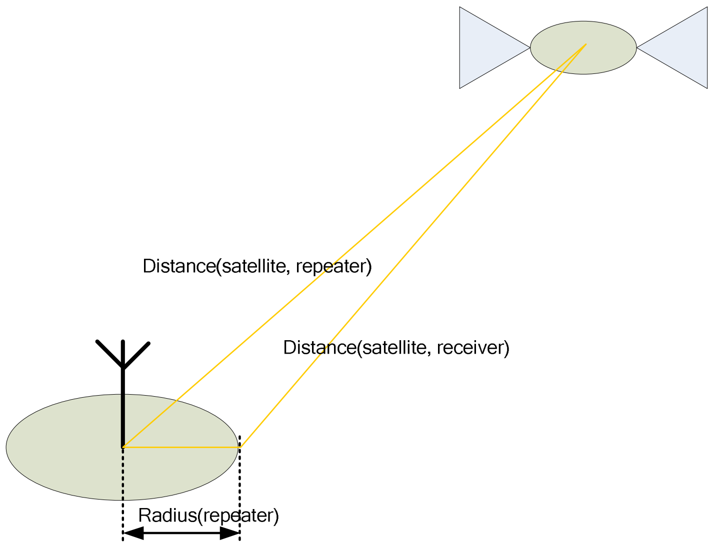


Figure 7.24: Delay calculation principle for SFN operation with 1 repeater

The distance is a function of longitudinal difference between repeater centre and satellite, latitude of the repeater centre and radius of the repeater. A parametric analysis shows the impact of the longitudinal difference (see figure 7.25).

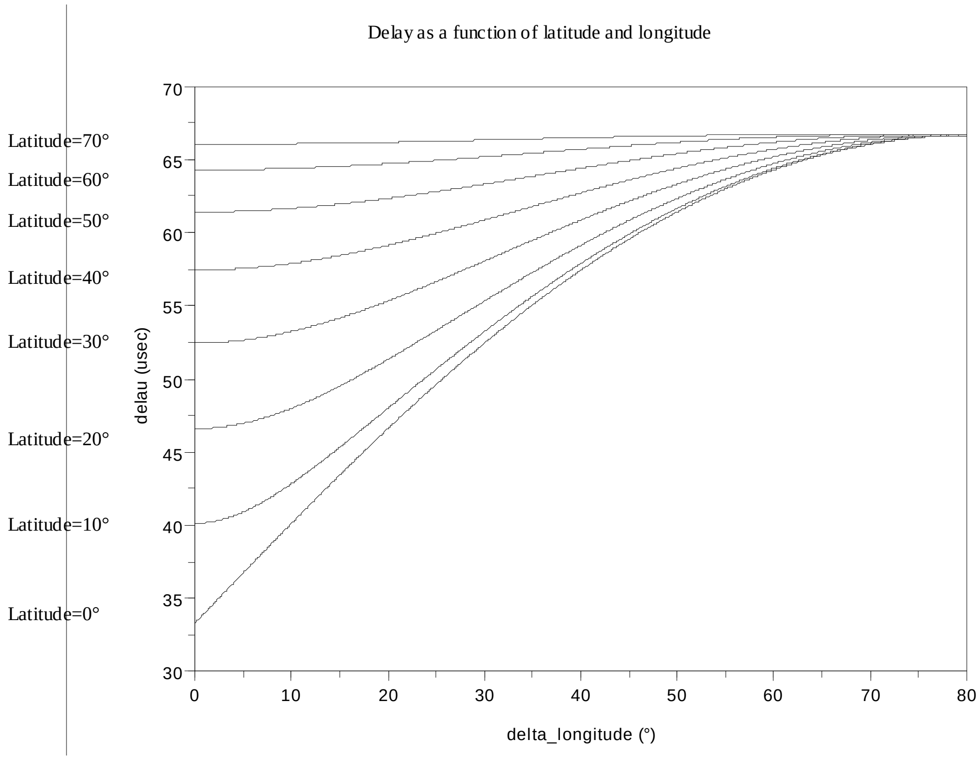


Figure 7.25: Delay as a function of latitude and longitude difference between satellite and terrestrial repeater cell, for a cell radius of 10 km

Taking the asymptotic values given with a delta_longitude at around 70°, the influence of the repeater radius can be displayed.

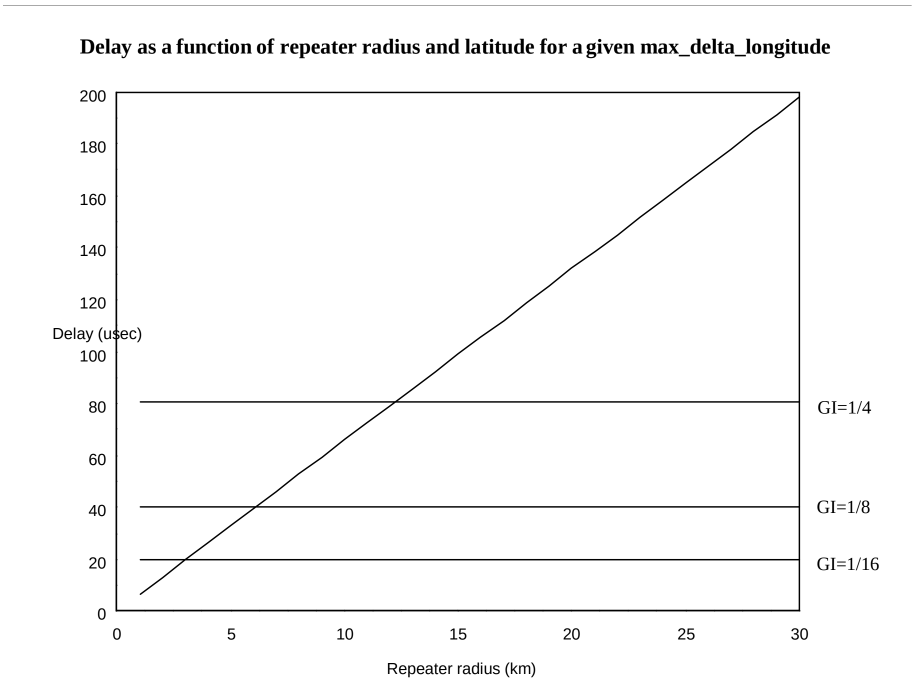


Figure 7.26: Delay as a function of the repeater radius for 70° delta_longitude

It must be recalled that these are minimum distances in the worst direction. In the opposite direction, the distance would be larger so that the "iso-δ zone" will not be a circle. The way this delay must be actually taken into account depends on network planning options:

- guard interval margins: reduction of the GI actual time due to lower constraints on repeaters synchronization; a typical value is 10 % requiring terrestrial transmitter precision of ±5 % of the guard interval value (e.g. for 2 k mode 5 MHz, a precision of 1,12 µs is required for the shortest guard interval 1/32) leading to a decrease of 10 % of the GI actual value. Room for delay spread on the terrestrial transmission component must also be considered: this value is function of the propagation environment;

- propagation losses: the effect of the delay reduces when the relative power of the components increases; so, propagations losses have to be considered.

Practical values for OFDM 2k mode, 5 MHz bandwidth are given in table 7.10.

Table 7.10: Max cell radius to ensure SFN between the satellite and terrestrial network at edge of one repeater

Max Cell radius	Max delay in us	GI = 1/4	GI = 1/8	GI = 1/16	GI = 1/32
12 km	79,8	80,64	40,32	20,16	10,08
6 km	39,9	80,64	40,32	20,16	10,08
3 km	19,65	80,64	40,32	20,16	10,08
1 km	6,55	80,64	40,32	20,16	10,08

For networks having several repeaters, the same principles apply, with additional contributions from the other receivers. See figure 7.27 for details.

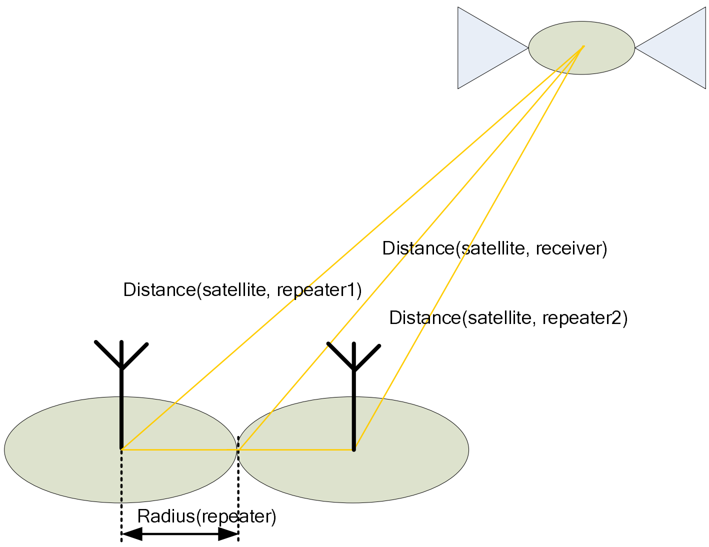


Figure 7.27: Delay calculation principle for SFN operation with 1 repeater

For the worst case assumption, the difference in distance is now:

$$\delta d = 2 * \text{radius(repeater)}$$

Provided the repeaters are synchronized with the satellite, they can be placed with distance equal to twice the maximum cell radius. Distances between repeaters could of course be made greater but then their reception would not be continuous, leading to a "spotted" map.

Depending on the type of network, the typical distances between repeaters would be in 2k FFT mode, 5 MHz:

- for 3G-based networks: 2 km allowing guard intervals of 1/32 or higher;

- for broadcast-centric networks: 12 km allowing guard intervals of 1/8 or higher.

7.5.1.2 Recommendation on network equipment

7.5.1.2.1 Principles of SFN architecture

In order to ensure exact synchronization between all transmitters, the transmitters are time synchronized by SHIP, providing functionality similar to MIP.

- In traditional DVB-H deployment, the MIP is inserted by dedicated equipment called MIP inserter located downstream of the IP encapsulator.
- In the context of DVB-SH, the SHIP insertion is highly recommended to be placed inside the IP encapsulator as the SHIP is also in charge of conveying SH services signalling that may be related to the DVB-H layers.

The SHIP enables SFN synchronization in a similar way as the MIP for DVB-T and most of the recommendations found in [DVB-T-IG] also apply as presented in figure 7.28. The SHIP includes two key parameters to enable synchronization:

synchronization_time_stamp: The *synchronization_time_stamp* of SHIP contains the time difference, expressed as a number of 100 ns steps, between the latest pulse of the "one pulse per second" reference (derived from GPS) that precedes the start of the SH frame M+1 and the actual start (i.e. beginning of first bit of first packet) of this SH frame M+1 at the output of the SHIP inserter. The granularity is 100 ns, the maximum value is 0x98 967F, equivalent to 1 s.

maximum_delay: The *maximum_delay* contains the time difference between the time of emission of the start of SH frame M+1 of the DVB SH signal from the transmitting antenna and the start of SH frame M+1 at the SFN adapter, as expressed by the value of its *synchronization_time_stamp* in the SHIP. The value of *maximum_delay* must be larger than the sum of the longest delay in the primary distribution network and the delays in modulators, power transmitters and antenna feeds. The granularity is 100 ns, the maximum value is 0x98 967F, equivalent to 1 s.

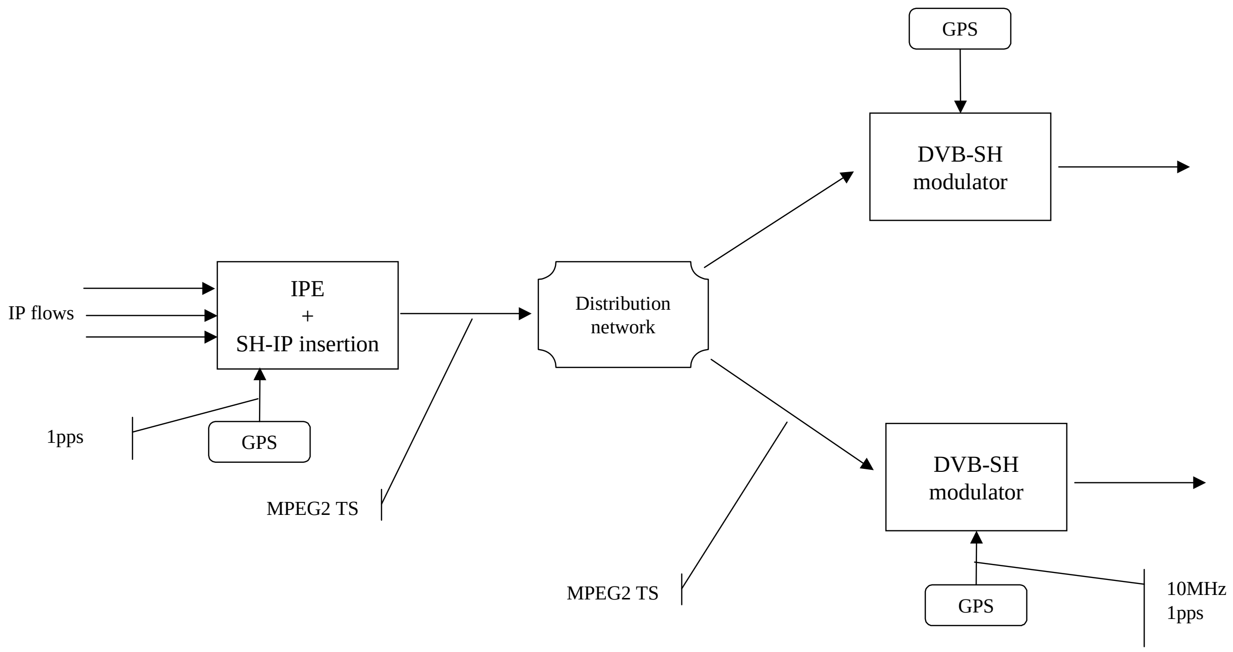


Figure 7.28: principle of SFN synchronization in a DVB-SH network

There exist particularities that are explained below.

7.5.1.2.2 DVB-SH particularities

Repetition period: the first particularity concerns the signalling period of the SHIP. There is no more any mega-frame since this has been replaced by the SH-frame. SH-frame has a repetition interval that does not depend on the mode (1k, 2k, 4k or 8k) but only on the selection of guard interval and modulation.

This repetition interval is always smaller than 1 s. In general, the repetition frequency of the SH frame (and the SHIP) is larger than for the MIP.

As more than 1 SH frame per OFDM super-frame may occur (QPSK 8k, 16 QAM 4k and 8k), it is recommended to check the frame-number (bit s23 and s24 of SHIP) to detect the first SH frame (number 0) of super frame, by the way, all exciters will transmit the same super-frame at the same time.

Interleaver control: the second particularity is the impact of the physical layer time interleaver. It spreads the data in time, with the full spread typically larger than the GPS 1pps impulse. This "latency" must not be confused with the delays introduced by the content delivery and the transmitter play-out, which are covered by the mechanisms above.

The following rules apply for SFN operation:

- modulators synchronize to the SFN network by evaluating the *synchronization_time_stamp* + *max_delay* and the GPS 1 pps and 10 MHz signalling, identical to DVB-H; this will take into account all transit delays brought by the digital and analogue processes;
- no compensation is made for the interleaver latency as the structure of the time interleaver is inherently self-synchronizing. The shortest tap L[0] (as seen by the receiver) of the time interleaver is (by nature) always set to 0, forcing the very first OFDM symbol of an SH frame to carry already some useful information;
- if the start of an SH frame is indicated in the *synchronization_time_stamp* (not equal to 0xFFFFFF), then the TPS bit s35 is immediately set in this super-frame.

7.5.1.2.3 Possible implementations

- To sign the emission date of the following SH-frame M + 1, the SHIP inserter will copy into the 3-byte valid *synchronization_time_stamp* field of the SHIP packet contained in the SH-frame of index M the value that will be reached by its own counter when it will emit the first bit of the first packet of the SH-frame M + 1.
- An offset value, entered by the human operator of the distribution network, is then inserted by the SHIP inserter in the 3-byte field *maximum_delay*. This "offset" field must be greater than the maximum delay spread introduced by the network and is expressed as a certain number of 100 ns long (1/10 MHz) periods.
- Due to its a priori knowledge of the PID of the SHIP, each channel modulator will extract the SHIP from each valid (synchronization_time_stamp ≠ 0xFFFFFF) SH-frame M of the incoming MPEG-TS and add both time field *synchronization_time_stamp* and *maximum_delay*. Then it waits for its local counter to reach the resulting value before inserting the associated following SH-frame M + 1 into the time interleaver.
- The instant for "emitting the associated following SH-frame" shall be understood as the instant of the first sample of the guard interval of the COFDM symbol that carries the first IU of the SH-frame. It is up to each modulator to take into account the transit delays brought by the digital and analogue processes, including processing time but not the delays for the time interleaver.

7.5.2 Transmitter configuration in non-SFN hybrid networks

For configurations without SFN between satellite and terrestrial component (e.g. SH-A non-SFN or SH-B), the receiver has the possibility to combine the signals from different reception paths.

The same rules for the use of the (terrestrial) SFN configuration apply as in SFN operation; however the satellite and terrestrial ground component have to meet less stringent timing and frequency accuracy requirements.

To allow combining between satellite and terrestrial signals, the alignment between the two components has to be ensured on the base of an SH frame. Therefore, the following requirements apply:

- at the terrestrial transmitter location, the time difference between the satellite and terrestrial component must be in the range between -5 % and +5 % of an SH frame duration. Typical values for SH frame durations are 100 ms or higher, thus the required time difference is between -5 ms and +5 ms;
- no specific requirements are made for the frequency synchronization between the two components;
- the requirements to compensate the latencies of two different interleavers on satellite and terrestrial transmission is a function of the time interleaver and therefore described in clause 7.2.3.4.2.

7.5.3 Receiver synchronization and re-synchronization

To successfully decode one FEC packet, various modules in the receiver have to be synchronized to the block transmission structure of the turbo code. This clause tries to summarize the different steps for a successful decoding and presents different possible strategies.

The use of time slicing introduces some constraints on the synchronization, as receivers will have to either completely re-synchronize at the beginning of each burst, or use prediction from internal oscillators.

This clause describes various strategies to synchronize both the OFDM and the TDM part with the SH frame. The first two methods do not reuse any existing knowledge from previous time-sliced bursts; the third one reuses information derived from previous time-sliced bursts.

7.5.3.1 SH frame synchronization strategy 1 (without a priori information)

If no prediction of the SH frame structure is provided by the receiver, the synchronization to the SH framing has to be recovered at each switch-on.

For OFDM, the TPS bits are used to recover the framing. By reading the TPS of the OFDM super frame, the receiver discovers the following information:

- s_{23} and s_{24}: OFDM frame number;
- s_{35}: OFDM super frame number in SH frame.

However, the bit s_{35} acquisition is only required in modes where the number of super frames per SH frame is larger than 1, i.e. in QPSK 1k, QPSK 2k, and 16QAM 1k modes. For all other modes, the bit s_{35} is not used.

After a maximum of 2 OFDM super frames (for the combinations as mentioned above) and 1 OFDM super frame for all other modes, the receivers is able to determine the SH framing and, thanks to the coding parameters already acquired, derive exact boundaries of the code words.

Actually, the maximum duration can be 2 (respectively 1) OFDM super frame(s) if the receiver uses this strategy.

For TDM, the SOF preamble is used to derive the boundaries of the SH frame. The maximum duration is identical to OFDM as the SH frame duration has been chosen identically for OFDM and TDM.

7.5.3.2 SH frame synchronization strategy 2 (without a priori information, for OFDM only)

The second strategy is only applicable to OFDM. It is based on the alignment of interleaver and capacity units with the OFDM symbols and founds on the pattern of the scattered pilots. Without knowing the SH framing, the de-interleaver can immediately start the de-interleaving process, as soon as the first group of 4 OFDM symbols has been acquired.

This is possible due to the self-synchronizing structure of the convolutional interleaver. Each 48 interleaver units (48*126 bit = 6 048 bit) can be handled as a group and fed into the de-interleaver. For any 4 OFDM symbols, the number of bits to be processed is an integer multiple of 6 048 bit:

- 1k, QPSK, 4 symbols equivalent to 756*4*2 bit = 6 048 bit;
- 8k, 16QAM, 4 symbols equivalent to 6048*4*4 bit = 96 768 bit = 16*6 048 bit.

After a sufficient de-interleaver duration (to receive enough parity data for decoding), it is possible to perform a limited number of "blind decoding trials".

7.5.3.2.1 Option 1

The number of tests is limited because the code words are aligned with the pattern of scattered pilots and, as a consequence, are aligned with the de-interleaver cycles. In practice, the number of cycles is equal to:

$$\frac{N_{BIL}}{6048}$$

(With N_{BIL} being the turbo block size after interleaver and puncturing), and is completely independent of the modulation.

The procedure is then the following:

- wait for 2 code words during $2 * \frac{N_{BIL}}{bit\_rate}$;
- test for successful decoding;
- if the test is not successful, shift by one cycle by waiting for 6 048 bits during a duration of $\frac{6048}{bit\_rate}$ and retry the decoding after a waiting time of $2 * \frac{N_{BIL}}{bit\_rate}$.

The maximum wait duration is $\frac{N_{BIL}}{6048} * \left(\frac{2 * N_{BIL} + 6048}{bit\_rate} \right)$, the average is half of this value.

7.5.3.2.2 Option 2

As soon as the received has done the scattered pilots acquisition, it is able to number the symbols in a row of 4 and to position the start of a code word with regard to the output of the smallest duration tap L[0]. The maximum number of tests depends on the modulation, FFT size and denominator of the fractional code rate. They are given in table 7.11 for code rate k/n where denominator n value is below or equal to 5.

The duration of the tests is implementation dependent and depends on the turbo-decoding strategy. Typical implementations will require reception of an equivalent of 2 code words for each try, implying reception duration of 2 code words multiplied by the number of tries.

The procedure is then the following:

- wait for 2 code words
- test for successful decoding
- if the test is not successful, move to next position without waiting and restart the test

In case of successful test decoding, the receiver knows the code word boundaries and also the CBCOUNTER_FB which gives the position index of the code word inside the current SH frame. So the receiver knows exactly the SH-framing.

These two strategies (frame duration and decoding tries) can be applied in parallel and the first successful one will be used to minimize the acquisition time. As can be seen in table 7.11, code word acquisition is generally faster than OFDM frame or OFDM super frame acquisition except for the 1k mode.

Table 7.11: Typical synchronization times for strategy 1 and strategy 2

MOD	carriers	1/code_rate	Maximum number_of_tries	Option 1 Nof_tries*(2 CW + 6 048) (ms)	Option 2 Nof_tries * CW (ms)	SH frame acquisition time (ms)
16QAM	8k	2	1	16	3,58	243,712
16QAM	4k	2	1		3,58	121,856
16QAM	2k	2	1		3,58	60,928
16QAM	1k	2	2		7,17	45,696
QPSK	8k	2	1	32	7,17	243,712
QPSK	4k	2	1		7,17	121,856
QPSK	2k	2	2		14,34	91,392
QPSK	1k	2	4		28,67	45,696
16QAM	8k	3	8	35	43,01	243,712
16QAM	4k	3	4		21,50	121,856
16QAM	2k	3	3		16,13	60,928
16QAM	1k	3	3		16,13	45,696
QPSK	8k	3	4	70	43,01	243,712
QPSK	4k	3	3		32,26	121,856
QPSK	2k	3	3		32,26	91,392
QPSK	1k	3	6		64,51	45,696
16QAM	8k	4	1	61	7,17	243,712
16QAM	4k	4	1		7,17	121,856
16QAM	2k	4	2		14,34	60,928
16QAM	1k	4	4		28,67	45,696
QPSK	8k	4	1	122	14,34	243,712
QPSK	4k	4	2		28,67	121,856
QPSK	2k	4	4		57,34	91,392
QPSK	1k	4	8		114,69	45,696
16QAM	8k	5	8	94	71,68	243,712
16QAM	4k	5	7		62,72	121,856
16QAM	2k	5	5		44,80	60,928
16QAM	1k	5	5		44,80	45,696
QPSK	8k	5	7	188	125,44	243,712
QPSK	4k	5	5		89,60	121,856
QPSK	2k	5	5		89,60	91,392
QPSK	1k	5	10		179,20	45,696

7.5.3.3 SH frame synchronization strategy 3 (with prediction)

This strategy is applicable to both OFDM and TDM. The receiver may provide frame prediction according to the following schemes. The prediction times are considered to be feasible with consumer product oscillators, assuming a frequency accuracy of (±) 50 ppm.

For OFDM, the use of 2k mode, 5 MHz bandwidth with QPSK modulation and guard interval 1/8 has been selected for this example:

- prediction of the OFDM symbol number is possible over **one** second within an interval of (±) 12,4 % (50 µs at a total symbol length of 403,2 µs per OFDM symbol);
- prediction of the OFDM regular pilot structure (4 OFDM symbols) is possible over **one** second within an interval of (±) 3,1 % (50 µs at a total length of 4 OFDM symbols of 1 613 µs);
- prediction of the OFDM frame structure (68 OFDM symbols) is possible over **ten** second within an interval of (±) 1,8 % (500 µs at a total length of 68 OFDM symbols of 27,4 ms);

The length of the SH frame in OFDM frames depends on the selection of the modulation and guard interval and the selection of the FFT mode. The following extreme cases are possible:

- "very short SH frame": For 8k FFT and 16QAM modulation, one SH frame maps to 1 OFDM frame;
- "very long SH frame": For 1k FFT and QPSK modulation, one SH frame maps to 16 OFDM frames;

Combined with the highest (9,14 MHz) symbol rate available, this maps to the following prediction accuracies:

- "very short SH frame": Prediction of the SH frame structure is possible over **ten** seconds within an interval of (±) 0,73 % (500 µs at a total length of 68 OFDM symbols of 68,54 ms);
- "very long SH frame": Prediction of the SH frame structure is possible over **ten** seconds within an interval of (±) 0,36 % (500 µs at a total length of 16*68 OFDM symbols of 137,09 ms).

It can be concluded that, with low requirements on the receiver oscillators used, the receiver is capable of predicting the SH frame alignment for both OFDM and TDM over a time span of more than 10 s and even with oscillators worse than those used for the calculation exercise. Using this proposed scheme, acquisition time is only dependent on the acquisition time of the demodulators, but no longer on the SH framing, allowing a good prediction of code word boundaries.

7.6 System throughput calculations

DVB-SH capacity at MPEG TS interface level may be calculated from the parameters defined in the waveform definition [1]. This is performed in calculating first the number of MPEG TS packets in an SH Frame, and then the duration of the SH frame. The ratio of these two numbers give the bit rate capacity of the system.

Please note that the throughput calculation is independent of the selection of FFT mode in OFDM. To support seamless hand-over between OFDM and TDM, the SH frame length of the TDM part has been aligned to the SH frame length in OFDM. Therefore, the TDM parameters can not be calculated independently but imply a selection of OFDM parameters which is represented in the following tables.

7.6.1 Calculation of the number of MPEG TS packets per SH Frame

Each turbo code word contains exactly 8 MPEG TS Packets.

For OFDM, table 5.11 of the waveform definition [1] gives the number of turbo code words in an SH frame as a function of the FEC code rate.

For TDM, table 5.12 of the waveform definition [1] gives the number of Capacity Units (CU) per SH Frame. A CU is a set of 2 016 bits. From table 5.8 of the waveform definition [1], the number of CU per turbo code word may be calculated for each code rate. The total number of turbo code words in an SH frame is then the integer part of the division of the number of Capacity Units (CU) per SH Frame by the number of CU per turbo code word.

7.6.2 Calculation of the SH Frame duration

The SH frame length is defined as a function of the length of OFDM frames. Therefore, the frame length of both an SH-frame in OFDM and TDM mode are identical. The number of OFDM Frames per SH frame is defined in the table 5.10 of the waveform definition [1]. Each OFDM Frame is composed of 68 OFDM Symbols. The duration of an OFDM symbol is given in tables 5.22, 5.24 and 5.26 of the Waveform Definition document [1].

7.6.3 Typical MPEG-TS bit rates for OFDM

The following tables give the MPEG TS bit rates in Mbps of the OFDM waveform for various DVB-SH configurations.

7.6.3.1 Channel Bandwidth 5 MHz, OFDM

Table 7.12

	GI	1/5	2/9	1/4	2/7	1/3	2/5	1/2	2/3
QPSK	1/4	1,333	1,481	1,679	1,876	2,222	2,666	3,357	4,443
	1/8	1,481	1,646	1,865	2,085	2,468	2,962	3,730	4,937
	1/16	1,568	1,742	1,975	2,207	2,614	3,136	3,950	5,227
	1/32	1,616	1,795	2,035	2,274	2,693	3,231	4,069	5,386
16QAM	1/4	2,666	2,962	3,357	3,752	4,443	5,332	6,714	8,887
	1/8	2,962	3,291	3,730	4,169	4,937	5,924	7,460	9,874
	1/16	3,136	3,485	3,950	4,414	5,227	6,273	7,899	10,455
	1/32	3,231	3,591	4,069	4,548	5,386	6,463	8,139	10,772

7.6.3.2 Channel Bandwidth 1,7 MHz, OFDM

Table 7.13

	GI	1/5	2/9	1/4	2/7	1/3	2/5	1/2	2/3
QPSK	1/4	0,427	0,474	0,537	0,600	0,711	0,853	1,074	1,422
	1/8	0,474	0,527	0,597	0,667	0,790	0,948	1,194	1,580
	1/16	0,502	0,558	0,632	0,706	0,836	1,004	1,264	1,673
	1/32	0,517	0,574	0,651	0,728	0,862	1,034	1,302	1,723
16QAM	1/4	0,853	0,948	1,074	1,201	1,422	1,706	2,149	2,844
	1/8	0,948	1,053	1,194	1,334	1,580	1,896	2,387	3,160
	1/16	1,004	1,115	1,264	1,413	1,673	2,007	2,528	3,346
	1/32	1,034	1,149	1,302	1,455	1,723	2,068	2,604	3,447

7.6.4 Typical MPEG-TS bit rates for TDM

The following tables give the TDM MPEG TS bit rates in Mbps of the TDM waveform for various DVB-SH configurations.

7.6.4.1 Channel Bandwidth 5 MHz, TDM, with 15 % roll-off and QPSK in OFDM

Table 7.14

		TDM	Modulation	QPSK	Roll Off :	15 %	OFDM modulation :		QPSK
OFDM GI		1/5	2/9	1/4	2/7	1/3	2/5	1/2	2/3
1/4		1,530	1,728	1,925	2,222	2,567	3,110	3,900	5,184
1/8		1,536	1,755	1,975	2,249	2,633	3,127	3,950	5,266
1/16		1,510	1,684	1,917	2,207	2,556	3,078	3,891	5,169
1/32		1,556	1,676	1,915	2,214	2,573	3,112	3,890	5,146

Table 7.15

		TDM	Modulation	8PSK	Roll Off :	15 %	OFDM modulation :		QPSK
OFDM GI		1/5	2/9	1/4	2/7	1/3	2/5	1/2	2/3
1/4		2,320	2,567	2,913	3,308	3,900	4,690	5,826	7,800
1/8		2,359	2,633	2,962	3,346	3,950	4,718	5,924	7,899
1/16		2,323	2,614	2,904	3,311	3,891	4,705	5,866	7,841
1/32		2,334	2,573	2,872	3,291	3,890	4,668	5,805	7,779

Table 7.16

		TDM	Modulation	16APSK	Roll Off :	15 %	OFDM modulation :		QPSK
OFDM GI		1/5	2/9	1/4	2/7	1/3	2/5	1/2	2/3
1/4		3,110	3,456	3,900	4,443	5,184	6,221	7,800	10,417
1/8		3,127	3,511	3,950	4,498	5,266	6,308	7,899	10,532
1/16		3,136	3,485	3,891	4,472	5,227	6,273	7,841	10,455
1/32		3,112	3,411	3,890	4,428	5,146	6,224	7,779	10,353

7.6.4.2 Channel Bandwidth 5 MHz, TDM, with 15 % roll-off and 16QAM in OFDM

Table 7.17

		TDM	Modulation	QPSK	Roll Off :	15 %	OFDM modulation :		16QAM
		1/5	2/9	1/4	2/7	1/3	2/5	1/2	2/3
1/4		1,481	1,679	1,876	2,172	2,567	3,061	3,851	5,134
1/8		1,536	1,646	1,865	2,194	2,523	3,072	3,840	5,156
1/16		1,510	1,626	1,859	2,207	2,556	3,020	3,833	5,111
1/32		1,436	1,676	1,915	2,154	2,513	2,992	3,830	5,146

7.6.4.3 Channel Bandwidth 5 MHz, TDM, with 25 % roll-off and QPSK in OFDM

Table 7.18

		TDM	Modulation	QPSK	Roll Off :	25 %	OFDM modulation :		QPSK
OFDM GI		1/5	2/9	1/4	2/7	1/3	2/5	1/2	2/3
1/4		1,432	1,629	1,827	2,074	2,419	2,913	3,653	4,888
1/8		1,426	1,536	1,755	2,030	2,359	2,852	3,566	4,718
1/16		1,452	1,568	1,801	2,033	2,381	2,904	3,601	4,821
1/32		1,436	1,556	1,795	2,035	2,394	2,872	3,591	4,787

7.6.4.4 Channel Bandwidth 1,7 MHz, TDM, with 15 % roll-off and QPSK in OFDM

Table 7.19

		TDM	Modulation	QPSK	Roll Off :	15 %	OFDM modulation :		QPSK
OFDM GI		1/5	2/9	1/4	2/7	1/3	2/5	1/2	2/3
1/4		0,490	0,553	0,616	0,711	0,822	0,995	1,248	1,659
1/8		0,492	0,562	0,632	0,720	0,843	1,001	1,264	1,685
1/16		0,483	0,539	0,613	0,706	0,818	0,985	1,245	1,654
1/32		0,498	0,536	0,613	0,709	0,823	0,996	1,245	1,647

Table 7.20

		TDM	Modulation	8PSK	Roll Off :	15 %	OFDM modulation :		QPSK
OFDM GI		1/5	2/9	1/4	2/7	1/3	2/5	1/2	2/3
1/4		0,743	0,822	0,932	1,058	1,248	1,501	1,864	2,496
1/8		0,755	0,843	0,948	1,071	1,264	1,510	1,896	2,528
1/16		0,743	0,836	0,929	1,059	1,245	1,505	1,877	2,509
1/32		0,747	0,823	0,919	1,053	1,245	1,494	1,857	2,489

8 Services

8.1 General considerations

From the service point of view, the following similarities between DVB-SH and its DVB precursor can be noted:

- DVB-SH signal is intended to be received by a variety of mobile and portable devices;
- DVB-SH offers high data rates, even in moving conditions and reuses technical elements of the DVB-H specification so that harsh conditions of terrestrial, urban mobile propagations can be addressed in the same manner;
- DVB-SH is a broadcast-centric delivery system: the same content is delivered to an unlimited audience without the risk of network saturation;
- DVB-SH has the flexibility of narrow-casting thanks to the support of multicast protocols;
- DVB-SH is an IP-based system using MPEG2-TS as the baseline transport layer (and consequently, DVB MPE as the default encapsulation protocol). Existing DVB signalling such as PSI/SI, SFN, Time-Slicing, MPE-FEC are mostly preserved;
- DVB-SH can carry DVB-IPDC in essentially the same way as DVB-H does. In this framework, it can support two-way interactive services when coupled with an appropriate interactive channel (albeit, with the negative impacts of long delays incurred when in the satellite-only reception).

There are however key differences from DVB-H that should be kept in mind, due to the hybrid-network nature of DVB-SH:

- DVB-SH operates in frequency bands internationally allocated to Satellite. The time line of its deployment is not directly affected by the "Digital Switch-Over" of analogue TV;
- it was not envisaged that current set-top-boxes should be able to receive a DVB-SH signal. Therefore, DVB-SH does not require that the same MPEG2 multiplex be shared with DVB-T or DVB-S/S2 services, nor does DVB-SH requires the use of DVB scrambling at MPEG2-TS for Access Control;
- although the first deployments of DVB-SH will be based on MPE, the DVB-SH framing specifications has the provisions to allow migration to the new GSE protocol;
- when the 2 GHz S-band band is used, synergy with 3G telephony infrastructure would be exploited, especially in areas where such infrastructure already exists. Network planning for DVB-SH in urban areas could be similar to the 3G planning with the benefit that indoors coverage could be made essentially the same;
- a DVB-SH coverage is always composed of a satellite coverage complemented by a terrestrial coverage. The services offered in these two coverages are strongly linked but not necessarily the same;
- as a consequence of the above, DVB-SH services are a mix of Common services and Local services. Common services are services that are available in the SC and **must** be transmitted in the CGC (although possibly with different attributes, see clause 8.2.2 for more details). Local services are services that are available in the CGC only. Common services are usually those with very large audience while Local services have more fragmented audiences, possibly with geographical dependencies. A Local service package for one city/town may differ from the package for another city/town;
- there are challenges for DVB-SH due to higher mobility, satellite specific propagation channels, and, in some cases, higher frequency bands. Some of these constraints are addressed somewhat differently by SH-A and SH-B architectures;
- although the Common services are available in both the SC and the CGC, the service attributes may differ depending on the user location, more precisely between different reception modes: satellite-only, terrestrial-only and combined satellite-terrestrial receptions. For example, specific physical parameters may be selected so that higher user speed is possible with satellite-only reception. On the other hand, less user cooperation is required with terrestrial-only reception where LOS is not required;

- DVB-SH interactive services would rely mainly on a terrestrial return channel which could be independent from its CGC. It should be noted that the technical possibility exists, in the 2 GHz S-band, to establish a direct return path via satellite. Such possibility could be invaluable in catastrophic events leading to unavailability of terrestrial infrastructures.

8.2 Service classification

DVB-SH services could be considered as just a supplement to traditional TV broadcasting. In such a narrow view, DVB-SH just extends the service consumption from the homes to other places. Taking a more forward-looking view, it is expected that new services will naturally develop, exploiting close cooperation between the broadcast infrastructure, the mobile infrastructure and the content provisioning infrastructure. The characteristics of the whole spectrum of services that can be offered with DVB-SH with the above assumption and their implications on the network, the terminal and the user behaviour are discussed below.

8.2.1 Service categories

The services can be categorized using several criteria related to the way it is intended to be consumed. The following questions may be helpful in defining a particular focus for a particular DVB-SH implementation (see note).

1) is the service Common or Local?;

2) on which terminal category is the service most likely to be received? (Refer to clause 10 for terminal categories definitions and descriptions);

3) what are the likely use-cases? An important factor to take into account is the degree of "cooperation" that can be expected from the user. User cooperation means that the user can adapt his speed, can position/configure the terminal for maximum reception quality or accepts to refrain from certain actions when he is notified that the actual receive condition is not compatible with such actions. An example is a system that can warn the user that he is receiving under frequent signal blockages so that although the FEC is recovering the packet losses of the current service, the zapping time will be excessive due to the necessity to refill the FEC receive buffer for the new service;

4) another important fact to take into account is the expected speed of the terminals under use. A vehicle-mounted application will obviously be travelling at high speeds. However, hand-held devices such as mobile phones or portable media players will commonly be used by passengers in vehicles. Also, car manufacturers are increasingly supporting streaming audio from mobile handsets by Bluetooth into the car stereo system; again this must work "at speed" and the extension of this to video should also be considered;

5) does the service always require a display and what is the maximum size of such display? A radio service does not require the display to be switched on (or the display can be used for another application, e.g. games). If the display is switched off, the user may wish to put the terminal in his pocket (in the same way he is used to with his MP3 player). The video bit rate should be sufficient in order to avoid visible video artefacts;

6) does the service require "multi-tasking"? For example, is it required to be able to watch a program while answering an incoming phone call with the same device? Another example is the ability to download a program while watching another one;

7) is the service time-critical? Can the content be delayed by several seconds compared to the actual time it is occurring or shown in traditional broadcasting? An example of time-critical service is a live TV program with participation of the selected audience by phone (mobile or fixed phone calls). Such services would not tolerate a latency of more than 600 ms;

8) are feedbacks of terminal information available? Terminal information could be: terminal Class, terminal location (at least coverage location, e.g. SC coverage or CGC cell-id), max speed, on-time periods, receive quality parameters, etc. These feedbacks could be used by the service provider or the network provider. These feedbacks could be in real-time during interactive sessions of an interactive service but they can also be periodically scheduled.

NOTE: The guidance for the selection of key DVB-SH technologies and architectures according to the main service requirements is for further study.

8.3 Service attributes

8.3.1 Definition

Service attributes describe the properties attached to a service as perceived by the end user in various reception conditions. This covers classical parameters found in DVB-H but also new parameters required by the DVB-SH specificities.

DVB-H classical attributes:

- bit rate profile (CBR, VBR, etc.);
- quality of service (Bit Error Rate, Frame Error Rate, ESR(5), etc.).

DVB-SH specific attributes:

- coverage (satellite, terrestrial);
- zapping time;
- end-to-end delay;
- ease of use in various mobility situation (moving speed, static reception, various degrees of user cooperation, etc.).

8.3.2 Application

The service attributes can be defined at several interfaces.

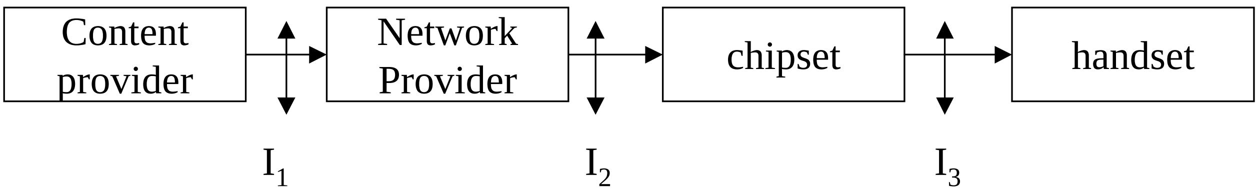


Figure 8.1: Application of service attributes concept

- **interface between content and network provider (I_1):** a service level agreement (SLA) is contracted between the content provider and the network operator so that the content operator requests specific quality, the coverage, the type of target terminals;
- **interface between network operator and chipset (interface I_2):** network operator transmits the "semantic" attached to the content using over-the-air ESG metadata in a non real time manner. The ESG from DVB-H needs to be updated with the added semantics required in DVB-SH;
- **interface the chipset and the handset (interface I_3):** only the chipset is able to know the real-time parameters such as if and when the user passes from one area covered by the satellite only to another covered by the CGC. This real time information is very "terminal-dependent" and is not be subject of the standardization process but left for implementation (the chipset may get this information by different means). By crossing the real time information with the semantics conveyed by the ESG, the handset is able to process the content and display it to the end user in an agreed manner. Different policies can be found:
    - the ESG display may differentiate those content that can be found on the satellite coverage from those that are available on terrestrial only coverages so that the user is aware that some may not be found everywhere. So the user is aware that some content may not be available everywhere. He could adopt his consumption behaviour based on this information (for a commuter between a rural and urban area, he would know where the rural content is aware, then he selects it);

- for those contents that are not available everywhere, some hysteresis behaviour is possible at the boundary. It may be useful for the terminal to show the content inside the ESG as "selectable" only when its reception is stable enough;
- zapping time may differ between those contents that are available via satellite (and which could necessitate a longer zapping time in specific degraded reception conditions) and those that are available via terrestrial and should have a "terrestrial" zapping time. For the satellite contents, decision could be made depending on actual reception conditions to display immediately (typically under CGC) or to delay display (in case conditions may be less favourable, for instance under satellite-only coverage). But another policy could be to display immediately, whatever the actual quality of the stream is, in order to favour immediate perception of the user. So zapping time could be made variable in terms of content attributes and terminal conditions.

Table 8.1: Service attributes at interfaces

Service attributes	I_1	I_2	I_3
Bit rate profile	X		
Quality of service	X		
coverage (satellite, terrestrial)	X	X	X
zapping time requirements	X	X	X
end-to-end delay requirements	X	X	X
Target terminal (moving, fixed, cooperative…)	X	X	X

8.4 Consequences of hybrid architecture on implementation of a DVB-SH service offering

The hybrid nature of DVB-SH systems has consequences on the several service aspects: handover implementation, service discovery and access, Electonic Service Guide (ESG). It should also be noted that the regulations in some countries could allow the deployment of the CGC in anticipation of the satellite launch.

8.4.1 Handover issue

For the Local services which are available only terrestrially, handover can be treated in the same way as in DVB-H. It should be noted that in the 2 GHz S-band, the CGC transmitters are likely to implement an SFN over large areas, at least at the level of a city coverage.

In a SH-A configuration, it is most likely that SFN between SC and CGC is implemented. Handover is not necessary in this case for the Common services.

In a SH-B configuration, handover is required since the Common services must be transmitted on two different frequencies for the SC and the CGC, using (by definition) two different modulations and possibly different FEC and interleaver parameters (the same is true for SH-A *whenever SFN is not used*).

It should be noted that there are two differences between normal handover and the present case:

- firstly, normal handover assumes that there is only one front-end in the receiver. In SH-B, the TDM/OFDM configuration implies two demodulators working in parallel, one for the satellite signal and the other for the terrestrial signal (the RF front-end may be shared);
- secondly, the satellite TS and the terrestrial TS are not necessarily signalled by DVB-SI as two different TS. When the soft combining property is configured, the two signals coming from the satellite and terrestrial origins are merged before the Turbo decoder and appears at this decoder output as a single TS (see clause 7.2.2.3.3 Maximal ratio combining and complementary code), despite the fact that the actual content of this TS will vary as a function of the presence or absence of the CGC signal. When soft combining is not used, handover is merely switching between two TS outputs. It is recommended that even in this case the PSI/SI be kept identical between the two TS.

8.4.2 Network and service discovery mechanisms

Network discovery in DVB-SH is based on DVB PSI/SI with the introduction of new descriptors (to be detailed in a new revision of EN 301 192 [9]).

Due to the possible use of SFN between the SC and its CGC, all the transport streams, whether transmitted via satellite or via terrestrial repeaters, belong the same network in the DVB sense.

Independently of the use of SFN, a DVB-SH network may be divided into "regions", within each a different terrestrial frequency plan is used. A simple example of the partitioning of an SH-network into regions is given in figure 8.2.

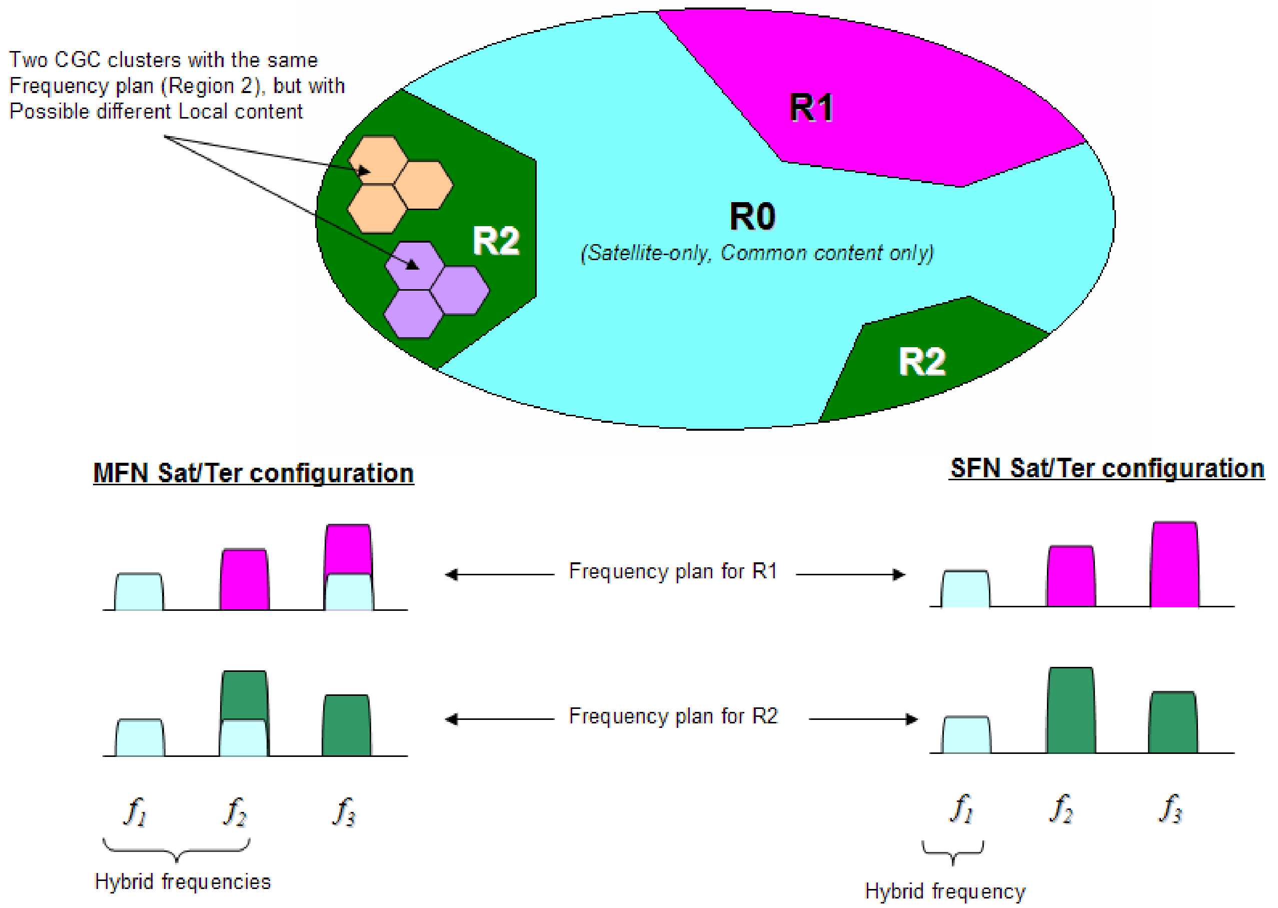


Figure 8.2: Illustration of the concept of partitioning an SH-network into regions

As illustrated by the figure above, a Region should **not** be interpreted as a contiguous area, or as a cluster of CGC coverages in SFN mode. For example, in Region R2, the figure shows two disjoint clusters of CGC coverage in SFN mode. The Local content in these two clusters may be different while their frequency plan is identical, by definition of the term "Region". When the SC and the CGC operate in MFN for the Common content, the frequency of the terrestrial retransmission (the "hybrid frequency") may be different between Regions, as depicted in the left part of the examples. When the SC and CGC operate in SFN for the Common content, the physical parameters chosen for the two "terrestrial-only" frequencies may differ between Regions, as shown in the right part of the examples. This may be imposed by interference constraints at the border of the satellite beam (see discussion on "exclusion zones" in clause 11). This is illustrated in the figure by indicating that frequency f3 has more terrestrial capacity than frequency f2 in Region R1 (and vice versa for Region R2).

The information on the validity of a frequency plan when the receiver is in a given region is signalled through a mechanism similar to the one used to signal the Cell_ID in DVB-T/H.

Usually, the Common services of a particular network are managed by a single IP Platform, possibly separated from the Local services which can be managed by several different IP Platforms. The user may have to select several IP Platforms.

It is desirable that, in normal operation, the need to monitor and parse SI information is limited to a minimum and that the discovery and selection of services are based on metadata signalled at the IP level.

8.5 Other considerations

DVB-SH terminal is expected to enjoy a regime of free circulation and use all over Europe as it is the case with mobile phones and receivers of broadcast services.

Terminals with transmit capability on satellite S-band (1 980 MHz to 2 010 MHz) may require compliance to the following ETSI standard (see note).

It is possible that, through roaming agreements, a Common service in one satellite coverage could be made available also as a Local service in another satellite coverage. It is highly desirable that such feature be offered in the way that looks transparent to the user. In cross-border situations it is desirable that such feature be treated as a special handover.

EXAMPLE: EN 301 442 [i.32]: "Satellite Earth Stations and Systems (SES); Harmonized EN for Mobile Earth Stations (MESs), including handheld earth stations, for Satellite Personal Communications Networks (S-PCN) in the 2,0 GHz bands under the Mobile Satellite Service (MSS) covering essential requirements under Article 3.2 of the R&TTE directive".

9 Network configurations

9.1 Considerations on network configurations

Two main type of networks can be considered. On one side we have network that are designed to target portable devices (handsets, handheld devices) and on the other side, network that are designed to target vehicular reception.

9.1.1 Mobile TV network targeting portable devices

The Mobile TV service to handsets are/will be mostly deployed, managed and commercialized with the support of mobile operators. The typical handset will be a cellular handsets with a 2,2" to 4" screen that will allow to display mobile TV on top of regular features such as voice telephony, SMS or Internet access. Depending on countries, these handsets will be subsidized by the mobile operator that will get revenues from the mobile TV (flat fee). Interactive services, possibly VoD and other service will transit using the existing 3G network and will provide additional revenues to the mobile operators.

Another Mobile TV offer can also target non connected devices like Portable Multimedia Players. In this case the service, likely to be controlled by broadcasters, is an extension of broadcast digital TV paradigm to portable receivers essentially based on live channels reception combined with PVR functionalities. Both free to air or pay TV model would apply.

In both cases, services can be provided directly from a satellite in rural areas but a terrestrial repeater network should be deployed to provide indoor coverage in urban areas where the satellite signal is insufficient. The terrestrial repeaters being designed for a smooth integration in existing 2G or 3G cellular sites. Depending on the requested indoor coverage quality, the number of TV channels, the number of cellular sites to be equipped with a low power DVB-SH repeaters will range from 1/4 up to 1/1. The terrestrial repeaters can be fed either via satellite signal using a DVB-S2 radio interface (typically in Ku or Ka band) or via terrestrial IP network.

In rural areas, satellite ensures nationwide direct reception. To improve indoor coverage, domestic gap-fillers can be used in rural areas to improve satellite reception. Repeaters in urban areas, mostly for indoor coverage, support satellite coverage. These repeaters re-transmit the nationwide channels and will therefore offer indoor coverage identical to a co-located UMTS system. To increase the system capacity in urban areas, adjacent carrier transmitters complement the satellite signal, allowing additional Local content. In each urban area, these complementary channels can be either national, regional or local ones. Finally, for handset based service, a cellular network is used to stream in unicast TV channel with limited audience TV and Video on Demand (VoD) as well as for interactivity return channel.

9.1.2 Mobile TV network targeting vehicle

The Mobile TV service to vehicle are mostly deployed, managed and commercialized directly by the broadcasters or the satellite operators, possibly without the support of mobile operators. The typical terminal will be installed in a car with typically a 7" screen, hence higher data rate per channel is expected. This screen can be coupled with other services such as GPS, Video player, MP3 player, etc. Possibly, limited interactive services will be possible using a return channel directly to the satellite.

The objective of the infrastructure is to optimize outdoor coverage across a territory. Services will be provided directly from a satellite in most areas but a terrestrial repeater network should be deployed to provide coverage in dense urban areas where the satellite signal could be insufficient (street canyon, underpass, etc.). As there is no need in this configuration for indoor coverage (except in tunnels), the repeater network will mainly use medium power DVB-SH repeaters that in most case will not be collocated with existing cellular infrastructure. Typically, only around 10 repeaters will be required to cover a typical urban metropolitan area. The terrestrial repeaters can be fed either via satellite signal using a DVB-S2 radio interface (typically in Ku or Ka band) or via terrestrial IP network.

Satellite ensures nationwide direct reception. Suitable spatial technologies (large antennas, high-power platforms, etc.) are used to provide the required net capacity. Repeaters in dense urban areas support satellite coverage. These repeaters re-transmit at the frequency of the satellite carrier. According to frequency scheme selection, a similar complement of urban channels can be implemented. Finally, a return link from the vehicular terminal up to the satellite can be used to provide interactive service using for example the ETSI standardized GMR1and2 waveform already operational on Thuraya and AceS system.

9.2 Synchronization of Satellite and CGC for Common content

9.2.1 Introduction

Basic mode of operation of DVB-SH networks rely on hybrid architecture when satellite broadcast content is retransmitted terrestrially in areas when satellite link availability is reduced. When those terrestrial repeaters are deployed, SFN operation is of great importance and is required in the following conditions:

- when operating in SH-A mode, there is a need for dual SFN operations : between the different terrestrial repeaters, like in DVB-T or DVB-H networks, and between the satellite and the terrestrial repeaters, in a way called hybrid SFN operation;
- when operating in SH-B mode, there is a need of synchronicity between the different terrestrial repeaters and the satellite but the requirements are not as stringent as for SH-A mode.

9.2.2 Terrestrial SFN

The terrestrial SFN operation is described in TR 101 190 [i.22] "Implementation guidelines for DVB terrestrial services". Only basic principles are recalled here, and some differences between the two systems are described in more details.

9.2.2.1 Principle

In a SFN, all transmitters are synchronously modulated with the same signal and radiate on the same frequency. Due to the multi-path capability of the multi-carrier transmission system (COFDM) signals from several transmitters arriving at a receiving antenna may contribute constructively to the total wanted signal.

However, the limiting effect of the SFN technique is the so-called self-interference of the network. If signals from far distant transmitters are delayed more than allowed by the guard interval they behave as noise-like interfering signals rather than as wanted signals. The strength of such signals depends on the propagation conditions, which will vary with time. The self-interference of an SFN for a given transmitter spacing is reduced by selecting a large guard interval. It should be noted that the impact of delayed signals outside the guard interval may depend on receiver design. As an empirical rule, to successfully reduce self-interference to an acceptable value the guard interval time should allow a radio signal to propagate over the distance between two transmitters of the network. In order to keep the redundancy due to the guard interval down to a reasonably low value (25 %), the useful symbol length has also to be large given the transmitter spacing. On the other hand a smaller guard interval would lead to a higher number of transmitters. The SFN operation is spectral and power efficient as it uses the same frequency channel overall, and increases the coverage or the signal over noise ratio. The price to pay is an accurate time and frequency synchronization all over the coverage area.

9.2.2.2 SFN operation

The same content is received from distribution network and transmitted at the same time by all the transmitters of the SFN, with the same waveform strictly and exactly. Due to the nature of the signals, SFN operation have some constraints recalled below:

- *frequency synchronization*: if f_k denotes the ideal RF position of the k^{th} carrier, then each transmitter should broadcast this k^{th} carrier at $f_k \pm (\Delta f / 1\,000)$;
- *time synchronization*: the Guard Time has been introduced to cope with the different echoes in terrestrial channels, and also for SFN purpose. Thus, to avoid Guard Time wasting, a strong synchronization constraints must be put on the transmitters : a few µs difference should be the maximum (less than 10 % of the Guard Time).

The different constraints impose an absolute time reference like the GPS for all transmitters, and the transmission of start signal ("top") to allow all transmission at the same time. This is performed through the Mega Frame in DVB-T/H and through the SH Frame in DVB-SH.

9.2.2.3 SHIP solution

The general architecture of the SFN synchronization scheme is provided here in figure 9.1.

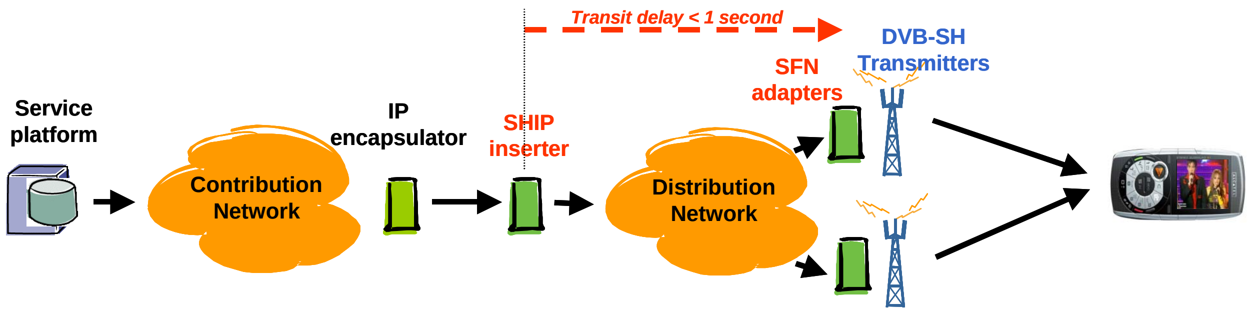


Figure 9.1: SFN synchronization architecture

Distribution network characteristics:

- heterogeneous network: satellite, terrestrial;
- differential transit delay performance from one path to another.

Synchronization principles:

- SH Frame Information Packet inserter: insertion of a GPS based timestamp (±0,1 us accuracy) in the SH-FRAME indicating the transmission time of the beginning of the next SH Frame;
- SFN adapters in the transmitters (repeaters): buffering of incoming MPEG-TS packets and transmission of SH Frame aligned with GPS relative time stamp.

The definition and specification of the SHIP is provided in EN 302 583 [1].

The complete mechanism is described in clause 7.6.1.2.

9.2.3 Synchronization of Satellite and Terrestrial repeaters for the Common Content

9.2.3.1 SH-B and SH-A/MFN cases

In this configuration, content is broadcast on two separate carriers, each having dedicated modulation schemes. Two demodulators are required in the receiver; at this stage, no specific time or frequency synchronization is required. Each demodulator outputs a time de-interleaved SH Frame, one for the satellite path and one for the terrestrial path. Typical combination scheme consists in combining content prior to FEC decoding (see clause 7.3.2.3). It benefits from the FEC code combining performance and provides straightforward seamless reception. This requires time synchronization of both SH Frames, and controlled difference between symbol rates.

DVB-SH framing provides TDM and OFDM SH Frames of identical period; this eases the synchronization process in the receiver, as it provides simple repetitive and predictable processing. At network level, only relative delays have to be handled. Typically, the satellite component has the longest delay: it is the combination of the transmission path (about 250 ms), plus the length of the Physical Layer interleaver (potentially up to several seconds). Thus, at terrestrial repeater level, these satellite delays must be compensated to minimize the level of buffering memory with the receiver. Residual variation delay is less than a few milliseconds, which much less that the size of the memory required for physical layer time de-interleavers. Synchronization between transmit signals (satellite and terrestrial repeaters) is performed using the SHIP information, in relation with configurable geographical-dependant information in each repeater site. In the receiver, time synchronization between both components is ensured by the SH-Frame (Signalling Field for the TDM component, and OFDM frame for the OFDM component).

9.2.3.2 SH-A case

9.2.3.2.1 Principle

The global satellite coverage and the geographically distributed overlap of hybrid satellite/terrestrial zones require specific dispositions to ensure deriving maximum benefit of SFN capabilities to mobile broadcast services networks; this involves time phase and frequency shift control over the several possible paths used to transport broadcast signals to the User Equipment.

a) satellite-originating broadcast signals are everywhere pregnant and hence constitute a de facto reference to which the terrestrial segment must be adjusted to ensure the signals transmitted over the different paths arriving at the receiver should be coherent in time, and frequency;

b) DVB-SH is the reference ETSI standard applicable to hybrid satellite/terrestrial coverage architectures; its waveform definition [1] provides through the SH frame a "container" structure for SFN synchronization data (SHIP), adapted from the DVBT/H standards;

c) Each hybrid satellite/terrestrial network requires the elaboration of customized SFN synchronization data, that are dependent on the actual specifics of the satellite orbital behaviour, which must distinguish the following cases:

   - geostationary orbit (GEO);

   - near-Geostationary orbit;

d) Relative Time phase and frequency shift variations between the satellite direct path and the indirect repeated last mile path affect the architecture performances and engineering optimization (cell size, etc.) of the ground network, and must be tuned carefully.

Hybrid SFN engineering considerations lead to ensure that:

- time phase performances synchronization between direct and indirect path of some fraction (typically a few percents) of the Guard Interval duration

- frequency synchronization performances between direct and indirect paths of some fraction (typically a few thousandths) of subcarrier spacing (the which e.g. is 1 709 Hz for a 7 MHz band and a 4 K FFT size)

This requires that all equipments be synchronized onto a unique and stable clock reference and/or a unique and stable frequency reference. This is ensured by the synchronization of all terrestrial equipment (IPE/ Satellite gateway modulation stages/ Repeater including frequency Up and down converters) onto the GPS 1 pps clock and/or 10 MHz frequency references.

9.2.3.2.2 Application with near GEO

With active house keeping, GEO satellite orbit should be kept within less than ±0,1 ° in both directions. The near GEO satellites may be inclined by as much as some few percents (typically up to (±) 5/6°. Though, correction procedures can be simplified for GEO satellite, we propose the analysis for the worst case, that is to say near GEO orbits.

9.2.3.2.2.1 Time delay

The direct path through this satellite experiences changes in the absolute time delay involving four location parameters:

a) Satellite gateway location

b) Cell location in the hybrid satellite/terrestrial coverage zone

c) Ground repeater location within the terrestrial cell

d) UE location in the cell

Maximum cell range resulting from CGC engineering results from ground network planning architecture and dimensioning choices. The following schematic displays the several path components contributing to relative delays between the direct satellite broadcast path and the indirect repeated last mile path. The distribution network for repeater content can be of different nature as for DVB-T/H networks. In the shown case, distribution network is using satellite S2.

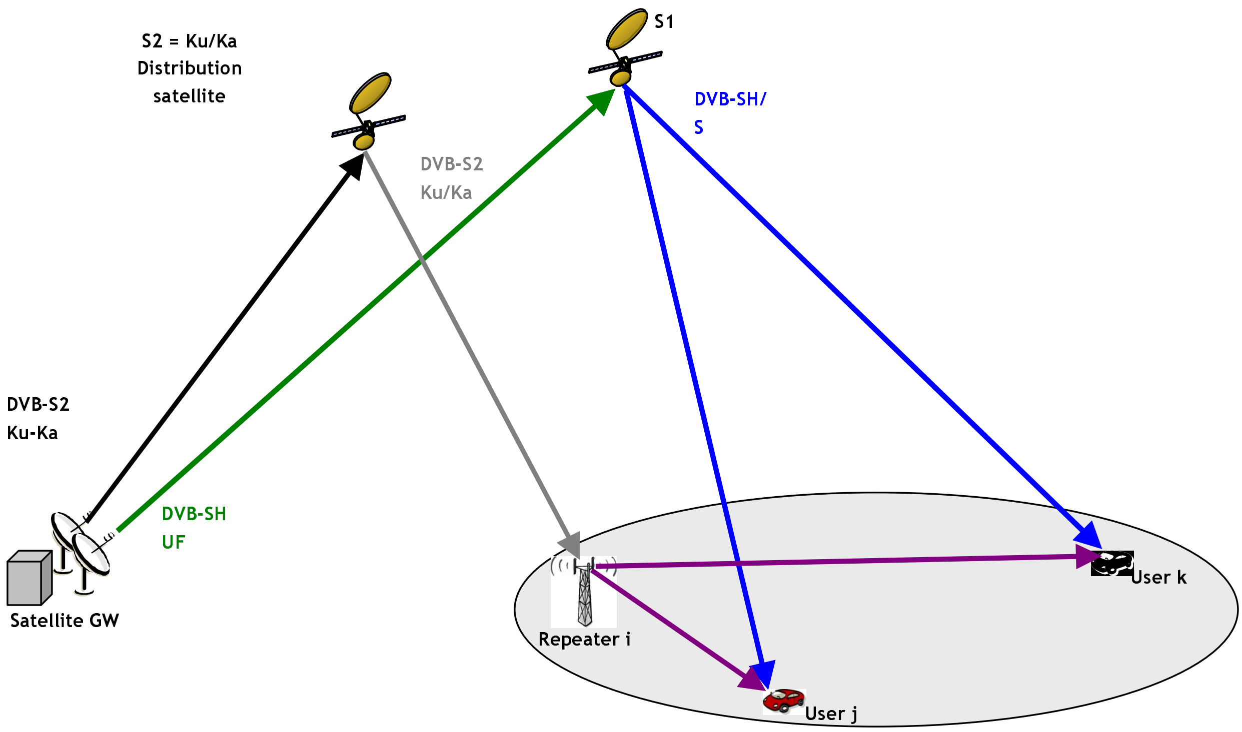


Figure 9.2: Overall system

The satellite apparent motion cause propagation time variation on the gateway-satellite path and on the satellite-repeater path (and on satellite to any location in the cell path). Depending on the satellite orbital characteristics and coverage considered in the hypothesis, the time delay variations may range to about ±4 ms in the whole coverage. A pre-compensation of the time delay variation must be done at the Gateway location. It is ensured by comparing the locally received signal with a local reference delayed by twice the gateway to satellite reference position path duration. This pre-compensation is generally sufficient for GEO satellites. For near GEO satellites, this pre-compensation reduces significantly the time delay variations on the coverage, however a differential delay of up to about ±400 µs remains between the direct and indirect paths. These differential delay variations must be compensated for at the level of each repeater using satellite ephemeris position to reach 10 % of the guard time (around 11µs in QPSK 5 MHz 2k GI=1/4). The technique used to convey this information is out of the scope of the current standard, however data could be sent by in-band techniques like SHIP private section.

In a near GEO case with 6 ° inclination, the following absolute delay variations can be observed, as shown in following figure, during 24 hours. Two repeaters at different locations are used.

- DT1 : delay variation : satellite to repeater1 minus nominal value;
- DT2 : delay variation : satellite to repeater 2 minus nominal value;
- DTG : delay variation : satellite to Gateway minus nominal value.

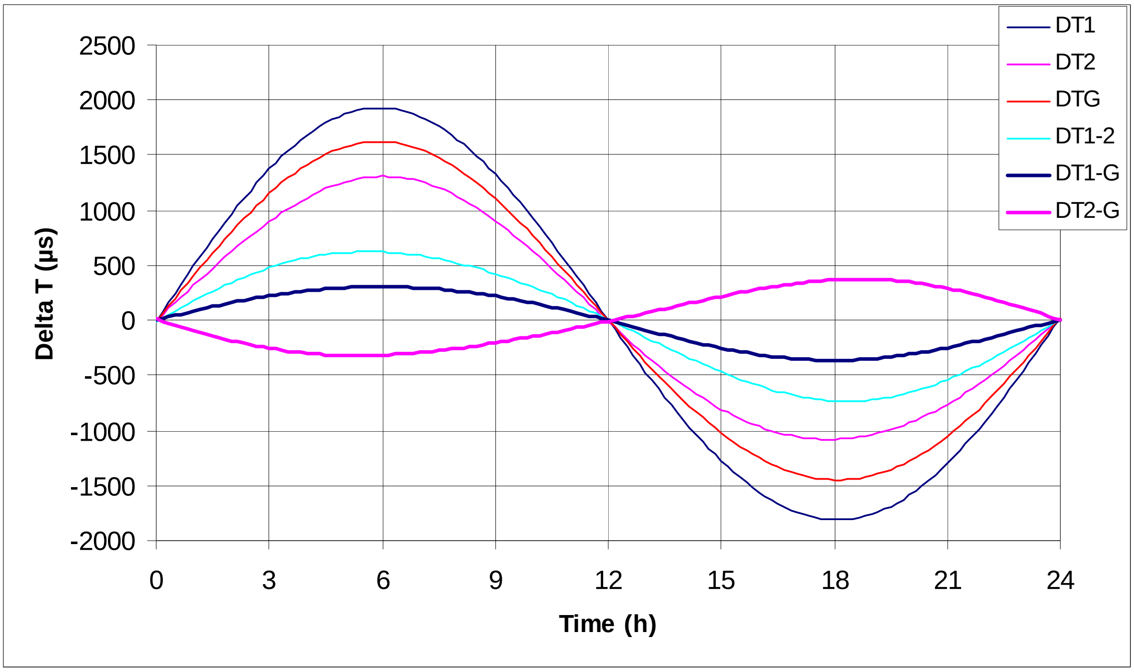


Figure 9.3: Examples of absolute delay variations (one way).

The differential following delay variations are shown in next figure.

- DT1-G = DT1-DTG;
- DT2-G = DT2-DTG;
- DT1-2 = DT1-DT2.

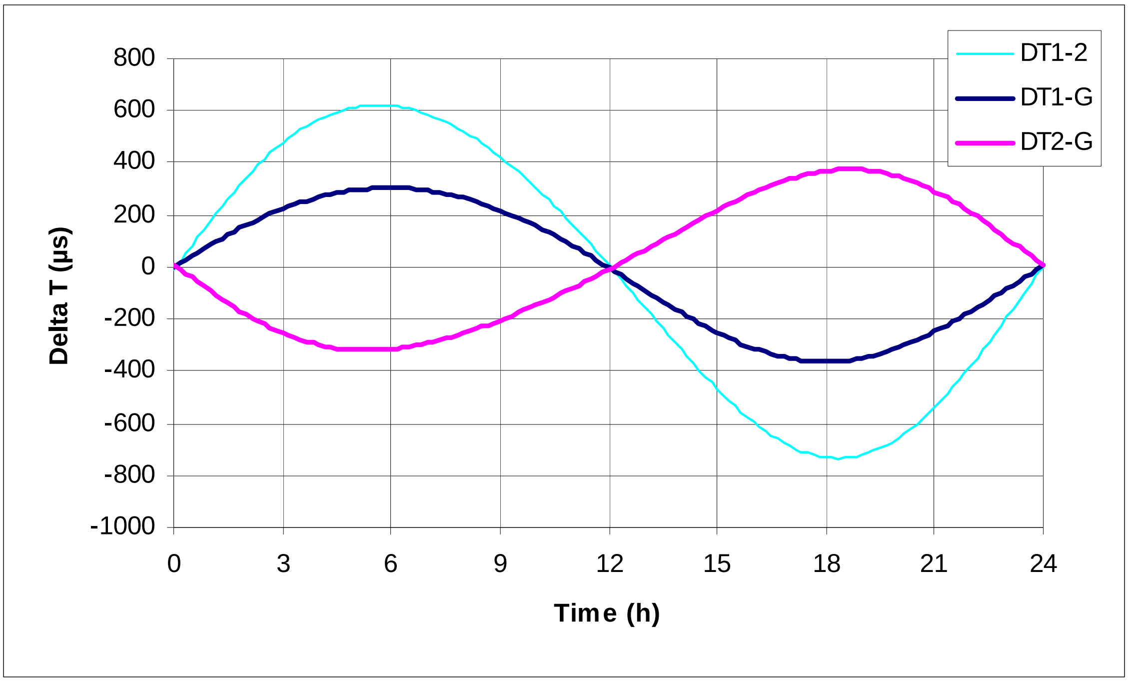


Figure 9.4: Examples of differential delay variations

9.2.3.2.2.2 Frequency shift

In parallel to the time variations, the satellite apparent motion results in Doppler shifts of the frequency by about a few hundred Hz (typical value circa ±300 Hz) which also require compensation. The absolute Doppler variations are shown here below.

- DF1 : Delta F satellite/repeater 1;
- DF2 : Delta F satellite/repeater 2;
- DFG : Delta F satellite/Gateway.

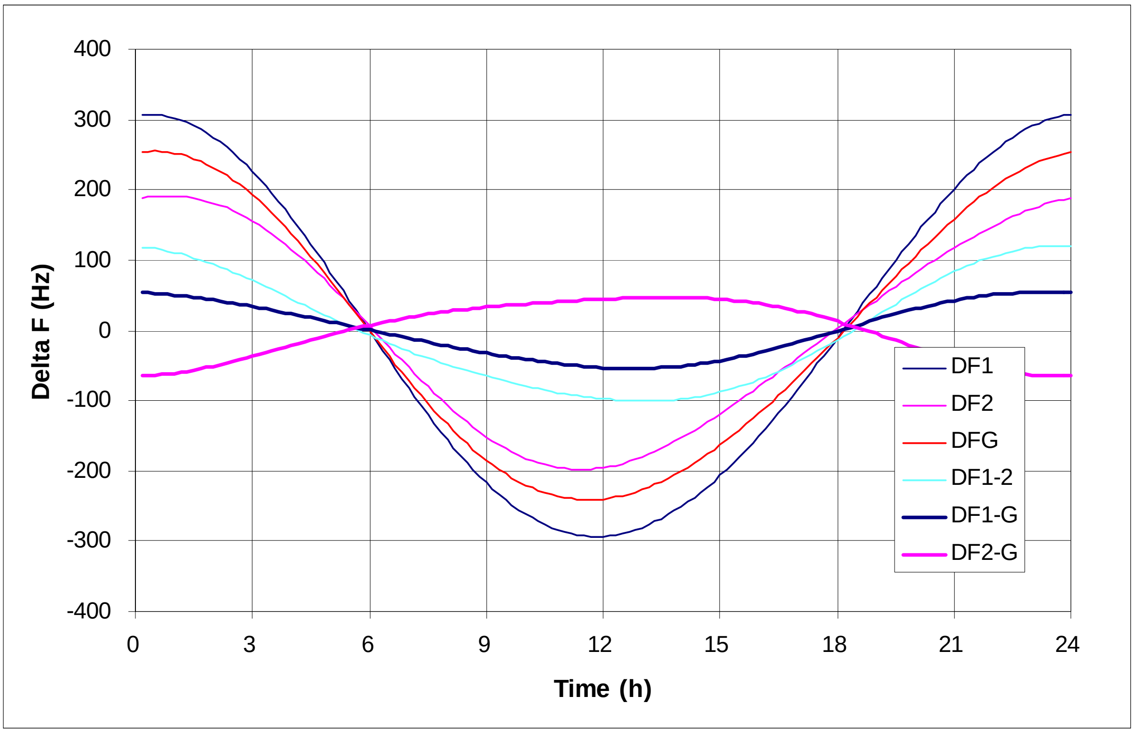


Figure 9.5: Examples of absolute Doppler variations (2,2 GHz on downlink)

A pre-compensation of the Doppler shifts must be done at the Gateway location. It is ensured by comparing the locally received signal with a local reference. This pre-compensation is generally sufficient for GEO satellites. For near GEO satellites, this pre-compensation reduces significantly the Doppler shifts on the whole coverage, however it remains a differential delay up to about ±100 Hz. These differential delay variations must be compensated for at the level of each repeater using satellite ephemeris velocity data the same way the time compensation is achieved.

9.2.3.2.3 Architecture

The satellite gateway provides signal processing for the satellite direct and indirect paths, and elaborates all SFN-related data required by the CGC elements (repeaters). The main services offered by the gateway are:

- transmission of data received from the IP Encapsulator towards the User Equipment via the satellite (S-Band), and towards the terrestrial repeaters through a commercial distribution satellite (Ku/Ka-Band);
- global SFN control and Time and frequency shifts corrections computation in a synchronized way between the satellite path and the repeater path.

The same MPEG TS are broadcast by S-band satellite (S1) and by a commercial satellite (S2). The MPEG-TS transmitted by the gateway to S1 will be DVB-SH modulated. This DVB-SH modulated signal, after transposition by S1 to a S-band carrier frequency f_o, is directly received by the User Terminals (direct path) under S1 coverage. The same MPEG-TS transmitted by the gateway to S2 will be DVB-S2 modulated in the Ku/Ka-band. S2 retransmits this signal in the Ku/Ka-band. The repeaters under the S2 coverage then demodulate the received DVB-S2 signal, and retransmit the extracted MPEG-TS using the same DVB-SH modulation scheme at the same previous S-band f_o frequency: the terrestrial DVB-SH modulated signal are then be also received by the User Terminal.

The SFN time and frequency synchronization processing in the gateway as illustrated by figure 9.6 consists in:

- comparing actual received S band signal with an internal reference;
- compensating time and frequency in the DVB-SH modulator sending to S1 ensuring perfect time and frequency compensation for the gateway location; and
- delivering to the repeaters through adequate means the relevant ephemeris information (one means could be to introduce such ephemeris into SHIP private functions that is sent over the global MPEG2 TS) for correcting remaining differential time variation and frequency shift, if required, at the repeater level.

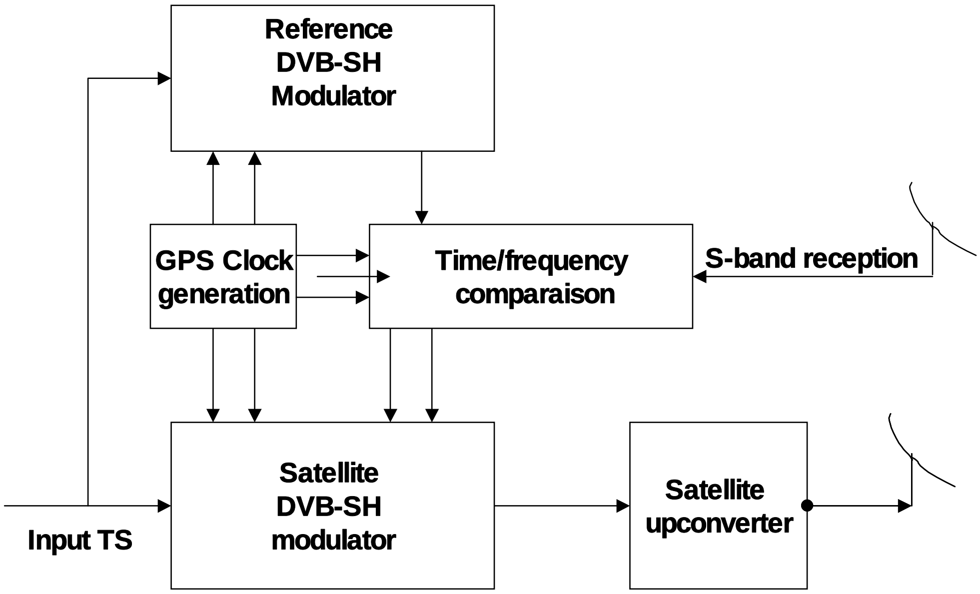


Figure 9.6: Hybrid SFN satellite gateway modulator

9.3 Signalling

As presented in clause 4, different signalling information are used in a DVB-SH system:

- physical layer: SHIP (all modulations), TPS (OFDM) and Signalling Field (TDM) are used;
- section: PSI signalling;
- link: MPE and MPE-IFEC;
- higher: ESG, files, etc.

Relevance of SHIP, TPS and SF signalling are detailed in clause 7. Relevance of MPE and MPE-IFEC signalling are detailed in clause 6. ESG are protected by MPE-IFEC, files are protected by IPDC content delivery mechanism. So this clause concentrates on PSI signalling relevance. The Service Information (SI), is not protected by the MPE-FEC so SI reception quality will heavily depend on burst and error rates witnessed before MPE-IFEC decoding.

The main PSI/SI needed by an IPDC/DVB-H terminal are:

- **NIT** Network Information Table;
- **INT** IP/MAC Notification Table;
- **PAT** Program Association Table;
- **PMT** Program Map Table; and
- **SDT** Service Description Table.

Table 9.1: PSI/SI tables reception performance

Table	Average size (bytes)	Size (TP)	Repetition period (ms)	Reason for acceptable performance quality	Remarks
NIT	3 000 to 6 000	20 to 37	10 000	Static	Size depends on number of TS and cells
INT	500 to 13 000	4 To 72	30 000	Static over HF Repetition increased over NHF	Size depends on neighbour cell signalling policy and scenario
PAT	24 to 64	1	100	Repetition period	Size depends on scenario (SH-A/B)
PMT	140 to 1 400	1 to 8	100	Repetition period	Size depends on scenario (SH-A/B) and local content policy
SDT	50 to 450	3	2 000	Static over HF Repetition increased over NHF	Size depends on scenario (SH-A/B) and local content policy (service_availability)

Concerning PAT/PMT, the size of the respective tables and the repetition rate (10 times per second) are such that robustness is not a problem, even in ITS channels where TS PER of 10 % are typical. SDT, NIT and INT tables could be quasi-static and the receiver does not have to actually receive them each time it is switched on or each time it performs handover, provided the tables are stored in the receiver.

Case 1: *SI tables do not depend on geographical location within a country*

This case is applicable for the hybrid frequency. In this case, NIT, INT and SDT are defined to cover a full spot beam so they could be stored in the receiver and updated only when they have changed. Any change of content would then be signalled in the PMT so a receiver would immediately be aware of this and could start downloading the updated tables. Robustness will not be a problem in this case. In this case, a receiver in SI acquisition mode would likely be tune on during the acquisition time in order to optimize the reception probability. In typical ITS environment, the average TP error rate is around 10 % to 15 % but these errors are distributed over 20 % of bursts (or 80 % of burst are error free). By listening continuously to burst, the receiver will quickly acquire the table since the repetition interval is in the order of a few seconds.

Case 2: *SI tables do depend on geographical location within a country*

This case is applicable for non hybrid frequencies where SI tables may differ between different areas of the network. It may be necessary to actually receive these tables more frequently. This is particularly true for the INT and the SDT, since the NIT is unique throughout the network and will not change (except when roaming between network, case not described here). If one can assume that the receiver accesses the SI each time it is switched on and in connection with each handover, then this SI has to be receivable in a robust way without too much delay (preferably before the next burst). Due to the fact that the SI is tailor-made for the specific area, the size of the SI tables can be made highly limited in size. It is then possible to repeat the tables much more frequently than the required minimum (every 2 s form the SDT, every 10 s for NIT, every 30 s for INT). If the table sizes are small, they could be repeated, e.g. every second, and the probability of correct reception would increase dramatically thanks to the redundancy provided by the repetitions. This optimization procedure is also detailed in EN 302 304 [3], clause 9.3 and PSI/SI tailoring will be detailed in the DVB-SH PSI specifications and implementations guidelines (to be published).

In conclusion, the PSI/SI to be used for DVB-SH will most probably be quasi-static over the hybrid frequency, since all content related information is sent over IP. This information can, therefore, in principle, be stored in the receiver, which will make access much less time critical. By the redundancy of the repetition of the PS/SI tables, correct reception is guaranteed, sooner or later in all reception conditions. In cases where fast access of SI, not previously stored, is required, this can be accomplished over non-hybrid frequency by increasing the repetition rate of the NIT, INT and SDT. With a repetition rate of, for example, one second, correct reception of SI tables can be obtained within a few seconds, also in very bad channel conditions.

9.4 Considerations on the use of repeaters and their feeder links

This clause contains guidelines and recommendations for the use of repeaters (Complementary Ground Components) in DVB-SH networks.

9.4.1 TR(a) On-channel regenerative repeaters.

The TR(a) CGC on-channel regenerative repeaters are broadcast infrastructure transmitters which complement reception in areas where satellite reception is difficult especially in urban areas. They are basically designed as any broadcast transmitters. Local content insertion is possible.

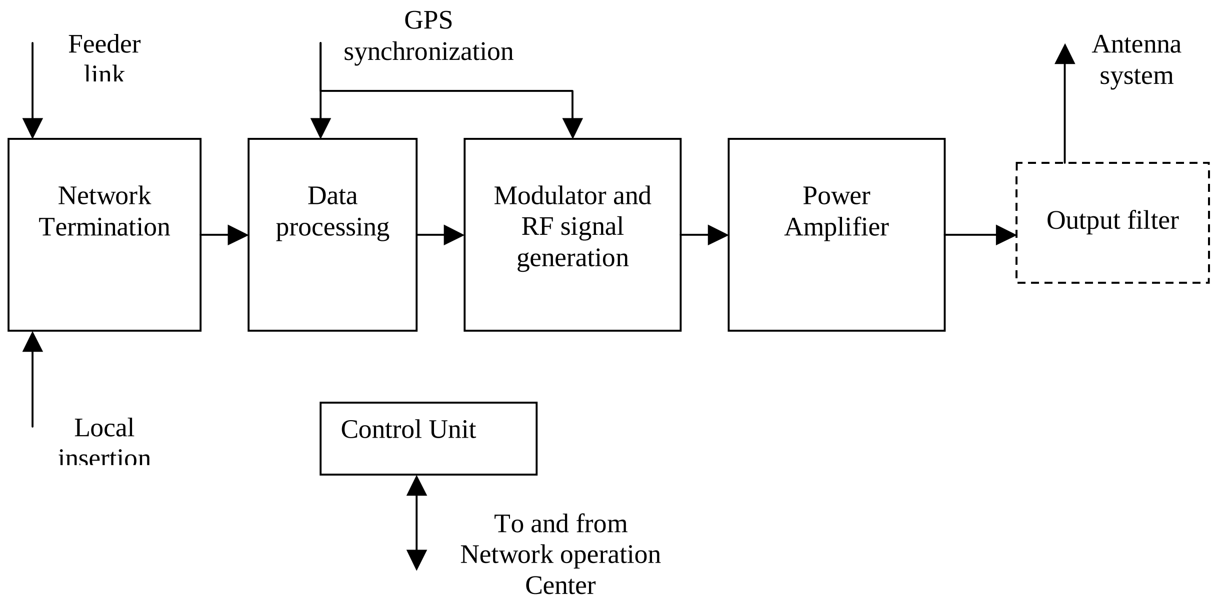


Figure 9.7: TR(a) functional block diagram

The TR(a) repeaters are composed of the main following functional blocks:

- A **network termination** function that interfaces with at least one feeder link. This feeder link ensures the backhauling function and brings the MPEG TS flows to the repeater in such a way that the SFN synchronization remains possible. The backhauling can be of any appropriate kind providing that the Quality of Service is compatible with the overall target (typically ten times better than the repeater to subscriber QoS). This aspect of feeder link QoS is not part of these guidelines. One of the most appropriate backhauling means is via a Ku or Ka-band geo-stationary satellite transponder (DVB-S or DVB-S2) thus offering the advantage of ensuring a wide area coverage of any number of repeaters. Other backhauling means are: Optical fibre network, microwave link, high speed DSL, etc. A mix of any compatible means is possible provided that the overall latency remains compatible with the SFN alignment. The network termination function ensures the interfacing with the backhauling network at the PHY and MAC levels. For adequate network management the network termination also ensures the monitoring of the quality of the incoming signal and provides the information to the control unit. A repeater can include several types of interfaces for either ensuring redundancy between feeder links or for the insertion of local contents or both.

- A **data processing function** that ensures the extraction of the selected bouquet(s) of MPEG TS from the backhauling flow(s). This data processing function might also be able to re-multiplex the MPEG TS in order to optimize the use of the feeder link. In particular local content can be extracted from the feeder link to generate the local multiplex. In that condition, only the envelope of all local contents is sent rather than the sum of all local multiplexes. This data processing function also achieves the SFN synchronization after extraction of the mandatory parameters of the SHIP packets. The time reference for the SFN synchronization is typically a 1 pps GPS pulse (or Galileo in the future). In case of loss of the time reference signal the TR(a) should stop transmitting (no radiated signal) or hold the transmission as long as the time shift can be guaranteed to be lower than 25 % of the guard interval and then stop transmitting. The extraction of the optional SHIP parameters is advisable. Note that the SFN synchronization as derived from the DVB-T/H assumes that the S band satellite is geo-stationary and that its residual relative movement can be compensated by the broadcast head end.

- A **control unit** function or **Operation and Maintenance** function that monitors and controls the repeater and is able to be linked to a centralized network management. If not linked to such a network operation centre the control function should be able to record the key parameters in such a way that the history can be later recovered for analysis. If no link to the network operation centre is possible then the use of optional SHIP parameters could be envisaged.
- A **modulator** function that generates the COFDM waveform followed by an up-converter or directly at RF. The RF frequency must also be synchronized by means of a GPS (Galileo) signal to ensure the correct SFN function.
- A **power amplifier** or set of power amplifiers to amplify the carrier signal up to the desired level. The output power limit is determined by the applicable safety regulation on one hand and to the radio planning target on the other hand.
- An optional **output filter** in order to comply with the output spectrum limits

NOTE 1: For example, ETSI EN 302 574-1 in the case of the S-UMTS band.

- The **antenna system** that radiates the RF signal.

Typical TR(a) parameters:

- Feeder interfaces:
  - L-band interface from Ku or Ka-band LNB (950 MHz to 2 150 MHz)
  - ASI
  - Ethernet (10/100 baseT or Gigabit Ethernet, optical or electrical)
- Synchronization:
  - GPS receiver
- Output power (S-UMTS case):
  - More than 45 dBm per carrier per cell site
- Other characteristics:
  - MER : > 23 dB

NOTE 2: The MER (Modulation Error Ratio) value includes all the signal impairments as measured at the antenna level. The indicated value assumes that there is no non-regenerative means between the repeater and the subscriber terminal or that a non-regenerative devices inserted has no significant effect on the MER value seen by the terminal.

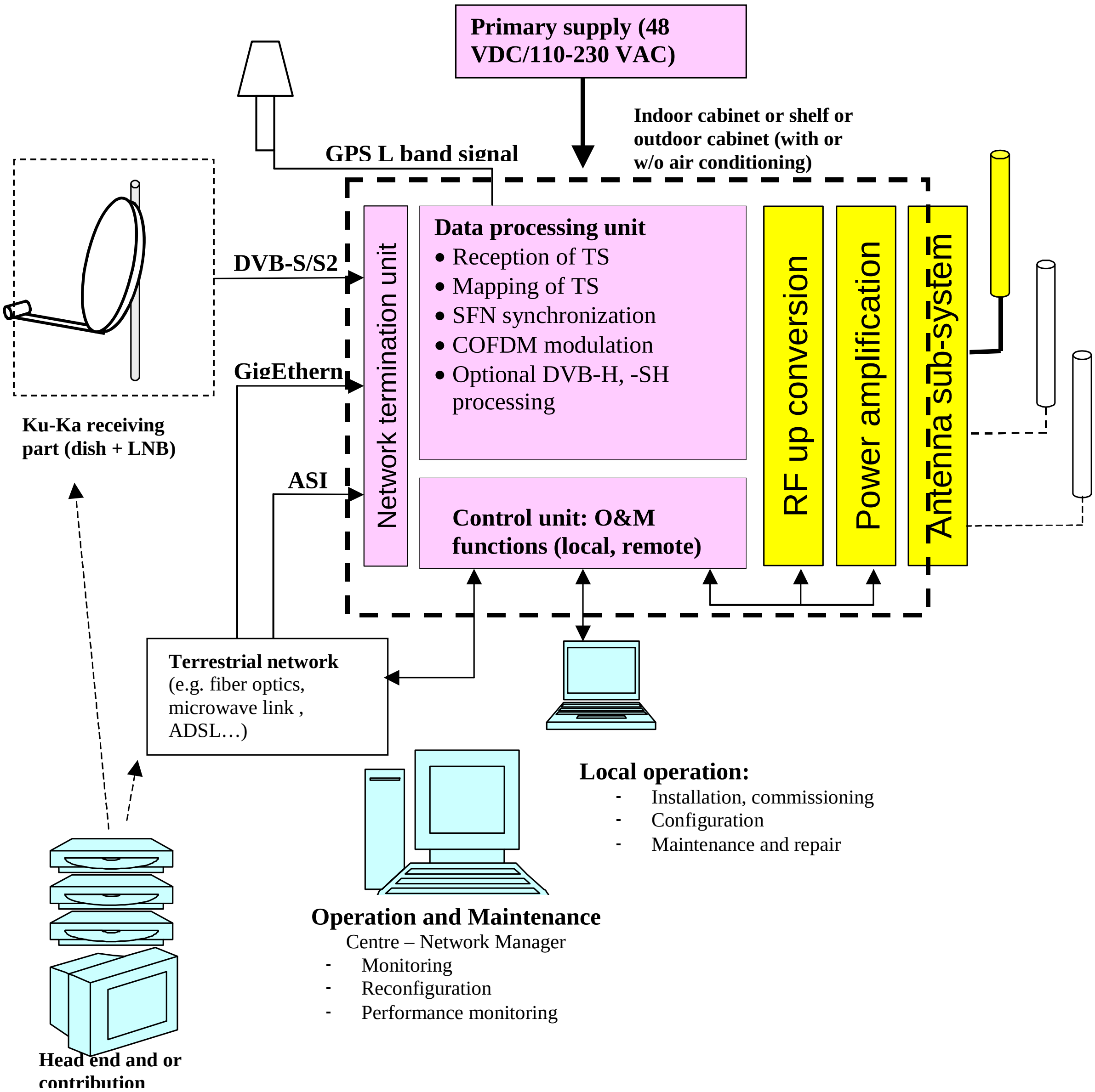


Figure 9.8: Example of implementation diagram of TR(a)

9.4.2 TR(b) Non-regenerative gap-fillers

9.4.2.1 Generalities

The non-regenerative gap-fillers (or simply gap-fillers) are personal or public devices of limited coverage providing local on-frequency re-transmission and/or frequency conversion; typical application is indoor enhancement under satellite coverage; no local content insertion is possible. The main benefits of the gap-fillers, when compared to regenerative repeaters, are easier deployment and lower cost. The delay induced by the whole process of reception, amplification and transmission must be substantially shorter than the guard interval of the used DVB-SH mode (a typical delay is 5 µs), so that a receiver receiving both signal from a gap filler and signal from a satellite or a regenerative repeater does not have to deal with interference but with a constructive addition of signals. Two difficulties arise when using a non-regenerative gap filer is the **filtering** and the difficulty to ensure the **remote management** of the gap-fillers when the deployment and operation have to be kept simple and economical.

9.4.2.2 Filtering issues in non-regenerative gap-fillers

The main obstacle in the deployment of on-channel gap-fillers is a problem inherent to its logic. The transmitted signal may be fed back to the input of the gap-fillers, thus creating a feedback-loop which generates two kinds of problems: ripple in the transfer function of the device, and, at worst, instability of the device.

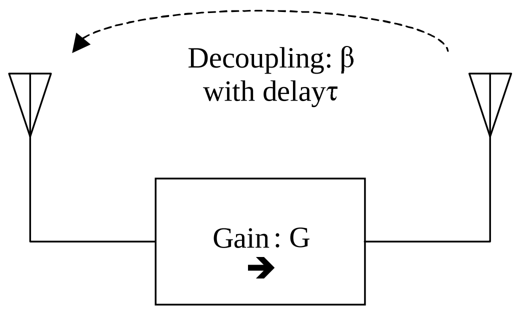


Figure 9.9: Illustration of an on-channel gap-filler coupling effect

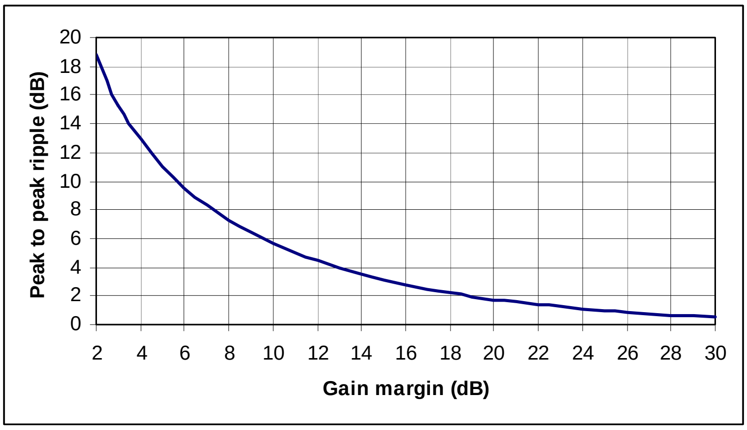


Figure 9.10: Peak to peak ripple as a function of gain margin

The amplitude ripple value in dB is given by:

$$R = 20\log\left(\frac{1+10^{-\frac{M}{20}}}{1-10^{-\frac{M}{20}}}\right)$$, M: margin in dB = -20 logGβ (with βG < 1)

There is also a group delay distortion and ripple but of negligible impact in OFDM: if the relative delay between input signal and re-injected loop-back signal is τ then the time delay variation in function of the frequency *f* is given by:

$$\theta = -\frac{d\varphi}{d\omega} = \tau.G\beta\frac{G\beta - \cos 2\pi f\tau}{1+G^2\beta^2 - 2G\beta\cos 2\pi f\tau}.$$

Common values are 20logG around 70 dB and -20logβ greater than 80 dB.

Some on-channel gap-fillers include internal echo-cancellers in their devices. This element adds an internal decoupling to the external decoupling, thus allowing for a higher total effective decoupling. Improvements of more than 15 dB have been reported. The use of echo-cancellers in on-channel non-regenerative repeaters thus brings two benefits:

- the deployment of repeaters in sites where it would otherwise reveal itself unfeasible; and
- a reduction of in-band distortion thus improving quality of the received signal in the area covered by the repeater.

If no filtering is included in the gap-filler transfer function then other channels and also close or adjacent channels belonging to other services will be re-transmitted thus impacting the coverage of these other services. Proper site spectrum survey is thus necessary. In certain cases however this on the contrary can be advantageously used in the sharing of gap filling infrastructures between different services or applications. Typical examples are subways, tunnels and large indoor public areas (commercial malls, airports).

Regarding the gap-filler itself it is also important to note that protection filtering might be required in order to protect it against strong nearby sources that could drive it into non-linear behaviour and thus non-linear distortion and inter-modulation generation.

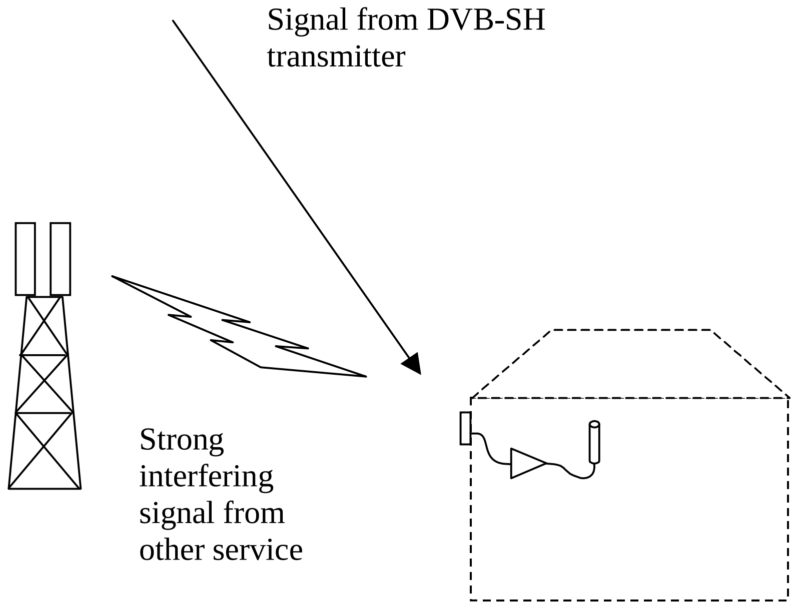


Figure 9.11: Filtering issues in non-regenerative gap-fillers

Recent gap-filler designs include efficient filtering in order to properly select the channel(s) to be retransmitted but this does not always overcome the issue of input overloading by strong out of band signal.

9.4.2.3 Frequency synchronized transposing non-regenerative gap-fillers

Such gap-fillers receive a signal on frequency F1 and re-transmit it on frequency F2 without any SFN resynchronization apart from the F2 carrier frequency. Demodulation at F1 then modulation at F2 of the received signal is not recommended so as to maintain the through time latency as low as possible. The carrier frequency F2 has to be synchronized to ensure the SFN between all the transposing gap-fillers this can be efficiently done by using a GPS locked frequency reference.

9.4.3 TR(c) Mobile transmitters

These transmitters are used to build "moving complementary infrastructures" on-board public transportation such as trains. Depending on waveform configuration and radio frequency planning, local insertion may be possible. Depending on the application both types of repeaters/gap-fillers can be employed for mobile coverage: regenerative and non-regenerative transmitters. In all cases it must be kept in mind that temporary interference with fixed repeaters might occur.

9.4.3.1 Regenerative TR (c)

Obviously if local insertion is required then regenerative TR(a) type is needed. A typical service that could be offered e.g. on-board trains could be pre-recorded programs locally inserted. That kind of service would be similar to what is offered now on-board planes. The backhauling of the main content is the most critical issue since it requires a relatively high data rate link between the vehicle and the fixed network. However depending on the desired capacity this will be easily achievable and actually achieved for other data services (WiMAX). Because there is no need for SFN with fixed sources it is also possible in that particular application and in the case of a hybrid network to recover the signal from the S-band satellite preferably by means of a relatively high gain antenna (6 dBi to 8 dBi) on the roof of the mobile and to demodulate this signal before re-modulation on the same frequency. The re-generated signal would then be redistributed through a radiating distributed local installation at a much higher level compared to the residual signal coming direct from the satellite.

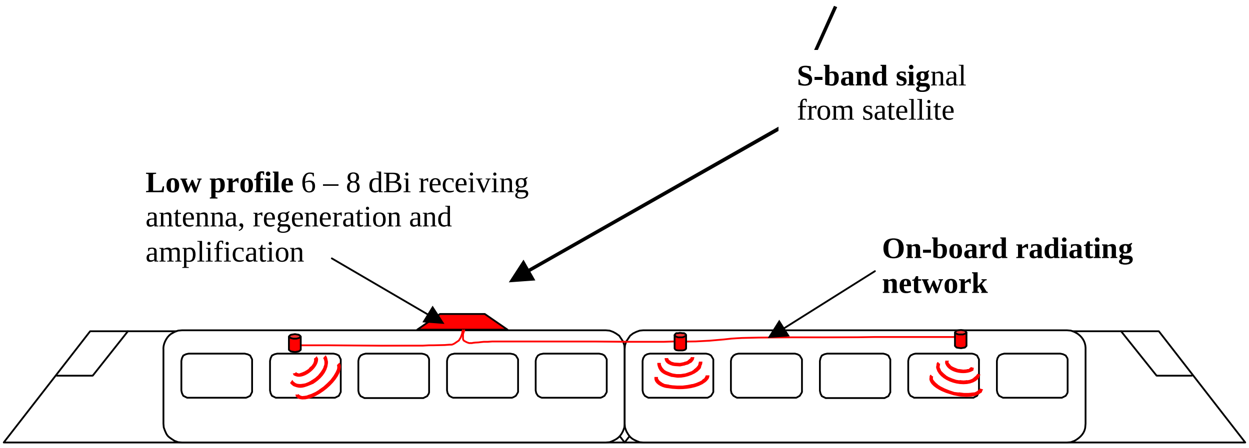


Figure 9.12: Example of possible TR(c) implementation on-board train

9.4.3.2 Non-regenerative TR(c)

If no local insertion is required then a non-regenerative solution could be envisaged. However attention has to be paid in the link budget and final C/N value depending on the main receiving antenna G/T. All the other drawbacks of non regenerative on-channel solutions (see TR(b)) apply but indeed in that case a shared installation with 3G networks could be contemplated. A transposing non-regenerating solution would help in solving the isolation constraint of the on-channel gap-fillers but the frequency availability or local usage might not be possible.

10 Reference Terminals

10.1 Top-level design considerations

10.1.1 Terminal categories

To address a wide range of market sectors, DVB-SH allows a large freedom in terminal implementations. Three main categories can be identified and are considered hereafter :

- category 1: car-mounted terminals (also called "vehicular");
- category 2: portable TV devices with 2 sub-categories:
  - 2a: large screen (≥ 10") portable devices, battery or mains powered;
  - 2b: pocketable (handheld) TV devices, mainly battery powered;
- category 3: handheld terminal with embedded cellular telecom modem (or "convergence" terminal).

Car-mounted terminals can especially benefit from the nation-wide coverage and allows designers to include many of the advanced features of DVB-SH (Seamless complementary satellite/CGC coverage, SH-B configuration, high-order modulations, long time interleaver, etc.).

Portable TV devices have large screen and are mainly stationary during reception. They could have attached antenna but also detachable antenna accessories. This latter case allows high reception performances thanks to optimization of the antenna position by the user (i.e. find a LOS reception from satellite or optimized position for good reception from CGC).

Handheld terminals can especially benefit from the outdoor and indoor coverages in built up areas (similar to 3G coverages) but are more challenging due to their small form factor, the large number of functions to be integrated, limited battery power and coexistence with other active radio functions.

Category 2b has common characteristics with handset. Most often, pocketable terminals embed a large number of multimedia features without necessitating coexistence with radio modems.

Other features which are at the discretion of the manufacturers/markets include:

- antenna dedicated and/or optimized to satellite link;
- number of antennas and branches for diversity gain;
- power of processor embedded in terminal;
- embedded memory for physical layer processing;
- host memory for PVR functions.

10.1.2 Mobility aspects

10.1.2.1 Mobile channels

DVB-SH receiver mobility environments are defined in clause 11.

10.1.2.2 Antenna and diversity considerations

Due to the relative narrow bandwidth of the frequency bands considered (see clause 4.1.3.1 on frequency allocations), antenna performances could be better than that of their wide-band UHF counter part. Of course, this statement is not valid in the case of multi-band antenna (antenna shared between different missions on different bands).

For handheld with integrated linearly polarized antenna, -3 dBi gain can be achieved with omnidirectional pattern (for example at 2,2 GHz/30 MHz bandwidth). If the antenna is external, antenna gain can be as high as 0 dBi for linear (0 dBic for circular antennas and omnidirectional for both). Polarization choice depends on targeted terminal (optimization for satellite link or not).

For vehicular or external dedicated antenna, 4 dBic gain can be reached at elevation optimized to target satellite orbit and position (for example, antenna for SDARS system). Further details are given in clause 10.4.4.

Antenna diversity reduces the effect of the fast fading Rayleigh-channel. In an antenna diversity receiver, output signals obtained from several antennas are linearly combined using adjustable complex-value weighting factors before being decoded. Implementations may differ:

- by the antenna system's characteristics: number of antennas, relative positions, orientation and characteristics of each antenna (polarization, radiation pattern, etc.);
- by the algorithm used to compute and eventually iteratively adapt the weighting factors.

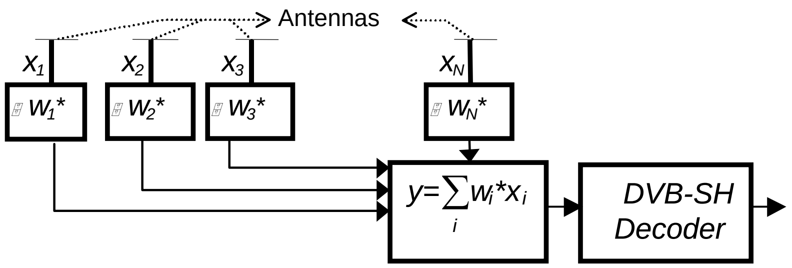


Figure 10.1: Antenna diversity receiver

In mobile reception conditions (especially for Terminal Category 1), antenna diversity gain is expected to permit a reduction of terrestrial repeaters transmit power by up to 6 dB for the same coverage. It should also allow increasing the mobile's maximum speed for correct reception. This result is given considering a multipath Rayleigh channel, and no long time interleaver benefit is considered. Results on cumulative diversity and interleaver gains are not yet available.

For satellite reception, 2,5 dB diversity gain can be achieved in LOS reception environment.

Of course, diversity with higher order (3 or 4) could be envisaged on Terminal Category 1.

In indoor reception (especially for Terminals Categories 2a, 2b and 3) the channel conditions are changing slower than for mobile reception. However, considering Rayleigh environment, diversity gain could be as high as 6 dB. First results show this level of performances. Nevertheless, test conditions have to be defined precisely and antenna system implementation is of prime importance. Indeed, some impairments could affect diversity gain like gain unbalanced between antennas, correlation between fadings depending on antennas respective positions on the terminal and considered environment.

10.1.3 Service aspects

As for DVB-H, a trade-off has to be made between burst length, time-slicing off time period, power saving effectiveness and access time.

Receivers should minimize "wake up" time (time between RF switch-on and start of demodulation) by implementing various techniques such as keeping and using formerly acquired demodulation parameters, using fine evaluation of the off time period, etc.

As for DVB-H, time slicing allows seamless handover by scanning adjacent cells signals during idle time periods. Another aspect in DVB-SH is the receiver Class (memory, see also clause 10.2) and its associated time interleaver depth capability.

Issues to be considered are:

- backward compatibility of receiver Class 1 with long interleaver used by the transmitter;
- time slicing burst length / off time period;
- access (or zapping) time.

Coexistence (especially for Category 3 receivers)

For handheld receivers with embedded cellular telecom modem, DVB-SH reception should not prevent or hinder cellular operations (initial synchronization, roaming, handover, communication). Reciprocally, reception of DVB-SH service should be possible in parallel with cellular operations. However, DVB-SH quality may be degraded by cellular uplink signal (see clause B).

Two states may be considered:

- cellular function in standby, i.e. intermittent uplink transmissions (cell search, handover in mobile conditions): DVB-SH reception if active should provide nominal picture quality; and
- cellular function in active communication, i.e. uplink traffic transmission: DVB-SH reception if active may or may not be displayed, depending on user selection. When DVB-SH service display is switch off during call, demodulation should continue in the background and display should be automatically resumed at the end of the cellular call.

Coexistence conditions described in clause B consider a worst case:

- DVB-SH full service while cellular handheld operation: standby and communication;
- low level DVB-SH downlink signal;
- maximum power cellular uplink;
- no DVB-SH time interleaver protection.

10.2 Memory requirements for DVB-SH processing

The DVB-SH receive baseband processing may follow a similar architecture as that of the DVB-H counterpart, namely analog-to-digital converter, OFDM demodulator, de-interleaver, Forward Error Correction (FEC), demultiplexing, multi protocol de-capsulation, MPE-FEC, IP filter and terminal host interfacing. New modified functions are mainly the Turbo decoder and the physical layer time de-interleaver.

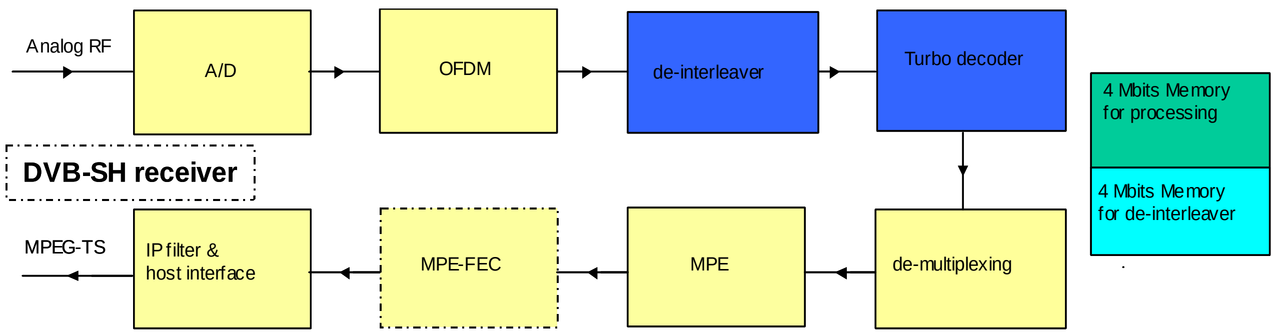


Figure 10.2

The Turbo decoder needs extra hardware which is available as an off-the-shelf building block from 3G technology development.

The main challenge from the Physical Layer Interleaver is its memory requirement which is dependent to the DVB-SH Receiver Class:

Class 1

Receivers Class 1 are required to handle an interleaver profile comprising one full SH-frame with 816 CU. Using the convolutional interleaver approach which typically halves the amount of memory needed, this transforms to:

- interleaver lengths of up to 240 ms (QPSK, uniform) and 120 ms (16QAM, uniform);
- capability to store and process up to 408 CU or 6 528 IU (can be handled with approx. 4 Mbits of memory).

Since support of DVB-H is likely for DVB-SH receivers, the memory dedicated to the DVB-H MPE-FEC could be allocated to the de-interleaver. In this case, the DVB-H MPE-FEC, if ever transmitted, is not processed. The choice between MPE-FEC processing or interleaving processing should be managed at system level. Therefore, DVB-SH Class 1 receivers could be realized with the same amount of memory than previous DVB-H receivers.

Class 2

Receivers Class 2 are required to handle an interleaver profile comprising 64 full SH-frames with 52 224 CU. Using the convolutional interleaver approach which typically halves the amount of memory needed, this transforms to:

- interleaver lengths of up to 30 s (QPSK, non-uniform) and 15 s (16QAM, non-uniform);
- capability to store and process up to 26 112 CU or 417 792 IU (can be handled with 256 Mbits memory).

10.3 DVB-SH reference receiver model

10.3.1 Reference model

The receiver performance is defined according to the reference model shown in figure 10.3.

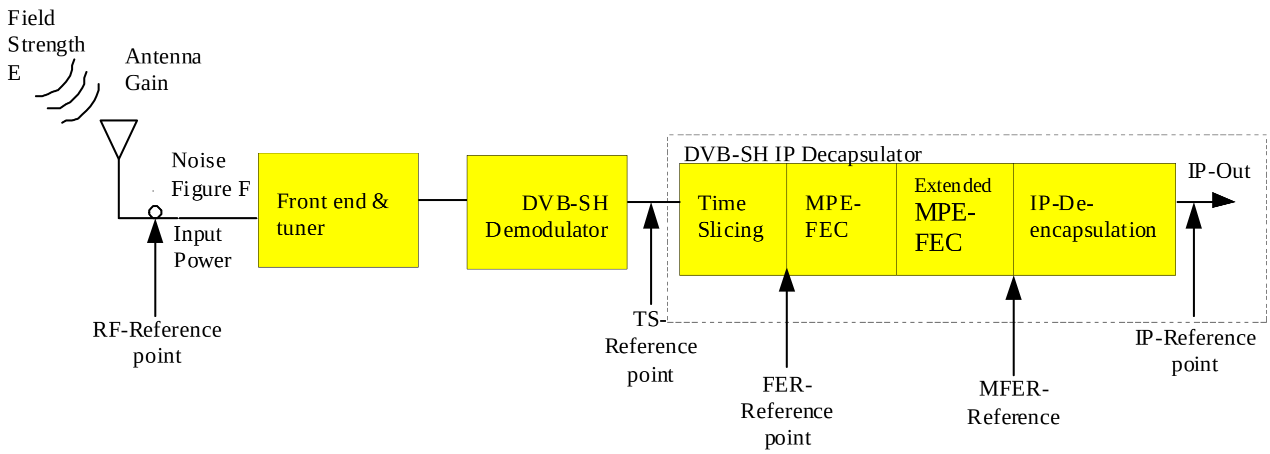


Figure 10.3: Reference model

Reference points are defined for:

- RF;
- transport stream;
- frame errors before MPE-FEC;
- frame errors after MPE-FEC and extended MPE-FEC;
- IP-stream.

All the RF receiver performance figures are specified at the RF-reference point, which is the input of the receiver.

10.3.2 Receiver for Vehicular terminals

Depending on terminal categories, various antenna and front-end solutions will be embedded in receiver. For terminal category 1, according to vehicular reception constraints (clause 4.2.6.1), antenna with optimized diagram pattern and polarization will be used.

Additional Low Noise Amplifier will be connected directly to antenna to optimize noise figure and sensitivity (see details in clause 10.4).

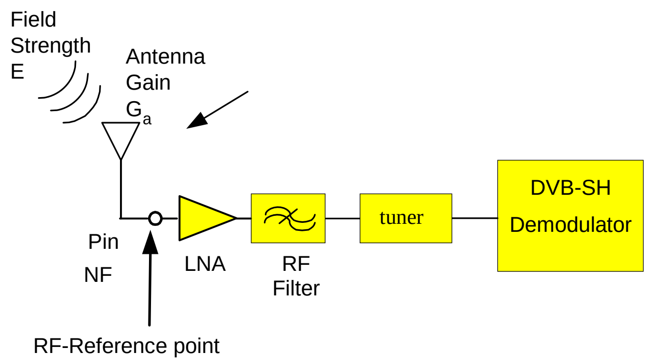


Figure 10.4: Radio receiver architecture for Terminal category 1

10.3.3 Receiver for Terminals with telecom modem

For category 3, if the terminal includes telecom modems, a filter is mandatory between the antenna and the (optional) LNA and/or the tuner to protect the latter elements from telecom signals blockers. This filter however degrades the DVB-SH receiver sensitivity.

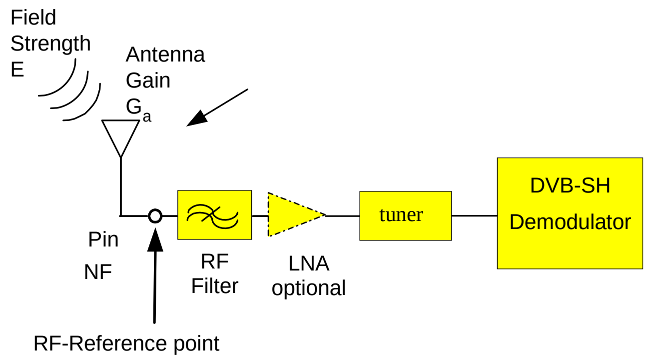


Figure 10.5: Radio receiver architecture for Terminal category 3

Basic receiver architectures for all terminal categories will be derived from both architectures presented above.

For SH-A system architecture, these synopsis will be used without any variants.

For SH-B system architecture, the tuner will have two outputs each of them will correspond to TDM and OFDM signals respectively. These outputs are generated after down conversion into the tuner from two RF different frequencies, one is provided by satellite(s), the other by CGC segment.

Demodulator for SH-B will have two inputs in accordance to figure 4.6 of clause 4.2.6.

For antenna diversity implementation, several receiver branches will be added to single one. After diversity combining, transport stream will be provided to an unique input of DVB-SH IP decapsulator.

10.4 Minimum signal input levels for planning

10.4.1 Noise floor for vehicular receiver

The receiver should have a noise figure better than 2 dB at the reference point, at sensitivity level of each DVB-SH mode.

This corresponds to the following noise floor power levels:

- Pn = -103,2 dBm [for 8 MHz OFDM channels, BW = 7,61 MHz];
- Pn = -103,7 dBm [for 7 MHz OFDM channels, BW = 6,66 MHz];
- Pn = -104,4 dBm [for 6 MHz OFDM channels, BW = 5,71 MHz];
- Pn = -105,2 dBm [for 5 MHz OFDM channels, BW = 4,76 MHz];
- Pn = -110,2 dBm [for 1,7 MHz OFDM channels, BW = 1,52 MHz].

10.4.2 Noise floor for handheld receiver

The receiver should have a noise figure better than 4,5 dB at the reference point, at sensitivity level of each DVB-SH mode

This corresponds to the following noise floor power levels:

- Pn = -100,7 dBm [for 8 MHz OFDM channels, BW = 7,61 MHz];
- Pn = -101,2 dBm [for 7 MHz OFDM channels, BW = 6,66 MHz];
- Pn = -101,9 dBm [for 6 MHz OFDM channels, BW = 5,71 MHz];
- Pn = -102,7 dBm [for 5 MHz OFDM channels, BW = 4,76 MHz];
- Pn = -107,7 dBm [for 1,7 MHz channels, BW = 1,52 MHz].

10.4.3 Minimum C/N requirements

This clause will contain practical C/N requirements when measured implementation losses are available. Until then, any values given below shall be considered provisional. For AWGN, it is recommended to add the following implementation losses to the C/N values given in clause 7.

Table 10.1

Modulation	Implementation loss (dB)
TDM- QPSK	0,5
TDM- 8PSK	1,0
TDM- 16APSK	1,5
OFDM- QPSK	1,1
OFDM- 16QAM	1,5

For TU6, the recommended implementation loss targets are 1,6 dB for QPSK and 2 dB for 16-QAM.

It is recommended to add a provision of 2 dB for implementation losses in Rice and Rayleigh channels to the C/N values given in clause 7.

10.4.4 Minimum input levels

10.4.4.1 Sensitivity for vehicular receiver

At RF-reference point, the receiver should observe reference criterion for a wanted signal level greater than Pn.

- Pn = -103,2 dBm + C/N [for OFDM 8 MHz];
- Pn = -103,7 dBm + C/N [for OFDM 7 MHz];
- Pn = -104,4 dBm + C/N [for OFDM 6 MHz];
- Pn = -105,2 dBm + C/N [for OFDM 5 MHz];
- Pn = -110,2 dBm + C/N [for OFDM 1,7 MHz].

where C/N is specified in clauses 10.4.2.2 to 10.4.2.3 and is depending on the channel conditions and DVB-SH modes.

10.4.4.2 Sensitivity for handheld receiver

At RF-reference point, the receiver should observe reference criterion for a wanted signal level greater than Pn.

- Pn = -100,7 dBm + C/N [for OFDM 8 MHz];
- Pn = -101,2 dBm + C/N [for OFDM 7 MHz];
- Pn = -101,9 dBm + C/N [for OFDM 6 MHz];
- Pn = -102,7 dBm + C/N [for OFDM 5 MHz];
- Pn = -107,7 dBm + C/N [for OFDM 1,7 MHz].

where C/N is specified in clauses 10.4.2.2 to 10.4.2.3 and is depending on the channel conditions and DVB-SH modes.

10.4.5 Antenna considerations

Receiver antennas must have high performance in various environments, receiving signal from satellite or/and CGCs. In this clause, intrinsic and integration characteristics are highlighted depending on the targeted terminal.

10.4.5.1 Antenna for terminal category 1

Antenna for vehicular terminals can be listed in two categories:

- antenna with circular polarization optimized for satellite reception; and
- antenna with linear polarization optimized for terrestrial reception.

Antennas with circular polarization for vehicular terminals have been developed for several systems like Satellite Digital Audio Radio System (SDARS) in US (XM radio, Sirius). These antennas have relatively good circularly polarized gain at high elevation angle and acceptable linearly polarized gain at the horizon. They efficiently receive both satellite and terrestrial signals.

Antenna gain in Left or Right Hand Circular Polarization (LHCP/RHCP) can be +2 up to +8 dBic, depending on angle aperture and directivity required. Antenna gain in linear vertical polarization is less than -1 dBi at horizon. These antenna are designed in various technologies: Small ground plane dependent patch etched on ceramics antennas, ground independent mast antennas such as quadrifilars, turnstile consisting in two 90° crossed dipoles able to generate circular polarization. Voltage Standing Wave Ratio (VSWR) can be less than 1,5 on 50 Ω load for narrow band operation, this performance maximize antenna efficiency. A Low Noise Amplifier (LNA) with a noise figure less than 1 dB can be connected directly at the antenna output to improve sensitivity.

G/T of more than -21 dB/°K can be achieved.

For antenna with linear polarization, the practical standard for car reception is $\lambda/4$ monopole which use the metallic roof as the ground plane. Planar structures may be also used. Due to the use of the metallic roof, antenna pattern (and consequently the gain) is very dependent of the position of the antenna on the vehicle. Considering the 2 GHz MSS band, gain on the whole demi-sphere in linear vertical polarization is more than 0 dBi. Due to depolarization losses, the gain in RHCP is less than -3 dBic.

Other constraints have to be considered: Number of radio systems integrated in vehicle (Telecom modem car kit, GPS..), larger antenna structures can be aesthetically displeasing when mounted on roof or glasses…

Diversity features that increase antenna number and can lead to use a mix between technologies described above.

10.4.5.2 Antenna for terminal category 2a

This clause deals with antenna for Portable digital TV sets. These terminals are defined to be equipped with screen size higher than 25 cm (more than 10" diagonal), stationary during reception and antenna attached to the receiver.

Integration of high performance antennas with circular polarization and high gain is now challenging due to the flat design of the terminal.

The manufacturer will have to make compromise between design and performances. The performances of this terminal could be between the one of category 1 and category 2b,3.

If performances are favoured, by using for example a detachable antenna accessory, the terminal G/T could be as high as to -25 dB/°K (position adjusted by the user in line of sight of the source).

Diversity will be implemented advantageously with regard to mechanical size and potential distance between antennas.

10.4.5.3 Antenna for terminal category 2b and 3

This clause deals with antennas for pocketable digital TV receiver and handheld with embedded Telecom modem(s), both being battery operated.

Two solutions could be met, one with an external antenna, the other with internal antenna. The majority of handset makers design terminals with internal antennas. A first consideration concerning antenna design is volume availability. Also, coexistence with a lot of components decreases expected gain due to dielectric and magnetic losses. This is why, low bulk antennas have less efficiency and more susceptibility to detuning. It is not conceivable to use multi band antenna that provide poor performances and coupling. Narrow band antenna will be considered as the basic antenna reference. For handheld 2b and 3 categories, better reception sensitivity will be reached with external antenna.

Antenna for handheld can also be ordered in two categories:

- antenna with circular polarization optimized for satellite reception; and
- antenna with linear polarization optimized for terrestrial reception.

Circular polarization: the antenna pattern could be optimized for some pre-defined elevations, the main drawback is the position given by the user to the terminal. This leads to favour omnidirectional pattern design rather than directional ones (at least in a hemisphere). The same assumptions have to be done with antenna with linear polarization.

Typical figures are given below for 2 GHz MSS band and in LOS measurement conditions:

- external RHCP antenna: gain 0 dBic (circular), gain in linear –3 dBi;
- internal RHCP antenna: gain -3 dBic (circular), gain in linear –6 dBi;
- external linear vertical polarized antenna: gain in linear 0 dBi, gain in circular –3 dBic;
- internal linear vertical polarized antenna: gain in linear -3 dBi, gain in circular –6 dBic;
- external RHCP antenna removable and adjustable by the user: gain +3 dBic, gain linear 0 dBi (antenna with directivity).

Antenna behaviour in multipath environment is more complex due to depolarization effect and integration of power in a sphere.

10.4.6 Maximum Input Power for Wanted and Unwanted Signals

The maximum allowed input level on the RF-reference point will depend on the antenna characteristics and RF front end architecture chosen (see clauses 10.3.2 and 10.3.3). Therefore, maximum value will differ with terminal categories.

For terminal category 3:

The maximum total average power from the wanted and unwanted signals should be less than +15 dBm. This value corresponds to a decoupling between DVB-SH antenna and modem antenna of more than 15 dB to 18 dB depending of the Telecom frequency band considered.

For terminal category 1 and 2a,2b:

The maximum total average power from the wanted and unwanted signals should be less than –25 dBm. This value corresponds to mobile phone up link emissions close to the DVB-SH terminal with a distance more than 1 meter from the source. Down link signal providing by base stations emissions will be below this value.

The assumptions considering Interoperability with other radio systems, are given in annex B. In particular, real interferer environments of DVB-SH receiver will be highlighted including interfering signals patterns (with number of carriers, level and frequency band). In this annex, recommendations will be done for receiver linearity and selectivity.

10.4.7 G/T considerations

Assumptions:

- vehicular receiver (category 1): NF=2 dB (LNA 0,8 dB, filter 2,5 dB, tuner 3 dB)
- handheld receiver (category 2b and 3): NF=4,5 dB (filter 1,5 dB, tuner 3 dB)

Table 10.2: Receiver typical G/T versus terminal category

Usage	Handheld	Handheld	Portable	Vehicular
Classification	category 3	category 2b	category 2a	category 1
Antenna polarization	L or C	L or C	L or C	C
Tuner	DVB-SH	DVB-SH	DVB-SH	DVB-SH
Antenna gain (dBi) or (dBic)	-3	0	2	4
Total Gain (dB)	-3	0	2	4
Antenna Temperature (K)	290	290	200	150
Active antenna	No	No	Yes	Yes
Noise figure (dB) max	4,5	4,5	3	2
Total noise temperature (°K)	817	817	489	320
G/T (dB/K)	**-32,1**	**-29,1**	**-24,9**	**-21,0**

Polarization losses are defined in the respective satellite and terrestrial budget links.

Table 10.2 gives typical figures of basic receivers (without diversity consideration) related to the terminal categories.

G/T calculations are based on various assumptions and the boundaries between classes of terminal are not so clear-cut.

Active antenna corresponds to an antenna with an attached LNA.

11 Network planning

DVB-SH system is a hybrid system with a satellite and a terrestrial components.

Terrestrial component network planning technique have been developed over the past 50 years for two distinct services

- Classical TV broadcast in UHF, VHF, from for high towers, using high power transmitter.
- Cellular, mobile system from GSM/DCS to UMTS and more recently from WiMAX technology.

Although based on the same physics, practice differ in some extent to the very different markets they address:

- High power for the first one, low/medium power for the second.
- Roof top reception for the first one, indoor handset service for the second one.

It is not the objective of this clause to propose an integrated framework, but to expose the main rules of each world for the network planning exercise.

On the other hand the satellite world, also existing for 50 years, has obviously specificities: line of sight, shadowing, high path loss, and uses Land Mobile Satellite propagation models, for availability computation.

The present document is presenting an overview of different methods of radio network planning and using different dimensioning parameters which values are compiled from different sources and experiences. The proposed values can not be seen as absolute references, especially on the cellular network approach, as the used parameters values can vary from a source to another, from a country to another. That is why a range of typical parameters values is proposed, and some examples provided for network planning

11.1 Introduction to DVB-SH network planning

As DVB-SH system is a hybrid system with a satellite and a terrestrial component, the coverage considered in the DVB-SH network planning can be split in three areas:

- Terrestrial only coverage: defined as the area served only by the complementary ground component.
- Satellite only coverage: defined as the area served only by the satellite component.
- Hybrid coverage: defined as the area where both the satellite and the terrestrial signals can be received simultaneously and potentially combined.

Concerning terrestrial coverage, there are typically two possible methods to perform radio network planning:

- The first one is a typical broadcast approach based on the computation of the minimum field strength requirements, for different classes of terminals and different propagation channels. The broadcaster method has been developed for the use of high transmitting towers with very high power, and aiming at roof top coverage rather than street level receivers.

- The second one is based on cellular network radio planning tools, and is using the minimum received power level as planning criterion. Most of these tools uses only this method, and allow network planners to determine (given a certain transmitting power per site) the number of sites necessary to cover a city with a given quality of service. The cellular network method has been developed for lower tower and also lowers to medium power to cover dense cities at street level receivers, and targeting indoor coverage. This clause will give also in some examples the link budgets with the required transmitter power in some typical examples.

The link between the two methods is insured through the formula below with E representing the field strength and P_{imin} the minimum received power level, and Ga the antenna gain in dBi.

$$E[dB\mu V/m] = P_{s\min}[dB(mW)] - G_a[dB] + 77{,}2 + 20\log_{10} f[MHz]$$

One of the objectives of this clause is to compile long time practices in network planning domain applicable to DVB-SH network planning, that is to say review parameters and their possible value ranges enabling both type of network planning. These approaches can also be extended for hybrid coverage considering the satellite as an additional repeater that allows a) reduce the number or the power of terrestrial sites or increase the quality of the service in the hybrid coverage area.

For satellite coverage, the number of transmitters is dependent on the number of satellites serving the satellite coverage area. While in the terrestrial case we have a cluster of cells serving a geographically limited area, for the satellite the coverage can be a continent-size area served with several linguistic or regional spots. The objective of the satellite network planning is to assess the required satellite power given a certain coverage area and a certain quality of service for a given data rate. This exercise is described in the clause "Satellite Link Budgets", but it is done in a different way as in terrestrial network planning.

- in terrestrial links, given the coverage probability in some conditions, the required shadowing margins are computed and allow to dimension the transmitter;

- in satellite links, given the satellite EIRP in some conditions, the obtained link margins are used in simulations to compute the availability in the given conditions. Thus, by reverse engineering, given a an availability, required satellite EIRP can be estimated.

Therefore, after this introduction, the clause comprises the following clauses:

- definition of reception condition;

- coverage definition, following a parallel track for the three approaches;

- network planning factors.

Then, following also a parallel track:

- network planning methods and examples based on minimum filed strength;

- network planning methods and examples based on minimum received signal level;

- satellite network planning; and

- for next release: hybrid network planning.

11.2 DVB-SH reception conditions

11.2.1 Introduction

Reception conditions depend on the environment, the mobility conditions and the kind of terminal. The environments relevant for DVB-SH system are defined in clauses 4.2.1 to 4.2.3 and clause A.7 by propagation channel models. DVB-SH terminal categories are defined in clause 10. Furthermore, DVB-SH introduces two classes of receivers with different capability of processing time diversity elements in the transmitted waveform (see clauses 6.5, 7.3.3, and 10.2).

Concerning the mobility conditions, two main kinds of reception can be defined:

- Pedestrian speed reception where the channel conditions exhibit relatively low levels of fast fading (from movement of nearby objects such as trucks) and slow deep fades due to slow movements of the antenna and nearby blocking objects (≤ 3 kmph). The portable terminal with external or integrated antenna is used indoors or outdoors at a mean height of 1,5 m above ground.

- Mobile reception applies to the use of Car-mounted Terminals (Terminal Category 1) with speeds higher than 3 kmph. It is assumed that the receiving antenna is external and at a minimum height above ground level of 1,5 m. Terminal Category 2b and 3 used for direct reception within a car or train or any kind of vehicle could also be considered as a case of mobile reception. It can be called in vehicle reception.

11.2.2 DVB SH reception conditions

The five cases of reception conditions defined in DVB-H, called Class A, B1, B2, C and D (see EN 302 304 [3] and [19]), fit also the DVB SH terrestrial variety of contexts. An additional dimension can be introduced if we consider also the DVB-SH environments defined in clause 4: rural, urban, and suburban. These classifications can also be used for satellite or hybrid coverage, although not all reception cases are applicable in all environments due to the intrinsic power limitation of the satellite link. Conversely, some reception conditions in certain environments are only applicable to satellite coverage, due to the geographical limitation of the terrestrial coverage. An additional case can also be defined in DVB-SH; this case refers to the satellite indoor coverage achieved due to a domestic gap filler (defined as TR (b) in the system specifications) that receives and amplifies the satellite signal.

Note that this partition can apply equally to SH-A and SH-B system architectures.

In all scenarios, and in the rest of the document, receiver is placed at 1,5 m above ground.

Table 11.1 : Usage scenario

Reception Condition	Situation	Characteristics	Environment	Coverage	Channel Characteristics	Typical channel parameter Relevant DVB-SH parameters
Reception condition A	Outdoor pedestrian	Up to 3 kmph	Rural	Satellite	Stationary : Low delay/low spread	LOS :AWGN/Rice K>10 : Additional margin to cope with fading For shadowed, Rice below 7 dB Time interleaving to mitigate effects
					Low speed: large signal variation	LMS channel model at low speed. Time interleaving
			Urban	Terrestrial	Stationary : Rayleigh/Very low Doppler	TU6 channel? /low code rate improves. Antenna diversity also improves
					Low speed / Rayleigh /low Doppler	Higher margins to cope with slow fading effects
			Suburban	Terrestrial, Hybrid	For terrestrial same as above	For terrestrial same as above
					No hybrid channel model available	No hybrid channel model available
Reception condition B1	Light-indoor	Up to 3 kmph, lightly shielded building	Rural	See note (1)		
			Urban	Terrestrial	Channel is the same as Reception A with high penetration margins	TU6 channel? /low code rate improves. Antenna diversity also greatly improves
			Suburban	Terrestrial, Hybrid	Same as above for terrestrial No hybrid channel model available	
Reception condition B2	Deep-indoor	Up to 3 kmph, highly shielded building	Rural	See note (1)		
			Urban	Terrestrial	Channel is the same as Reception A with higher penetration margins as in B1	TU6 channel? /low code rate improves. Antenna diversity also greatly improves
			Suburban	Terrestrial	Same as above (lower margins)	Same as above
Reception condition C	Mobile (vehicle) with roof-top antenna	Up to 200 kmph	Rural	satellite	Large signal strength variation depending on environment	LMS channel model at medium/high speeds for different environments
			Urban	Terrestrial	Multiple Rayleigh fading paths Delay spread depends mainly on network characteristics	Channel models like TU6 cover this scenario at least for low or medium power repeaters. Critical SFN scenarios require channel models with higher delay spread.
			Suburban	Terrestrial, Hybrid	For terrestrial same as above No hybrid channel model available	For terrestrial same as above No hybrid channel model available
Reception condition D	Mobile (portable) in-car	Up to 130 kmph	Rural	See note (2)		
			Urban	Terrestrial	Multiple Rayleigh fading paths Delay spread depends mainly on network characteristics	Channel models like TU6 cover this scenario at least for low or medium power repeaters. Critical SFN scenarios require channel models with higher delay spread.
			Suburban	Terrestrial	Same as above	Same as above

NOTE: According to clause 10 classification, terminals used for the different reception conditions are the following: reception conditions A, B1, B2 and D are associated to terminal categories 2a, 2b and 3 (handset, handheld and portable):
(1) for Reception conditions B1 and B2 in the rural environment, under satellite coverage, it is assumed that the satellite signal is assisted by terrestrial repeater (TR (b)). The link budget applies to these TR(b), not to the end-user terminal;
reception condition C is associated to terminal category 1 (vehicular):
(2) for Reception condition D in the rural environment, under satellite coverage, it is assumed that the satellite signal is assisted by terrestrial repeater (TR (c)). The link budget applies to these TR(c), not to the end-user terminal.

11.3 Coverage definition

In general, coverage is defined by a reference area and a percentage of that area where the signal is received with a given required quality of service a given percentage of time. Consideration is also given to the edge of the signal area (whether satellite beam, or terrestrial transmit signal). The edge is taken as the contour beyond which the signal degrades (either from leaving the satellite beam footprint, or moving too far from the terrestrial transmitter) below the point required to deliver the given QoS level. DVB-SH coverage is made up of three different components: satellite only coverage, terrestrial only coverage, and hybrid coverage.

The coverage area definition is differing, first between satellite and terrestrial coverage, and second between the broadcaster and cellular network approaches.

11.3.1 Broadcaster approach

In defining the coverage area for each reception condition, a three level approach is taken (see EN 302 304 [3]).

Receiving location coverage: the smallest unit is a receiving location with dimensions of about 0,5 m. In the case of portable antenna reception, it is assumed that optimal receiving conditions will be found by moving the antenna or moving the handheld terminal within 0,5 m in any direction. Such a location is regarded as covered if the required carrier-to-noise and carrier-to-interference values are achieved for 99 % of the time.

Small area coverage: the second level is a "small area" (typically 100 m × 100 m).

Coverage area: the third level is the coverage area. The coverage area of a transmitter, or a group of transmitters, is made up of the sum of the individual small areas in which a given class of coverage is achieved.

11.3.2 Cellular approach

Cell coverage: rather than using the three level approach coverage area is defined as the aggregation of several circular (approximately) cells defined by their radius. The transmitter is in the centre of each cell and can use omnidirectional antenna or trisector or any other type of antennas.

Cell radius: when using cellular network methods, coverage are is defined as a cell of radius R, defined, as the extreme points where the link budgets is satisfied to meet the required level of signal.

Cell edge: is defined at distance R of the transmitter.

11.3.3 Satellite coverage

For satellite-only coverage: the basic covered area is the satellite antenna footprint, typically tailored to a satellite spot beam. However, not all points in the satellite coverage area are reached with the same quality. Depending on the service definition, service coverage will have to take into account environments such as shadowing, which can be caused by trees, tall buildings, or even low buildings in the case of low satellite elevation angle.

11.3.4 Quality of coverage

The coverage is characterized by a certain quality of service characterized by a quality criterion. For each propagation channel (Gaussian, Rice, TU6, etc.) and each modulation and coding state, this criterion is satisfied by a minimum C/N value, which is used in network planning.(called network planning value).

The cell coverage probability is defined as the percentage of locations inside the cell where the criterion is satisfied or where C/N is above the minimum corresponding value.

Under terrestrial coverage (TU6), clause A.12 utilizes two quality criteria:

- FER 5 %, equivalent to MFER 5 %, which is used in DVB-H: 5 % of frames in error; and
- ESR5; defined in clause A.8: 1 s maximum in error every 20 s, which is more demanding and a fulfilment criterion defined by ESR(5).

As shown in clause A.12, less than 1 dB in C/N separates the two criteria. In the computations associated to TU6, an ESR(5) fulfilment criterion is chosen.

For satellite channel, ESR(5) fulfillment criterion is chosen.

BMCO [19] defines the different qualities of service corresponding to different coverage probabilities, and depending on the different reception conditions. They are defined as "good" and "acceptable":

- the terrestrial coverage area is declared *"good"*:
  - at least 99 % of receiving locations within the area are covered for reception conditions C and D;
  - at least 95 % of receiving areas at the edge of the area are covered for reception conditions A and B;
- the terrestrial coverage area is declared *"acceptable"*:
  - at least 90 % of receiving locations within the area are covered for reception condition C and D;
  - at least 70 % of the receiving locations at the edge of the area are covered for reception conditions A and B.

Table 11.2: Coverage probability for terrestrial coverage (broadcast method)

	"Good"	"Acceptable"
Reception Condition A	95 % edge	70 % edge
Reception Condition B	95 % edge	70 % edge
Reception Condition C	99 % overall	90 % overall
Reception Condition D	99 % overall	90 % overall

In the cellular coverage, there is no specific qualification of coverage quality or required percentage of coverage (cell or edge of cell)With given percentage of edge or overall coverage. There are some coverage obligation under the regulation process.

Usually, coverage probability can be defined on the whole cell, or at the edge of cell coverage. There is no specific rule.

The most usual qualities of coverage are 90 % and 95 % overall coverage for all reception conditions. Hence, other coverage probability can be used upon request of operator.

Under satellite coverage: **under satellite coverage, there are no typical values of coverage probability.**

In the satellite clause, the obtained coverage probability are provided in the different examples.

11.4 Network planning factors

11.4.1 Introduction

The network planning factors constitute the ensemble of parameters that are needed in the different network planning computation and tools; As they are common to the different methods, they are described in this clause. One can distinguish three categories of criteria:

- channel dependant factor: relative to the shadowing and in building penetration margins for instance;
- technology dependant factor : bandwidth, noise factor, antenna gain, etc.; and

- network architecture factor or implementation optional parameter: SFN topology gain, antenna diversity gain.

11.4.2 Channel dependant factor

The first factor is the in building or in vehicle entry or penetration losses, applicable to reception conditions B and D.

Depending on the different terrestrial planning methods, the range of values is somehow different, and so the taken values are give in each corresponding clause.

The other factor is called location correction factor, or shadowing margins or fading margins in the satellite links.

Correction location factor is defined by the product: $C_l = \mu * \sigma$ in EN 302 304 [3] and [19] where: μ is called the distribution factor, and σ is the standard deviation of the distribution. of the signal variation.

The possible values are provided in clause 11.5.3.1.

The term shadowing margins is used currently in the cellular network approach. The most commonly used propagation model is Log normal attenuation with the standard deviation σ.

Considering the distance to transmitter d and R the cell radius, probability that signal exceeds a defined threshold at distance d in an area dA (elementary area) is given by:

$$P_e = \frac{1}{2}\left[1 - erf\left(\frac{-C_l + 10nLog(\frac{d}{R})}{\sqrt{2}\sigma}\right)\right]$$

where: C_l is the shadowing margin associated with the coverage probability and σ is the shadowing standard deviation.

So, when being at edge of cell, d = R and the equation becomes:

$$Pe = \frac{1}{2}\left[1 - erf(a)\right]$$

with $a = {}^{-C_l}\!\!\diagup_{\sigma\sqrt{2}}$

It must be noticed that the ratio C_l / σ is the location correction factor μ.

The relationship between overall coverage probability and shadowing margins is given by the Jakes formula, obtained by integrating the previous equation all over the cell.

$$P_c = \frac{1}{2}\left[1 - erf(a) + \exp\left(\frac{1-2ab}{b^2}\right) \times \left(1 - erf\left(\frac{1-ab}{b}\right)\right)\right]$$

where: P_c is overall cell coverage probability, a the same variable as above and

$$b = \frac{10n \times \log(e)}{\sigma \times \sqrt{2}}$$

n is the path loss exponent (usually 3,5 in cellular networks);

e the exponential constant.

This probability is calculated at each location with the appropriate values, taking into account standard deviation.

There are many ways to use these formulas, whether a cell edge coverage or a whole cell coverage is required.

As the requirements are given in terms of whole coverage, the margins will be computed accordingly, and the cell edge coverage indicated.

In the satellite link budgets, there are different ways to consider the link margins:

- for stationary reception, a Rice channel is assumed, leading to similar computation as above, providing availability versus link margins. Then link budgets give the possible link margins, and the corresponding availability;
- for mobile reception (above 3 kmph), the link margins are given by the link budgets, and simulations provide the availability depending on conditions.

11.4.3 Technology dependant factors

By this we mean the antenna gain, frequency band and the receiver noise factor.

The different possible frequency bands have been explored in clause 4.

In this context, the computations are done in S band (2 170 MHz to 2 200 MHz) all over the document, but L band is also a candidate band, and the results can be easily adapted.

The noise bandwidth can have all the values described in the DVB-SH standard. For the computed examples provided in this clause, a noise bandwidth of 4,76 MHz is assumed, corresponding to the OFDM 5 MHz case.

The terminal related factors are the Noise Factor, the antenna gain,, the antenna polarization, and the antenna temperature, resulting in the factor of merit G/T, used in the satellite link budgets.

The different terminal categories and their related characteristics are described in table 10.2.

11.4.4 Network architecture or implementation optional factors

This clause concerns some factors depending on some specific system architecture, mainly SFN gain and on implementation antenna diversity gain, that may additionally be taken in account by the Network planning tool.

a) The SFN configuration

The SFN topology can be considered in two cases:

- terrestrial only coverage;
- hybrid satellite/terrestrial coverage.

In the terrestrial only coverage, it is an important feature of the DVB-SH. The SFN gain will depend on the network topology. When overlap is made possible by construction and the cell transmitters are properly synchronized, there is SFN gain with respect to a single-cell case. Similarly, the UMTS networks are also built as Single Frequency Network, probably more known under the name of networks with frequency reuse factor "1".

In UMTS networks, the handover procedure occurs when a mobile is at the edge of two or three cells, where signal received from different nodes B are of similar strength; Thanks to the Rake receiver, terminal can acquire and combine the different signals coming from the Nodes B, increasing signal strength by the amount of the so called Soft Handover Gain (SHG). This point is studied in the literature. Different references provide analytical and experimental studies of Handover gain, leading to 3 dB to 4,6 dB Soft Handover gain.

In DVB-SH/H SFN networks at the cell boundaries, the signals can be added coherently, thanks to the Guard Interval, leading to SFN gain.

This gain can be translated into the following advantages:

- increase of the coverage probability when dealing with equal transmit power as for a single cell;
- decrease of Transmit power while keeping the same coverage and cell radius (equivalent to shadowing margins reduction);
- increase of cell coverage.

The resulting so called SFN gains are in a range between 3 dB and 5 dB.

The network planner will take into consideration the requirements related to the SFN configuration. Namely, the deployment of the transmitters must be such that the signal coming from different stations can be added coherently (i.e. relative delay falls within the guard interval). When this is not respected the network will endure an additional level of self interference that may degrade the overall performance.

The hybrid SFN topology and possible associated gain is not studied in this release of the Implementation Guidelines.

b) Receiver antenna diversity

Antenna diversity has been studied and implemented for a long time in the telecommunication industry, and also by the car equipment manufacturers, as it can provide improvement of reception quality. The antenna diversity is not a standard feature. Implementation depends on wavelength and terminal size. Typically, the use of S band allows implementing dual antenna diversity on handsets, while it will be more adapted for PDA in L band for instance. According to various studies, a gain of 3 dB to 6 dB in outdoor conditions and 6 dB in indoor reception (reception condition B) can be expected using Maximum Ratio Combining technique.

As it is not a standardized feature, though mentioned in clause 10, it will not be taken into account in the different computations.

11.5 Terrestrial Network Planning based on minimum equivalent field strength (broadcast approach)

11.5.1 Introduction

Traditional TV broadcast is based on roof service characteristics, typically 10 m height, from high towers, high powered sites, to high gain Yagi type receiving home antennas. Quality of service provided is measured through the computation of the minimum median equivalent field strength.

Broadcast Network planning tools are usually proprietary to each broadcaster, that allow them to determine the transmitters optimal topology, and accordingly defined delivered quality of coverage.

For DVB-H, these methods have been extensively discussed at the BMCO forum and have been enhanced to take into account the specifics of indoor service planning to individual terminals, such as in building penetration losses, shadowing margins (hereunder defined as location correction factor), thus introducing near ground coverage (1,5 m) in urban and rural areas, and also technology terms such as emission antenna gain, and reception antenna diversity gain.

Only few DVB-H networks are actually deployed on large scale, thus providing little experience feedback; likewise actual cellular topologies and network planning based on these tools have limited background for broadcast services, and mostly rely on experiences from mobile radio services planning (detailed in the cellular approach clause).

The following document gives an illustration of applicable methods and typical parameterizations as recommended by the BMCO forum. Mobile services, derived methodologies are presented in the cellular approach In the following, different calculations are made for the different reception conditions identified earlier and are given for S band only, but they can be extended easily to L band or to any other usable frequency band.

11.5.2 Quality of coverage

The quality of coverage is provided in clause 11.2.4.

11.5.3 Network planning based on field strength computation methodology

To compute the median filed strength, the following reception cases are considered for terrestrial reception:

- reception condition A (outdoor portable);
- reception condition B1 and B2 (indoor portable);
- reception condition C Mobile (vehicle with roof top antenna);
- reception condition D (portable, in vehicle with no roof top antenna).

The C/N values will be the same as used in the previous clause on minimum signal level. To calculate the minimum median power flux density or equivalent field strength needed to ensure that the minimum values of signal level can be achieved at the required percentage of locations, the following formulas are used:

$$\Phi_{min} = P_{s\,min} - A_a$$

$$A_a = G + 10 \log 10 \ (\lambda^2/4\pi)$$

$$E = \sqrt{4\pi\eta \frac{P_{s\min}}{G_a}} \cdot \frac{f}{c}$$

or, in dB:

$$E[dB\mu V/m] = P_{s\min}[dB(mW)] - G_a[dB] + 77{,}2 + 20\log_{10} f[MHz]$$

This equation gives the link between the two possible methods,

where: $\eta = 120\pi\ \Omega$;

$E_{min} = \Phi_{min} + 120 + 10 \log (120\pi) = \Phi_{min} + 145{,}8$;

$\Phi_{med} = \Phi_{min} + P_{mmn} + C_l + L_o$ (Reception condition A or C);

$\Phi_{med} = \Phi_{min} + P_{mmn} + C_l + L_b + L_o$ (Reception condition B);

$\Phi_{med} = \Phi_{min} + P_{mmn} + C_l + L_v + L_o$ (Reception condition D).

In case of antenna diversity use, all value of minimum power flux density must be written differently:

$$\Phi_{med/div} = \Phi_{med} - G_{div}$$

$$E_{med} = \Phi_{med} + 120 + 10 \log (120\pi) = \Phi_{med} + 145{,}8$$

An equivalent formula is given here below:

$$E_{med} = NF + 10*Log(k_B T\ B) + C/N - G + 107{,}2 + 20*Log(f) + L_b \text{ (or } L_v) + \mu*\sigma + L_o \text{ (other losses)}$$

where:

C/N : RF signal to noise ratio required by the system (dB) for the required performance and modulation/coding scheme;

Φ_{min} : minimum power flux density at receiving location (dBW/m2);

E_{min} : equivalent minimum field strength at receiving location (dB⌈V/m);

A_a : effective Antenna Aperture;

G : antenna gain (dBi);

L_b : building penetration loss (dB);

L_v : vehicle entry loss (dB);

L_o : all other losses (polarization mismatch,.etc.);

P_{smin} : minimum receiver signal input power (dBW);

P_{mmn} : allowance for man made noise (dB);

C_l : location correction factor (dB);

Φ_{med} : minimum median power flux density, planning value (dBW/m2);

E_{med} : minimum median equivalent field strength, planning value (dB⌈V/m).

For calculating the location correction factor C_l a log-normal distribution of the received signal is assumed. The location correction factor can be calculated by the formula:

$$C_l = \mu * \sigma$$

where: μ is the distribution factor, and σ is the standard deviation of the distribution.

It must be noticed that this value of μ depends on the model assumed for the distribution (normally lognormal is considered for shadowing), as well as the standard deviation. The product is in fact the taken shadowing margins. Possible values are given in further clauses.

11.5.4 Network planning factors

11.5.4.1 Location Correction

11.5.4.1.1 Location correction for reception condition A

In reception condition A (portable outdoor), the location variation factor is given by:

$$C_l = \mu * \sigma$$

The common used value for σ is 5,5 dB (see ITU-Recommendation M.1225 [11]).

This gives the table for the different qualities of coverage.

Table 11.3: Location correction factor for reception condition A

Edge of area coverage (%)	Distribution factor (μ)	Location correction factor (dB)
70	0,52	3
95	1,64	9

11.5.4.1.2 Building penetration losses and location correction for reception conditions B1 and B2

In reception conditions B, signal are subject to in building penetration losses. The location correction factor at indoor locations is the combined result of the outdoor variation and the variation factor due to building attenuation. These distributions are expected to be uncorrelated. The standard deviation of the indoor field strength distribution can therefore be calculated by taking the root of the sum of the squares of the individual standard deviations. As a consequence, the location variation of the field strength is increased for indoor reception.

$$\sigma = (\sigma_o^2 + \sigma_p^2)^{1/2}$$

Table 11.4 gives the building penetration losses values proposed.

Table 11.4: Building penetration losses

Band	S band
Penetration loss	Loss
Condition B1	14
Condition B2	18

As an example, table 11.5 gives the location correction factor in deep indoor.(B2 reception) with overall:

$$\sigma = 10 \text{ dB}$$

Table 11.5: Location correction factor for reception condition B2 in S band

Edge of area coverage (%)	Distribution factor (μ)	Location correction factor (dB)
70	0,52	5,2
95	1,64	16,4

11.5.4.1.3 Location correction for reception conditions C and D

Mobile reception is defined as a reception with moving receiver or at location where large objects moving around the receiver.

Two cases are possible:

- mobile reception with handset terminal inside a vehicle, car or inside any moving object (Reception condition D);
- mobile reception with a vehicular terminal (Reception condition C).

Clause 7 includes the values that should be used for planning in mobile reception. It should be taken into account the great influence of Turbo code FEC on the C/N and maximum Doppler shift (and therefore maximum speed in a particular channel).Any way, link budgets are performed with speed below the limit.

(a) Location percentage requirements for mobile reception (conditions C and D)

For reception categories C and D, some additional margins are added to cope with mobile environments (see EN 302 304 [3]).

Table 11.6: Location correction factor for reception conditions C and D

Overall area coverage (%)	Distribution factor (μ)	Location correction factor (dB)
90	1,28	7
99	2,33	12,8

(b) Vehicle entry loss

For mobile reception inside cars or any other vehicle entry loss must be taken into account. For instance, an entry loss is 7 dB is taken for in car penetration losses. The location correction factors will be the same for in vehicle reception.

11.5.4.2 Network architecture and implementation optional factors

(a) The SFN possibility

The SFN gains appear at the cell overlap points. The field strength computation is made all over the area and not at specific points. So SFN gain will not be considered in this clause.

(b) Receiver antenna diversity

No diversity is used in the computation.

11.5.5 Field strength computation examples

The objective of the link budget is to provide the minimum field strength values at 1,5 m height. Normally, the exercise can be made for all possible values of C/N. We limit the examples to a few values corresponding to some situations in the clause A.12. Extrapolation from the proposed values is straightforward.

An example of each reception condition is provided in S band, only to limit the number of tables.

Antenna diversity is used for reception conditions A, B and D, though it can also be implemented for vehicle roof top reception.

Two modulation and coding schemes are studied:

- QPSK 1/3; and
- 16QAM 1/3.

The required C/N values are extracted from the clause 12. In the following clause, we give an example of table, and a synthesis for different examples. The C/N values are for 99 % of ESR(5) fulfilment, which corresponds to FER = 1 %, of course more constraining (between 0,5 dB and 1 dB) than the usual FER 5 % (or MFER 5 %).

In all the tables, the C/N values include implementation losses as follows:

- 1,6 dB for QPSK;
- 2 dB for 16 QAM.

It must be noticed that maximum speed in the tables (conditions C and D) is 50 kmph, below the CN + 3 dB Doppler limit.

11.5.5.1 Field strength computation for reception condition B2

The studied case is with Handheld Category 3 in reception condition B2(3 kmph) and the link budgets are computed for 99 % of ESR(5) fulfilment at different coverage qualities as indicated in the table 11.7.

Table 11.7: Median field strength in reception condition B2 with QPSK 1/3 and 16 QAM 1/3

Modulation and coding		**QPSK 1/3**	**16QAM 1/3**
Frequency band	(MHz)	2 182,5	
Equivalent Noise Bandwidth	B (MHz)	4,76	
Receiver Noise Figure	NF (dB)	4,5	
Minimum C/N required by system	(dB)	5,1	10,5
Min receiver signal input power	Psmin (dBW)	-127,60	-122,20
Min. equivalent receiver input voltage, 50 Ω	Us min (dBµV)	9,4	14,8
Antenna gain relative in dBi	G (dBi)	-3	
Effective Antenna aperture	Aa (dBm²)	-31,23	
Min. power flux density at receiving location	Φ min (dBW/m²)	-96	-91
Min. equivalent field strength at receiving location	Emin (dBµV/m)	49	55
In building or in car penetration losses	Lb (dB)	18	
Location correction factor for 70 % edge of coverage	Cl (dB)	5,2	
Minimum median power flux density at 1,5 m	Φmed (dBW/m²)	-73	68
Minimum median equivalent field strength at 1,50 m	Emed (dBµV/m)	**73**	**78**
Location correction factor for 95 % edge of coverage	Cl (dB)	16,4	
Minimum median power flux density at 1,5 m	Φmed (dBW/m²)	-62	57
Minimum median equivalent field strength at 1,5 m	Emed (dBµV/m)	**84**	**89**

11.5.5.2 Synthesis

Table 11.8 gives a synthesis of the previous link budgets based on minimum filed strength.

Table 11.8: Overall synthesis (Field Strength in dBµV/m)

Reception conditions/terminal	Quality of Service cases	QPSK 1/3	16 QAM 1/3
Condition A/ Handheld Cat 3	70 % edge	52	58
	95 % edge	58	64
Condition B2/ Handheld Cat 3	70 % edge	73	78
	95 % edge	84	89
Condition C/Vehicle Cat 1	90 % overall	48	53
	99 % overall	54	59
Condition D/ Handheld Cat 3	90 % overall	61	67
	99 % overall	67	73

11.5.6 Use of field strength based Radio Network Planning Tool example for DVB-SH Network Design

For further study.

11.6 Terrestrial Network Planning based on minimum received signal level (cellular approach)

Network Planning tools allow to compute precise site distribution and cell ranges. They use propagation models to predict with high accuracy path loss, and different cell ranges. These models integrate field based calibration. Some Practical Network Planning tools, include a propagation model called Standard Propagation Model, based on Cost Hata model with different correction terms allowing range of application down to 200 m from the BTS.

Others suppliers provide tools based on ITU-R Recommendation P.526 [27] model.

The complete description, parameterization of the tool, and procedures of use is not in the range of the present document.

A network planning tool will compute the coverage using an intermediate parameter called "the minimum signal level for coverage". This parameter corresponds to the minimum signal power level that will ensure service reception. It takes into account propagation characteristics associated to the considered type of scattering environment. The network planning tool will plot coverage maps.

This clause gives an overview of the common practices of the cellular industry (cellular operators), and provides a first level approximation of DVB-SH networking, including the computation of cell range, given the transmitter EIRP.

After some considerations on network topology, we provide the basic equations used in the network planning tools.

11.6.1 Cellular network topology

Without going into complex topological details, a cellular network is composed of an aggregation of cells that can be considered in first approximation as circle of radius R.. This radius defines the edge of cell and is considered in the link budget computation..

Figure 11.1 shows a regular cellular network with trisectorized cells and in SFN possible topology.

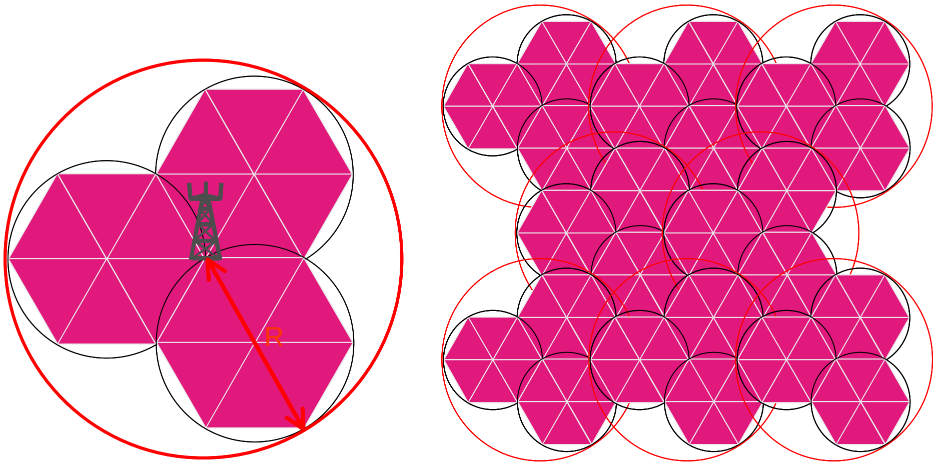


Figure 11.1: Trisectorized cell topology

For the trisectorized antenna on the left of the figure, the coverage area of the three sectors is approximated by a regular hexagon. This approximation makes it easy to show the overall coverage area in the figure on the right, as the hexagons completely cover the plane area without overlap.

The surface of a cell is the following S= 1,95 R^2.

Given the intersite distance D, relationship between D and R is the following: D = 3R/2.

This formula can be used to provide estimates of the different of required sites to cover an area.

11.6.2 Quality of coverage

Given a certain quality criterion, different whole cell coverage probability. can be considered. The 90%coverage probability means that 90 % (for instance) of the cell can receive at least the minimum signal level required for insuring the target C/N value.

Usually 90 % and 95 % overall cell coverage are considered. This corresponds respectively to 77 % and 88 % edge of cell coverage. But other values can be as well required.

Usually, network dimensioning is made for indoor service with handheld at 90 % or 95 % cell coverage., Such indoor coverage probabilities imply de facto in very high outdoor coverage probability, above 99 %.

11.6.3 Network planning tool

A typical numerical network planning tool uses the plotting of the Minimum Received Signal Level.

Starting by the following equation:

$$P_n = NF + 10 * \log (k * T_0 * B);$$

$$P_{s\ min} = P_n + C/N.$$

where: NF is the Noise Factor of the receiver;

K: Boltzmann constant;

T_0: receiver temperature;

B: noise bandwidth of the signal;

C/N is the required C/N for a given quality of service, depending on the modulation, coding and propagation channel.

So the minimum received signal input power at antenna input can be written:

$$P_{amin} = P_{s\,min} – G + L_p = C/N – G + L_p + NF + 10^* \log (k * T_0 * B).$$

It must be noticed that power unit is dBm,

Where: G is the antenna gain in dBi and L_p is possible polarization mismatch losses.

On the other hand the basic link budget equation can be written as follows, considering the received C/N:

$$C/N = EIRP_{Tx} – PL – Ms – L_b – L_v + G – NF – 10^* \log (k * T_0 * B) + G_{SFN} \qquad \text{(equation 1).}$$

where: $EIRP_{Tx}$: transmitter EIRP;

PL : path loss;

Ms : shadowing margins, following a Log Normal law of standard deviation σ;

L_b : building penetration loss (dB) (if any);

L_v : vehicle entry loss (dB) (if any);

L_p : polarization losses (if any).

$$EIRP_{Tx} = Tx_{power} – (losses) + Tx\ (antenna\ gain)$$

where: transmitted power is in dBm, losses in dB, and antenna gain in dBi.

Path loss is computed using different path loss models that are included in the network planning tool. There is a variety of PL models (Cost 231 Hata, Welfish Ikegami, Xia-Bertoni) that are usually on field calibrated. For the purpose of global link budgets for rough order of magnitude estimates, the Cost Hata model is used in the link budgets with some calibrated factors.

After some elementary algebra, equation (1) can be written:

$$C/N – G + L_p + NF + 10^* \log (k * T_0 * B) + Ms + L_b + L_v – (G_{SFN}) = EIRP_{Tx} –PL \qquad \text{(equation 2).}$$

The left part of the equation is the Minimum Received (Rx) Signal level, independent of transmitted power.

Other definitions can be introduced:

- System Gain (SG): SG = $EIRP_{Tx} – P_{amin}$;
- MAPL (Maximum Allowable Path Loss) = SG – Ms - L_b – L_v + G_{SFN}.

Leading to equivalent definition of the Minimum Received (Rx) Signal level.

11.6.4 Network planning factors

11.6.4.1 Shadowing margins

In cellular network radio planning it is usual to distinguish between different environments, while this classification does not exist in EN 302 304 [3].

- Dense Urban (DU): dense cities like Paris;
- Urban (U): cities like immediate Paris suburbs called also "Dense Individual, Mean collective";
- SubUrban (SU): residential areas, called also "Mean individual, mean collective" including as well industrial zones;
- Rural (RU): open areas : small towns and villages, highways, fields.

The last case is not applicable for DVB-SH, as rural areas will be under satellite coverage.

Shadowing margins follow a Lognormal law of standard deviation usually in the range of 6 dB to 10 dB, or even more for some authors [i.11], [i.10]. For network planning exercise, a common value of 8 dB is assumed for urban and suburban areas in [i.11], [i.12], [i.13], [i.14], and [i.15].

Concerning in building penetration losses, there is also a wide range of values. For instance, deep indoor penetration losses are in the range of 18 dB to 21 dB, and sometimes more.

In the following clauses, we give the possible ranges of values, and give an example for network planning purpose.

11.6.4.1.1 Shadowing margins for reception condition A

As proposed value for standard deviation is 8 dB, table 11.9 gives the shadowing margins for various conditions and different cell coverage.

Usual coverage probability values whole cell are 90 % and 95 %.

Table 11.9: Shadowing margins for reception condition A DU/U/SU

Cell Coverage (%)	Shadowing margins (dB)
90	5,5
95	8,7

11.6.4.1.2 In building penetration losses and shadowing margins for reception condition B

In building penetration margins values depend on the different materials and the environments. Table 11.10 presents the possible range of values, and the sued values in the network planning. But the taken values can vary from an operator to another.

Table 11.10: Indoor penetration margins in S band for reception condition B

Environment/B1 or B2	DU/B2	U/B2	SU/B1
Indoor penetration margins range (dB)	18 to 21	15 to 18	13 to 15
proposed value for network planning	18	15	13
NOTE: For the shadowing margins, the same formula is applied when dealing with the combination of indoor and outdoor signal variations.			

The following standard deviation values are taken:

- outdoor standard deviation: 8 dB for DU, U and SU;
- indoor standard deviation: 6 dB for all cases.

The shadowing margins are provided in table 11.11.

Table 11.11: Shadowing margins in reception conditions B1 and B2

Environment	DU	U	SU
Standard deviation (dB)	10	10	10
Shadowing margins @ 90 % cell coverage	7,6	7,6	7,6
Shadowing margins @ 95 % cell coverage	11,6	11,6	11,6

11.6.4.1.3 Shadowing margins for reception conditions C and D

In reception condition C, the same shadowing margins will be used as for reception condition A, as we are in outdoor reception.

In reception conditions D, an in vehicle penetration loss of 7 dB is required, and the shadowing margins are the same as for condition C.

11.6.4.2 Network architecture optional parameters

This clause concerns some criteria depending on some specific system architecture, mainly SFN gain and antenna diversity gain.

(a) The SFN configuration

The used topology is similar to the UMTS one. As the link budgets are computed at cell edge, SFN gain can be taken into account and 4,7 dB gain is used in the link budgets.

(b) Receiver antenna diversity

The antenna diversity gain will not be taken into account as already stated in clause 11.4.4.

11.6.5 Link budgets examples for DVB-SH network planning based on minimum received signal level

For cellular network, global estimation uses link budgets for budgetary purposes. As it gives only a rough order of magnitude of the coverage, Cost Hata model is also commonly used below 1 km.

In the link budgets, the power of each transmitter is provided, with different losses (cable losses, diplexer losses if any: between 3 dB and 4 dB overall), the antenna gain (18 dBi per sector in the following examples), resulting in the transmitted EIRP.

Diplexer losses are considered when the DVB-SH transmitter is sharing equipments with other BTS (3G for instance. When stand alone repeater, non diplexer loss is considered.

The results of the link budgets are the Minimum Signal Level for Network Planning (green line), and also the cell radius, and cell surface, given the reception condition, the terminal properties, and the required C/N.

Two modulation and coding schemes are studied:

- QPSK 1/3; and
- 16QAM 1/3.

The required C/N values are extracted from the Annex 12.In the following clause, we give an example of table, and a synthesis for different examples. The C/N values are for 99 % of ESR(5) fulfilment, which corresponds to FER = 1 %, of course more constraining (between 0,5 dB and 1 dB) than the usual FER 5 % (or MFER 5 %).

In all the tables, the C/N values include implementation losses as follows:

- 1,6 dB for QPSK;
- 2 dB for 16 QAM.

It must be noticed that maximum speed in the tables (conditions C and D) is 50 kmph, below the CN + 3 dB Doppler limit.

The following cases are provided:

- reception condition A: Outdoor pedestrian in dense urban area;
- reception condition B (B1 and B2): indoor pedestrian in dense urban area;
- reception condition C: vehicle at 50 kmph in suburban area;
- reception condition D: in car user at 50 kmph in suburban area.

As in the clause 11.5, we give an examples for reception condition B, and the synthesis for the other cases.

11.6.5.1 Link budgets for Reception condition B2

The studied case is with Handheld Category 3 and the link budgets are computed for 99 % of ESR(5) fulfilment over 95 % overall coverage in dense urban.

Table 11.12: Link budget for reception condition B2 @ 95 % with Tx power of 12 W and QPSK 1/3

SH-A Terrestrial link budget		
Radio interface parameters	**Unit**	**Dense Urban value**
Channel bandwidth	MHz	5,00
Frequency	MHz	2 182,50
Mode		2 048,00
Radio interface mod code		QPSK1/3
Antenna type		TRI
1/Guard Interval		8
Guard interval		0,125
Total number of subcarriers		1 705,00
Number of data subcarriers		1 512,00
Tu duration	µs	358,40
GI duration	µs	44,80
Ts duration	µs	403,20
Subcarrier spacing	kHz	2,79
Useful bandwidth occupancy	MHz	4,76
Useful data rate at MPEG2-TS interface	Mbit/s	2,490
Transmitting end		**Tx**
Power amplifier per carrier and sector	W	12,0
Tx Power at antenna input	dBm	40,8
Cable loss	dB	3,0
Diplexer loss	dB	0,7
Tx antenna gain	dBi	18,0
EIRP	*dBm*	*55,1*
ERP	*W*	*323,0*
Receiving end		**Rx**
Rx antenna gain	dBi	-3,0
Polarization mismatch	dB	0,0
Noise figure	dB	4,5
Antenna temperature	K	290,0
Ambient temperature	K	290,0
kT	dBm/Hz	-174,0
Equivalent Rx band	dBm/Hz	66,8
Rx noise floor	dBm	-102,7
Required C/N	dB	5,1
Rx sensitivity	dBm	-97,6
Minimum Rx level at antenna	*dBm*	*-94,6*
Minimum Signal Level for Network Planning	**dBm**	**-69,71**
System Gain	**dB**	**149,7**
Margins		
Average building penetration loss	dB	18,0
Target level of signal penetration		B2
Std shadowing outdoor	dB	8,00
Std shadowing indoor	dB	6,00
Std dev of Fading Margin	dB	10,00
Propagation constant		3,5
Shadow fading Margin - whole cell	dB	11,6
Coverage Probability - whole Cell		**95,0 %**
SFN network gain	dB	4,7
Rx gain (antenna diversity)	dB	0,0

SH-A Terrestrial link budget		
MAPL	**dB**	**124,8**
Cell range computation		
Cost 231-HATA model		
Cell range	**km**	**0,511**
Surface	km^2	**0,51**
H-BTS	m	30,0
H-MS	m	1,5
K1	dB	138,08
K2	dB	35,22
Kc	dB	-3,0
a(CPE)	dB	0,00
NOTE: Cell radius is 511 m in dense urban environment.		

11.6.5.2 Overall synthesis

As in the minimum field strength approach, the table 11.13 provides a synthesis concerning the Dense Urban case.

Table 11.13: Overall synthesis @ 95 % coverage

Reception Condition	A in dense urban		B2 in dense urban		C in suburban		D in suburban	
Modulation /coding	QPSK 1/3	16 QAM 1/3	QPSK 1/3	16 QAM 1/3	QPSK 1/3	16 QAM 1/3	QPSK 1/3	16 QAM 1/3
Power per sector (W)	7	27	12	32	7	25	7	25
Required C/N (dB)	5,1	10,5	5,1	10,5	3,1	8,5	3,1	8,5
Minimum Signal Level for Network planning (dBm)	-91	-85	-70	-64	-99	-94	-86	-80
Cell range (km)	1,718	1,771	0,511	0,474	4,748	4,788	1,964	1,981

It must be noticed, that its is easy to extrapolate the Minimum Signal Level to other values of C/N.

11.6.6 Use of radio network planning tool example for DVB-SH network design

The Radio Network Planning Tool allows to compute and show coverage maps based on the minimum received power level required to ensure service reception (depends on scattering environments).

The design is based on the use of a Radio Network Planning Tool (RNPT) and a geographical database.

The inputs of the network design are:

- the DVB-SH link budget for a precise configuration of the RNPT;
- the clutter and terrain model databases;
- the minimum signal levels for coverage (design level); and
- eventually the characteristics of existing 2G/3G sites if available for an eventual reuse of the antennas and feeders. (This is not a requirement: it can exist stand alone repeaters).

The outputs of the network design are:

- coverage maps with received power level or C/N+I level;
- number of sites;
- coverage statistics.

The goal of the network design process with the RNPT is to reach the minimum received power level defined in the link budget (i.e. coverage) for all scattering environments over the area considered in the design). A RNPT allows the designer performing the coverage of one area by positioning transmitting sites within the database.

The Minimum Rx signal level to be considered in the RNPT can be computed as follows:

Minimum Rx signal level = Tx EIRP – Maximum Allowable Path Loss (MAPL)

Using the Link budget formula provided above, it can be further computed from as follows:

Minimum Rx signal level = Min Rx input power + (Shadowing Margin + Building Penetration loss [see note 1]) - (SFN gain [see note 2] + Rx diversity gain)

NOTE 1: If Indoor coverage is targeted. Depending on the situation it can be also in vehicle penetration losses.

NOTE 2: SFN impact can also supported by the tool.

For a highly accurate network design especially in dense urbanized areas, ray-tracing simulations can be considered for a better reliability of the coverage of the network. For these simulations a deterministic propagation model (ray optical propagation model) is used, requiring a 3D building high-resolution database. This combination of ray-tracing and 3D vector database allows a higher accuracy and reliability of the coverage maps.

In some tools, SFN gain is added "manually", other advanced tools integrates already the SFN gain.

11.7 Satellite Network Planning

For the satellite network planning, the following reception categories are considered:

- Reception condition A (outdoor portable) for rural environment;
- Reception condition B1 with domestic gap filler;
- Reception condition C Mobile (roof top antenna).

For satellite network planning purposes, the following satellite to terminal propagation environments will be considered:

1) suburban or rural areas (villages);
2) open area or also rural (open field);
3) rural intermediate tree shadowing (road crossing a forest).

We have excluded from this analysis the urban and rural heavy tree shadowing environments as the satellite is not expected to deliver the DVB-SH service in these kind of environments.

11.7.1 Methodology for Satellite Coverage Calculation

The DVB-SH satellite coverage is a composite of the three environments mentioned earlier: suburban, open area and intermediate tree shadowing. Each environment has distinctive propagation characteristics and therefore DVB-SH physical and link layer performances are expected to be different. Moreover the required quality criteria in terms of ESR(5) is also dependent on the propagation conditions. For terrestrial channels that can be characterized as Rayleigh channels the requirement is 99 % of ESR(5) fulfillment; however, the satellite channel presents very different dynamics and therefore this requirement cannot be directly translated. At the present moment, we lack sufficient information to conclude on the required ESR(5) fulfillment necessary to guarantee a good video quality via satellite. The calculations presented in this clause are based on the assumption that a 90 % ESR(5) fulfillment is required, however the methodology is considered of general validity.

To get satellite coverage figure over a wide area it is necessary to:

- classify the satellite coverage region according to three environment categories listed above and associate a probability of occurrence $p_E(i) \le 1,\ i = 1,2,3$ for each of them over the satellite coverage;
- select typical mobile terminal speed for each environment;
- select an ESR(5) target fulfillment rate per environment $\left[F_{ESR5}(i)\right]_{\min}$, $i = 1,2,3$.

As explained earlier, this target fulfillment rate depends on the environment considered. For the calculations on the present document we will consider that for the satellite environments 90 % of ESR(5) fulfillment is required:

- define the system parameters required to compute the link budgets providing LOS *C/N* and *C/I* values as described in the following clause;
- define the DVB-SH waveform parameters required to perform the simulations as described in clause A.3;
- analyze the ESR(5) fulfillment rate $F_{ESR5}(i),\ i = 1,2,3$ achieved in each of the environments given the selected system and waveform parameters (see clause A.12 for typical results);
- compute the satisfaction index $\chi(i),\ i = 1,2,3$ for each satellite coverage environment defined as:

$$\chi(i) = \begin{cases} 1 \text{ if } F_{ESR5}(i) \ge \left[F_{ESR5}(i)\right]_{\min} \\ 0 \text{ if } F_{ESR5}(i) < \left[F_{ESR5}(i)\right]_{\min} \end{cases} \quad i = 1,2,3$$

- compute the overall satellite coverage C_{SAT} (in % as):

$$C_{SAT}(\%) = 100 \sum_{i=1}^{3} p_E(i)\ \chi(i);$$

where $p_E(i)$ is the probability of occurrence of each environment.

Note that the above described methodology assumes that the satellite elevation angle is not changing in an appreciable way over the coverage region. In case of large satellite beams, as it is the case of a global beam covering a large part of the continent, the above described procedure must be extended by splitting the satellite coverage over regions of similar elevation angle.

11.7.2 Basic formulas for Satellite link budgets calculation

As was done for the terrestrial case, the basic formulas for the calculation of the Signal to Noise ratio C/N in the satellite link are recalled hereafter. Satellite link budget is made of two components: uplink (from Gateway to Satellite) and a downlink (from Satellite to User Terminal). The formula for the uplink C/N calculation can be expressed as follows:

$$\left[\frac{C}{N}\right]_U = \frac{1}{kB} EIRP_{GW} \frac{1}{L_U} \left[\frac{G}{T}\right]_{SAT}$$

where:

k: Boltzmann's constant;

B: Noise Bandwidth;

$EIRP_{GW}$: Gateway Equivalent Isotropic Radiated Power;

L_U: Uplink losses;

$\left[\frac{G}{T}\right]_{SAT}$: Satellite Antenna Gain over the Satellite Noise Temperature;

$$EIRP_{GW} = \frac{P_{GWTX} G_{GW}}{L_T L_{FTX}}$$

where: P_{GWTX} is the transmitted power of the Gateway;

G_{GW} is the gateway antenna gain;

L_T is the de-pointing losses; and

L_{FTX} is the feeder loss between the transmitter and the antenna;

$$L_U = L_{FS} L_A$$

where L_{FS} is the free space losses (see formula below) and L_A is the atmospheric attenuation losses;

$L_{FS} = \left(\frac{4\pi R}{\lambda}\right)^2$ with R defined as the distance between the gateway and the satellite.

In a similar manner, for the downlink the following formula applies:

$$\left[\frac{C}{N}\right]_D = \frac{1}{kB} EIRP_{SAT} \frac{1}{L_D} \left[\frac{G}{T}\right]_{ES}$$

where $EIRP_{SAT}$: satellite Equivalent Isotropic Radiated Power;

L_D: Downlink losses;

$\left[\frac{G}{T}\right]_{ES}$: Terminal Antenna Gain over the Terminal Noise Temperature;

$EIRP_{SAT} = \frac{P_{SAT} G_{SAT}}{L_T L_{FTX}}$;

where P_{SAT} is the transmitted power of the Satellite;

G_{SAT} is the satellite antenna gain;

L_T is the de-pointing losses; and

L_{FTX} is the feeder loss between the transmitter and the antenna.

$L_D = L_{FS} L_A$;

where L_{FS} is the free space losses (same formula as above considering *R* as the distance between the satellite and the terminal); and

L_A is the atmospheric attenuation losses in the downlink.

Additional to the uplink and downlink C/N components, the link budget needs to take into account the interference contribution. In general, interference is modelled as an additional source of noise:

$$\left[\frac{C}{N+I}\right]^{-1} = \left[\frac{C}{N}\right]^{-1} + \left[\frac{C}{I}\right]^{-1}$$

where I is the interference component caused by multiple sources of interference: intermodulation (IM), co-channel interference (CC), and adjacent channel interference (ACI):

$$I = I_{IM} + I_{CC} + I_{ACI}\,.$$

In the following we will assume a transparent GEO satellite system (no modulation or coding on-board) which is limited by the downlink component (from satellite to user mobile terminal). This last assumption is consistent with the DVB-SH scenario where small hand-held or mobile terminals are targeted.

The basic link budget for transparent satellite systems is computed as follows:

$$\left[\frac{C}{N+I}\right]^{-1} = \left[\frac{C}{N+I}\right]_{uplink}^{-1} + \left[\frac{C}{N+I}\right]_{downlink}^{-1}.$$

When the downlink part is driving the link budget, the overall link budget can be approximated by the downlink component:

$$\left[\frac{C}{N+I}\right] \approx \left[\frac{C}{N+I}\right]_{downlink}.$$

11.7.2.1 Satellite link margin

In order to close the link, the available Signal to Noise plus Interference ratio needs to be greater than the required Signal to Noise ratio: for given waveform configuration (MODCOD):

$$\left[\frac{C}{N+I}\right] \geq \left[\frac{C}{N}\right]_{\text{REQUIRED MODCOD}}.$$

This required C/N guarantees the target BER and it depends on the Modulation and Coding and on the channel (in general considered as being AWGN).

Link margin is generally defined as the difference between the available Signal to Noise plus Interference ratio and the required signal to noise ratio:

$$M = \left[\frac{C}{N+I}\right](dB) - \left[\frac{C}{N}\right]_{\text{REQUIRED MODCOD}}(dB).$$

11.7.2.2 Satellite fade margin

When there is a fade in the downlink (cause by shadowing, partially blockage of the signal, etc.) we can expect the nominal signal power to be degraded by an attenuation α.

$$C_{fade} = C.\alpha$$

When the main source of interference comes from the intermodulation and from the co-frequency beams the fade affecting the carrier also affects the interference. Therefore the ratio C/I remains unchanged.

$$\left[\frac{C.\alpha}{I.\alpha}\right] = \left[\frac{C}{I}\right] \text{ if } I = I_{IM} + I_{CC}\,.$$

And therefore the overall C/(N+I) in the presence of fade affecting both the useful signal and the main sources of interference is as follows:

$$\left[\frac{C}{N+I}\right]_{Fadedownlink}^{-1} = \left[\frac{C\alpha}{N}\right]^{-1} + \left[\frac{C}{I}\right]^{-1}.$$

Now, by fade margin we refer to the value of α so that the link budget can be closed (i.e. is equal or greater than the required):

$$\left[\left[\frac{C\alpha}{N}\right]^{-1}+\left[\frac{C}{I}\right]^{-1}\right]^{-1}=\left[\frac{C}{N}\right]_{\text{REQUIRED MODCOD}}.$$

And therefore, the fade margin (in dB) can be calculated by:

$$M_{fade}=-\alpha(dB)=\left[\frac{C}{N}\right](dB)+10\log\left[\left[\frac{C}{N}\right]^{-1}_{\text{REQUIRED MODCOD}}-\left[\frac{C}{I}\right]^{-1}\right].$$

When the link budget is not limited by the interference, i.e. when C/I>> required C/(N+I) and C/(N+I)≈C/N; or when the interference is dominated by the components not affected by the same fade as the signal (i.e. adjacent channel interference or co channel interference from other systems), then the fade margin is the same as the link margin:

$$M_{fade}\approx\left[\frac{C}{N}\right](dB)-\left[\frac{C}{N+I}\right]_{REQUIREDMODCOC}\approx M\ .$$

11.7.3 Required margins computation

The above stated formulas can be used to calculate the Signal to Noise ratio received by the terminal in a AWGN channel, but in order to close the link in the propagation channels applicable to DVB-SH, it is required to assess the additional fade endured by the satellite signal due to the mobile propagation channel. Contrary to the terrestrial case, the satellite propagation channel is considered non-frequency selective. Therefore, the fade associated with the satellite to mobile propagation channel will homogeneously affect the satellite signal along all its bandwidth.

This clause analyses the required fade margins for the satellite link. As stated before, these margins depend on the reception conditions and on the environment considered.

For reception condition A, the methodology proposed here considers that the terminal is in direct LOS and in a static situation (neither the terminal nor the surrounding geometry changes or changes are very slow compared to physical and upper layer countermeasures available in DVB-SH). In this static condition, time diversity mechanisms such as the physical layer interleaver or the link layer FEC do not provide any advantage. Fade due to multipath needs to be counteracted via a modulation and coding that allows for a sufficient link margin (i.e. $M_{fade}>Fade$) or via spatial diversity mechanisms such as antenna diversity (the latter not analyzed in the following).

Concerning reception conditions C and D, the mechanisms introduced in DVB-SH to counteract the signal fluctuations in the satellite to mobile channel impact the link margin required to guarantee the service quality (ESR(5)). If in general the margin is calculated so that the link budget is closed under fade conditions for a certain modulation and coding (and a given BER), in the presence of physical layer FEC interleaver and/or link layer FEC, the quality criteria (ESR(5)) can be met even if the fade temporarily exceeds the available margin. In fact, within the physical layer or link layer protection time, the fade can exceed the available margin for a period of time as long as this period does not go beyond the correction capabilities of the interleaver or link layer. The correction capabilities depend on the specific implementation of the physical or link layer protection. For any given link margin, the overall signal quality performance depends on this correction capability and on the dynamics of the channel (i.e. kind of environment and terminal speed). Therefore, it is not possible to derive a required fade or link margin value of general validity. The margins proposed in the clause dedicated to mobile reception are based on the available physical layer simulation results for different environments. A methodology will be also proposed on how this results can be used to calculate the availability over the coverage area.

11.7.3.1 Margins required reception condition A

In portable reception conditions addressed in this clause we consider that the terminal is in a static situation where the fade due to multipath can be counteracted only with sufficient link margin.

In the case of reception condition A (i.e. outdoor pedestrian), satellite reception is associated to a rural environment. Direct reception of satellite signal to handheld terminals (terminal categories 2a, 2b) is achieved only when the terminal is in direct Line-of-Sight. In these conditions, the satellite-to-mobile terminal channel can be considered as a Ricean channel with a Carrier to Multipath Ratio that depends on the geometry of the environment. In order to calculate the required margin associated to the Ricean channel, the following calculations apply:

Let us consider a signal *s* with a Signal to Noise Ratio in AWGN channel hereafter called $[C/N]_{AWGN}$ (calculated according to the formulae given in the previous clause). This signal experiences a fade that instantaneously degrades the received *C/N*. This instantaneous *C/N*, hereafter called γ, can be expressed as follows:

$$\gamma = \left[\frac{C}{N}\right]_{AWGN} r^2 \text{ where } r \text{ is the fading process amplitude.}$$

In the case of a Ricean channel the distribution of the fading amplitude *r* is given by the Rice distribution (see also clause A.7):

$$p_R(r) = \frac{r}{\sigma^2} e^{-\left(r^2+s^2\right)/2\sigma^2} I_0\left(\tfrac{rs}{\sigma^2}\right),$$

where σ_2 is the variance of either the real or the imaginary components of the multipath; and

s is the amplitude of the LOS signal component.

To find the distribution of the instantaneous *Signal to Noise ratio* γ from the Rice distribution, we use the identity:

$$p_\Gamma(\gamma) = p_R(r)\frac{dr}{d\gamma}.$$

We also define *Rice factor K* as in clause A7 and we normalize to unity the LOS signal power:

$$K = \frac{s^2}{2\sigma^2};\ s^2 = 1.$$

Therefore, we obtain the following expression for the Probability Density Function (PDF) of the *instantaneous Signal to Noise ratio* γ:

$$p_\Gamma(\gamma) = \frac{K}{\left[\frac{C}{N}\right]_{AWGN}} e^{-K\left(\frac{\gamma}{\left[\frac{C}{N}\right]_{AWGN}}+1\right)} I_0\left(2K\sqrt{\frac{\gamma}{\left[\frac{C}{N}\right]_{AWGN}}}\right).$$

And the CDF of γ can be expressed by:

$$CDF_\Gamma(\gamma) = \int_{-\infty}^{\gamma} p_x(x)\,dx = 1 - Q_1\left(\sqrt{2K}, \sqrt{2K}\sqrt{\frac{\gamma}{\left[\frac{C}{N}\right]_{AWGN}}}\right).$$

where Q_1 is the Marcum's Q function:

$$Q_1(a,b) = e^{-\left(a^2+b^2\right)/2} \sum_{t=o}^{\infty} \left(\tfrac{a}{b}\right)^t I_t(ab) \quad b > a > 0$$

According to the above expressions, the following figure represents the *cdf* of the degradation of instantaneous Signal to Noise Ratio with respect to the Signal to Noise Ratio in AWGN for different values of Carrier to Multipath ratio *K*.

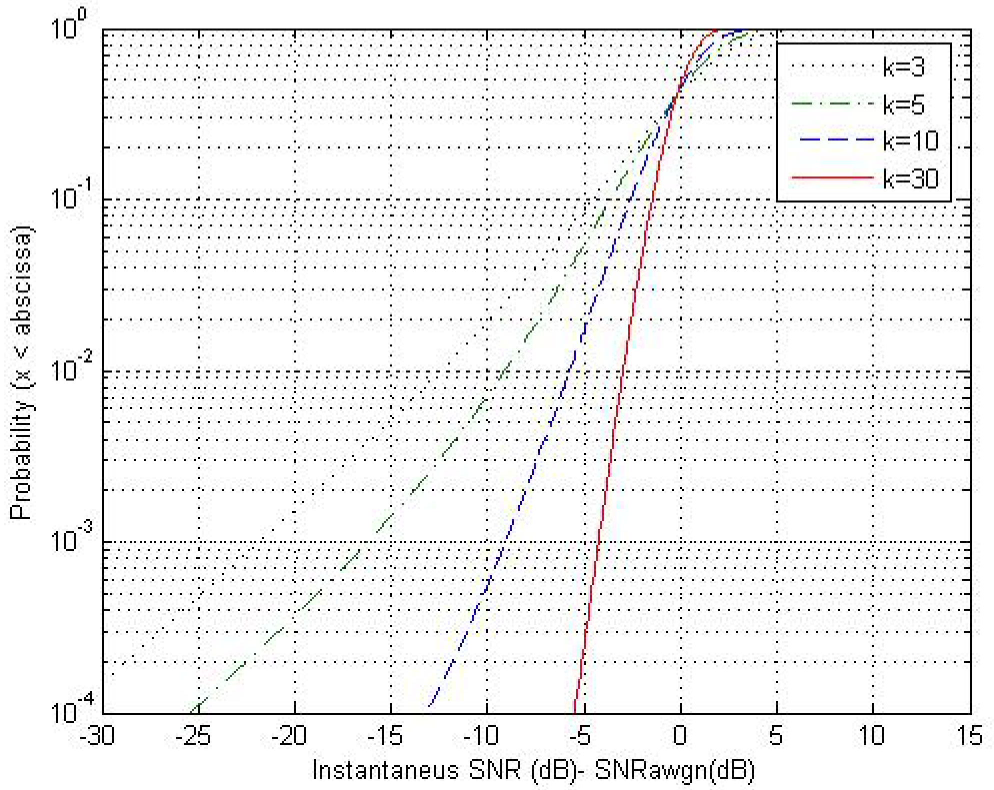


Figure 11.2: Ricean fade power loss CDF for different Rice factors K (in dB)

In a mobile channel, this instantaneous *Signal to Noise ratio* γ, is a time variant variable. When the terminal is in a static situation, as it is the case of category A reception, the Signal to Noise ratio becomes a position dependent variable. The time probability can be mapped into coverage probability and therefore the margins required to reach a certain percentage of the coverage can be calculated using the CDF above. This approach is homologous to the location correction factor used in the terrestrial coverage calculations seen earlier.

$$P_{out} = \Pr\left\{ \gamma < \left[\frac{C}{N}\right]_{REQ} \right\} = \Pr\left\{ \left[\frac{C}{N}\right]_{AWGN} - \gamma > \left[\frac{C}{N}\right]_{AWGN} - \left[\frac{C}{N}\right]_{REQ} \right\}$$

Where the quantity $\left[\frac{C}{N}\right]_{AWGN} - \left[\frac{C}{N}\right]_{REQ}$ represents the Fade Margin as defined earlier (in the case of no interference).

According to the previous calculations, the following figure summarizes the fade margin required for different values of Rice factor *K* and different percentages of covered area:

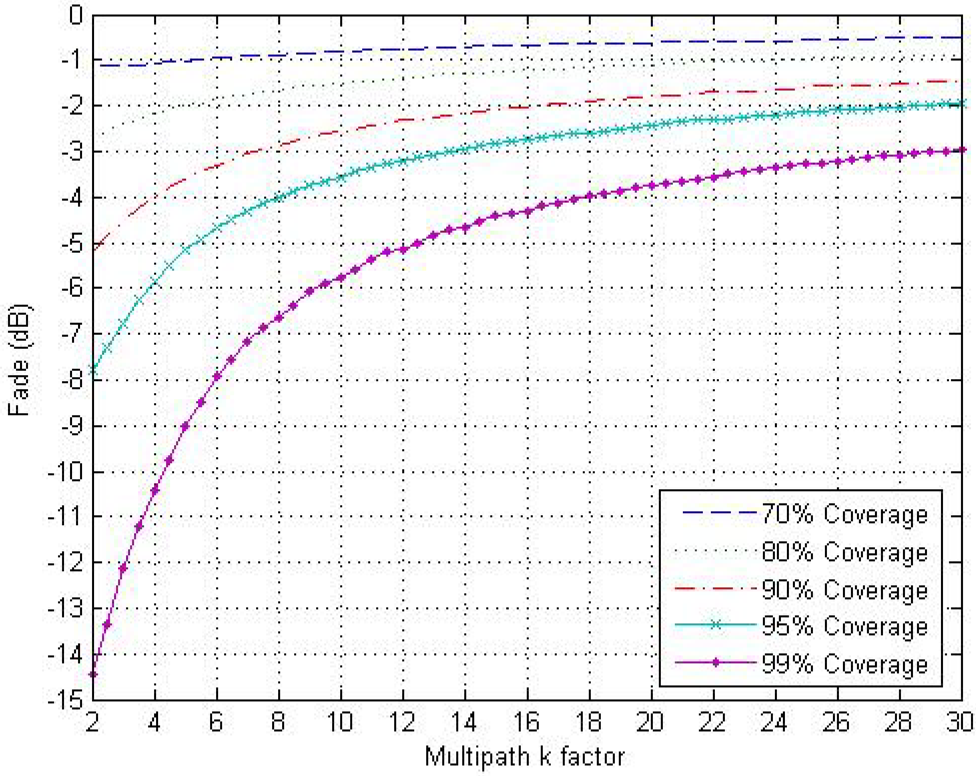


Figure 11.3: Fade margin required for different carrier to multipath ratios and coverage probability

Typical values for the Rice factor *K* range between 5 dB and 10 dB for hand-held terminal. Values can be higher in open areas.

It has to be noted that in the case of static reception conditions, the presence of a long physical layer interleaver or link layer FEC does not provide a real advantage in terms of additional coverage or availability since these mechanisms are effective to counteract the fluctuations of fade that occur only when the terminal is on the move. Instead the presence of mobile terminal antenna diversity may greatly help to reduce the required link margin for this case.

11.7.3.2 Margins required in reception conditions C

Mobile reception conditions for satellite refer to reception condition C for the rural environment. Mobile speeds analyzed in this clause cover also low speeds (3 kmph).

The Empirical Roadside Model briefly described in clause 4 allows determining the shadowing endured by the satellite link for a certain coverage probability in an equivalent way as the location correction factor does for the terrestrial network planning. However, as discussed previously, DVB-SH implements a long physical layer interleaver or/and a Link Layer FEC that alters the amount of shadowing that can be endured by the satellite link (without causing outage) when the terminal is moving with a certain speed. Therefore, this model cannot be directly translated into a required shadowing margin for our network planning calculations.

In order to take into account the time diversity introduced by the DVB-SH physical and link layer, the network planning needs to rely on the simulations of the physical and link layer over a channel model with representative fade/interfade durations. In the case of DVB-SH, many simulations have been performed using the time series generated with the well-known Perez-Fontan LMS model [16]. clause A.12 presents the results obtained for different environments, speeds and fade margins. These results give an indication of the achievable ESR(5) criteria in different environments for a given LOS *C/N*. The required margin for satellite reception is heavily dependent on the satellite elevation angle, the location, the type of environment, the mobile terminal speed and on the receiver physical and link layer configuration (overall redundancy, length and configuration of the physical layer interleaver and/or link layer FEC, etc.). Therefore, the tables in clause A.12 present a set of representative DVB-SH cases but cannot be considered to have general or universal applicability.

Clause A.12 simulations are organized by environments that are considered representative of DVB-SH usage (Intermediate Tree Shadowing, Suburban, etc.). Network planners need to take into consideration the percentage of time where a DVB-SH user is under each of the analyzed environments in order to derive the final availability.

11.7.4 Link budget examples

This clause presents several examples of satellite link budgets calculated following the formulae given above. It also presents the available fade margin for different physical layer configurations.

The selected link budget parameters are chosen to be representative of a DVB-SH GEO satellite system, although the link budget parameters (EIRP and interference contribution, etc.) always depend on the specific system configuration.

For these link budget calculations two satellite EIRP values have been selected: 63 dBW and 68 dBW which are representative of low and medium power satellites. It has also been considered an overall downlink C/I component of 14 dB, where the interference contribution comes from the inter-modulation noise and from the co-channel intra system interference. Both interference components are affected by fade in the same amount as the signal and therefore the distinction between fade and link margin defined in clause 11.7.2.2 applies.

In the following link budgets, there are four columns representative of the four different categories of terminals defined in clause 10. Main terminal characteristics (polarization, antenna gain, noise figure) are extracted from clause 10.

11.7.4.1 Link budget for SH-A

Table 11.14 represents the link budget for SH-A configuration in the case of a satellite with an effective EIRP towards the analyzed beam of 63 dBW and a 5 MHz channel.

Table 11.14: SH-A link budget for 5 MHz channel and 63 dBW EIRP

SH-A, 5 MHZ channel						
	Unit	**Handheld category 3**	**Portable category 2b (see note)**	**Portable category 2a**	**Vehicular category 1**
OFDM noise bandwidth	MHz	4,76	4,76	4,76	4,76
Up-link C/(N+I)	dB	19,5	19,5	19,5	19,5
Satellite transmission					
Tx frequency	GHz	2,2	2,2	2,2	2,2
EIRP effective/beam	dBW	**63,0**	**63,0**	**63,0**	**63,0**
Sat to RX Terminal propagation					
Free Space Loss	dB	190,7	190,7	190,7	190,7
Atmospheric attenuation	dB	0,5	0,5	0,5	0,5
Total attenuation	dB	191,2	191,2	191,2	191,2
Terminal Rx reception					
Terminal G/T	dB/K	-32,1	-29,1	-24,9	-21,0
Polarization losses	dB	3,0	3,0	3,0	0,0
Downlink Results					
C/N down-link	dB	-1,8	1,2	5,5	12,3
C/I down-link total	dB	14,0	14,0	14,0	14,0
Down-link C/(N+I)	dB	-1,9	1,0	4,9	10,1
Total C/(N+I)	dB	-1,9	0,9	4,7	9,6
NOTE: The results of this link budget are also representative of a handheld terminals with circular polarization.					

This link budget can be closed for different physical layer configurations, given different fade margins. In order to close the link, the required C/N and implementation margin values shown in clauses 7.2.2.6 and 10.4.3 respectively are chosen. Tables 11.15 to 11.17 show the link budget closure for three different physical layer configuration (Code rate 1/2, 1/3 and 1/5).

Table 11.15: Link budget closure for code rate 1/2 at physical layer and 63 dBW EIRP

Physical Layer	**Unit**	**Handheld category 3**	**Portable category 2b**	**Portable category 2a**	**Vehicular category 1**
		CODE 1/2 at PL			
DVB-SH mode		2K	2K	2K	2K
Number of carriers		1 705	1 705	1 705	1 705
Number of useful carriers		1 512	1 512	1 512	1 512
Useful symbol duration	msec	358	358	358	358
Guard Time fraction		¼	¼	¼	¼
Modulation Order		QPSK	QPSK	QPSK	QPSK
Physical Layer Coding rate		½	½	½	½
Useful bit rate at physical layer	Mb/s	3,38	3,38	3,38	3,38
OFDM noise bandwidth	MHz	4,76	4,76	4,76	4,76
Required C/N at physical layer at BER 10^-5	dB	1,4	1,4	1,4	1,4
Implementation Loss in AWGN	dB	1,1	1,1	1,1	1,1
Received C/(N+I)	dB	-1,9	0,9	4,7	9,6
LOS Margin at Physical Layer w.r.t. AWGN	**dB**	**-4,4**	**-1,6**	**2,2**	**7,1**
Available Fade Margin at Physical Layer	**dB**	**-4,7**	**-1,7**	**2,5**	**9,4**

Table 11.16: Link budget closure for code rate 1/3 at physical layer

Physical Layer	Unit	Handheld category 3	Portable category 2b	Portable category 2a	Vehicular category 1
		CODE 1/3 at PL			
DVB-SH mode		2K	2K	2K	2K
Number of carriers		1 705	1 705	1 705	1 705
Number of useful carriers		1 512	1 512	1 512	1 512
Useful symbol duration	msec	358	358	358	358
Guard Time fraction		¼	¼	¼	¼
Modulation Order		QPSK	QPSK	QPSK	QPSK
Physical Layer Coding rate		1/3	1/3	1/3	1/3
Useful bit rate at physical layer	Mb/s	2,25	2,25	2,25	2,25
OFDM noise bandwidth	MHz	4,76	4,76	4,76	4,76
Required C/N at physical layer at BER 10^-5	dB	-0,9	-0,9	-0,9	-0,9
Implementation Loss in AWGN	dB	1,1	1,1	1,1	1,1
Received C/(N+I)	dB	-1,9	0,9	4,7	9,6
LOS Margin at Physical Layer w.r.t. AWGN	**dB**	**-2,12**	**0,73**	**4,54**	**9,40**
Available Fade Margin at Physical Layer	**dB**	**-2,22**	**0,78**	**5,02**	**11,87**

Table 11.17: Link budget closure for code rate 1/5 at physical layer

Physical Layer	Unit	Handheld category 3	Portable category 2b	Portable category 2a	Vehicular category 1
		CODE 1/5 at PL			
DVB-SH mode		2K	2K	2K	2K
Number of carriers		1 705	1 705	1 705	1 705
Number of useful carriers		1 512	1 512	1 512	1 512
Useful symbol duration	msec	358	358	358	358
Guard Time fraction		¼	¼	¼	¼
Modulation Order		QPSK	QPSK	QPSK	QPSK
Physical Layer Coding rate		1/5	1/5	1/5	1/5
Useful bit rate at physical layer	Mb/s	1,35	1,35	1,35	1,35
OFDM noise bandwidth	MHz	4,76	4,76	4,76	4,76
Required C/N at physical layer at BER 10^-5	dB	-3,6	-3,6	-3,6	-3,6
Implementation Loss in AWGN	dB	1,1	1,1	1,1	1,1
Received C/(N+I)	dB	-1,9	0,9	4,7	9,6
LOS Margin at Physical Layer w.r.t. AWGN	**dB**	**0,58**	**3,43**	**7,24**	**12,10**
Available Fade Margin at Physical Layer	**dB**	**0,60**	**3,59**	**7,83**	**14,68**

In the same way, the tables 11.18 to 11.21 apply for the a medium power satellite radiating 68 dBW towards the beam.

Table 11.18: SH-A link budget for 5 MHz channel and 68 dBW EIRP

Physical Layer	Unit	DVB-SH-A, 5 MHZ channel			
		Handheld category 3	Portable category 2b	Portable category 2a	Vehicular category 1
FDM noise bandwidth	MHz	4,76	4,76	4,76	4,76
Up-link C/(N+I)	dB	19,5	19,5	19,5	19,5
Satellite transmission					
Tx frequency	GHz	2,2	2,2	2,2	2,2
EIRP effective/beam	dBW	68,0	68,0	68,0	68,0
Sat to RX Terminal propagation					
Free Space Loss	dB	190,7	190,7	190,7	190,7
Atmospheric attenuation	dB	0,5	0,5	0,5	0,5
Total attenuation	dB	191,2	191,2	191,2	191,2
Terminal Rx reception					
Terminal G/T	dB/K	-32,1	-29,1	-24,9	-21,0
Polarization losses	dB	3,0	3,0	3,0	0,0
Downlink Results					
C/N down-link	dB	3,2	6,2	10,4	17,3
C/I down-link total	dB	14,0	14,0	14,0	14,0
Down-lin k C/(N+I)	dB	2,8	5,5	8,8	12,3
Total C/(N+I)	dB	2,8	5,4	8,5	11,6

Table 11.19: Link budget closure for code rate 1/2 at physical layer

Physical Layer	Unit	Handheld category 3	Portable category 2b	Portable category 2a	Vehicular category 1
		CODE 1/2 at PL			
DVB-SH mode		2K	2K	2K	2K
Number of carriers		1 705	1 705	1 705	1 705
Number of useful carriers		1 512	1 512	1 512	1 512
Useful symbol duration	msec	358	358	358	358
Guard Time fraction		¼	¼	¼	¼
Modulation Order		QPSK	QPSK	QPSK	QPSK
Physical Layer Coding rate		½	½	½	½
Useful bit rate at physical layer	Mb/s	3,38	3,38	3,38	3,38
OFDM noise bandwidth	MHz	4,76	4,76	4,76	4,76
Required C/N at physical layer at BER 10^-5	dB	1,4	1,4	1,4	1,4
Implementation Loss in AWGN	dB	1,1	1,1	1,1	1,1
Received C/N	dB	2,76	5,35	8,49	11,57
LOS Margin at Physical Layer w.r.t. AWGN	**dB**	**0,3**	**2,9**	**6,0**	**9,1**
Available Fade Margin at Physical Layer	**dB**	**0,3**	**3,3**	**7,5**	**14,4**

Table 11.20: Link budget closure for code rate 1/3 at physical layer

Physical Layer	Unit	Handheld category 3	Portable category 2b	Portable category 2a	Vehicular category 1
		CODE 1/3 at PL			
DVB-SH mode		2K	2K	2K	2K
Number of carriers		1 705	1 705	1 705	1 705
Number of useful carriers		1 512	1 512	1 512	1 512
Useful symbol duration	msec	358	358	358	358
Guard Time fraction		¼	¼	¼	¼
Modulation Order		QPSK	QPSK	QPSK	QPSK
Physical Layer Coding rate		1/3	1/3	1/3	1/3
Useful bit rate at physical layer	Mb/s	2,25	2,25	2,25	2,25
OFDM noise bandwidth	MHz	4,76	4,76	4,76	4,76
Required C/N at physical layer at BER 10^-5	dB	-0,9	-0,9	-0,9	-0,9
Implementation Loss in AWGN	dB	1,1	1,1	1,1	1,1
Received C/N	dB	2,76	5,35	8,49	11,57
LOS Margin at Physical Layer w.r.t. AWGN	**dB**	**2,56**	**5,15**	**8,29**	**11,37**
Available Fade Margin at Physical Layer	**dB**	**2,76**	**5,75**	**9,99**	**16,84**

Table 11.21: Link budget closure for code rate 1/5 at physical layer

Physical Layer	Unit	Handheld category 3	Portable category 2b	Portable category 2a	Vehicular category 1
		CODE 1/5 at PL			
DVB-SH mode		2K	2K	2K	2K
Number of carriers		1 705	1 705	1 705	1 705
Number of useful carriers		1 512	1 512	1 512	1 512
Useful symbol duration	msec	358	358	358	358
Guard Time fraction		¼	¼	¼	¼
Modulation Order		QPSK	QPSK	QPSK	QPSK
Physical Layer Coding rate		1/5	1/5	1/5	1/5
Useful bit rate at physical layer	Mb/s	1,35	1,35	1,35	1,35
OFDM noise bandwidth	MHz	4,76	4,76	4,76	4,76
Required C/N at physical layer at BER 10^-5	dB	-3,6	-3,6	-3,6	-3,6
Implementation Loss in AWGN	dB	1,1	1,1	1,1	1,1
Received C/N	dB	2,76	5,35	8,49	11,57
LOS Margin at Physical Layer w.r.t. AWGN	**dB**	**5,26**	**7,85**	**10,99**	**14,07**
Available Fade Margin at Physical Layer	**dB**	**5,57**	**8,56**	**12,80**	**19,65**

11.7.4.2 Link budget for SH-B

Table 11.22 represents the link budget for SH-B configuration in the case of a satellite with an effective EIRP towards the analyzed beam of 63 dBW and a 5 MHz channel.

Table 11.22: SH-B link budget for 5 MHz channel and 63 dBW EIRP

	Unit	DVB-SH-B, 5 MHZ channel			
		Handheld category 3	**Portable category 2b**	**Portable category 2a**	**Vehicular category 1**
TDM occupied bandwidth	MHz	4,89	4,89	4,89	4,89
Up-link C/(N+I)	dB	20,0	20,0	20,0	20,0
Satellite transmission					
Tx frequency	GHz	2,2	2,2	2,2	2,2
EIRP effective/beam	dBW	63,0	63,0	63,0	63,0
Sat to RX Terminal propagation					
Free Space Loss	dB	190,7	190,7	190,7	190,7
Atmospheric attenuation	dB	0,5	0,5	0,5	0,5
Total attenuation	dB	191,2	191,2	191,2	191,2
Terminal Rx reception					
Terminal G/T	dB/K	-32,1	-29,1	-24,9	-21,0
Polarization losses	dB	3,0	3,0	3,0	0,0
Downlink Results					
C/N down-link	dB	-1,3	1,7	5,9	12,8
C/I down-link total (adjacent, co-channel, NPR, etc.)	dB	14,0	14,0	14,0	14,0
Down-link C/(N+I)	dB	-1,4	1,5	5,3	10,3
Total C/(N+I)	**dB**	-1,4	**1,4**	**5,2**	**9,9**

As before, this link budget can be closed for different physical layer configurations. Tables 11.23 to 11.25 show the link budget closure for three different physical layer configuration (Code rate 1/2, 1/3 and 1/5) according to the thresholds and implementation margins shown in clause 7 for the TDM case.

Table 11.23: SH-B link budget closure for code rate 1/2 at physical layer and 63 dBW EIRP

Physical Layer		**Handheld category 3**	**Portable category 2b**	**Portable category 2a**	**Vehicular category 1**
		CODE 1/2 at PL			
TDM Roll-off	-	0,15	0,15	0,15	0,15
TDM Symbol rate	-	4,25	4,25	4,25	4,25
Modulation	-	QPSK	QPSK	QPSK	QPSK
Physical Layer Coding rate	**-**	**1/2**	**1/2**	**1/2**	**1/2**
Useful bit rate at physical layer (minus 8 % of pilots)	Mb/s	3,91	3,91	3,91	3,91
TDM noise bandwidth	MHz	4,25	4,25	4,25	4,25
Required C/N at physical layer at BER 10^-5 in AWGN	dB	1,1	1,1	1,1	1,1
Implementation Loss in AWGN channel	dB	0,5	0,5	0,5	0,5
Received C/N	dB	**-1,4**	**1,4**	**5,2**	**9,9**
Link budget results					
LOS Margin at Physical Layer w.r.t. AWGN	**dB**	**-3,0**	**-0,2**	**3,6**	**8,3**
Available Fade Margin at Physical Layer	**dB**	**-3,2**	**-0,2**	**4,0**	**10,9**

Table 11.24: SH-B link budget closure for code rate 1/3 at physical layer and 63 dBW EIRP

Physical Layer		Handheld category 3	Portable category 2b	Portable category 2a	Vehicular category 1
		CODE 1/3 at PL			
TDM Roll-off	-	0,15	0,15	0,15	0,15
TDM Symbol rate	-	4,25	4,25	4,25	4,25
Modulation	-	QPSK	QPSK	QPSK	QPSK
Physical Layer Coding rate	**-**	**1/3**	**1/3**	**1/3**	**1/3**
Useful bit rate at physical layer (minus 8 % of pilots)	Mb/s	2,61	2,61	2,61	2,61
TDM noise bandwidth	MHz	4,25	4,25	4,25	4,25
Required C/N at physical layer at BER 10^-5 in AWGN	dB	-1,2	-1,2	-1,2	-1,2
Implementation Loss in AWGN channel	dB	0,5	0,5	0,5	0,5
Received C/N	dB	**-1,4**	**1,4**	**5,2**	**9,9**
Link budget results					
LOS Margin at Physical Layer w.r.t. AWGN	**dB**	**-0,7**	**2,1**	**5,9**	**10,6**
Available Fade Margin at Physical Layer	**dB**	**-0,8**	**2,2**	**6,5**	**13,3**

Table 11.25: SH-B link budget closure for code rate 1/5 at physical layer and 63 dBW EIRP

Physical Layer		Handheld category 3	Portable category 2b	Portable category 2a	Vehicular category 1
		CODE 1/5 at PL			
TDM Roll-off	-	0,15	0,15	0,15	0,15
TDM Symbol rate	-	4,25	4,25	4,25	4,25
Modulation	-	QPSK	QPSK	QPSK	QPSK
Physical Layer Coding rate	**-**	**1/5**	**1/5**	**1/5**	**1/5**
Useful bit rate at physical layer (minus 8 % of pilots)	Mb/s	1,56	1,56	1,56	1,56
TDM noise bandwidth	MHz	4,25	4,25	4,25	4,25
Required C/N at physical layer at BER 10^-5 in AWGN	dB	-3,9	-3,9	-3,9	-3,9
Implementation Loss in AWGN channel	dB	0,5	0,5	0,5	0,5
Received C/N	dB	**-1,4**	**1,4**	**5,2**	**9,9**
Link budget results					
LOS Margin at Physical Layer w.r.t. AWGN	**dB**	**2,0**	**4,8**	**8,6**	**13,3**
Available Fade Margin at Physical Layer	**dB**	**2,0**	**5,0**	**9,2**	**16,1**

The same approach can also be repeated for a satellite radiating 68 dBW per beam, obtaining 5 dB more margin for the same physical layer considered previously.

11.7.4.3 Example of availability calculation for reception condition A

The fade margin derived in the above link budget can be used to calculate the availability obtained for portable category of reception in a given area conditioned to the multipath factor over that area. Tables 11.26 and 11.27 show the availability obtained for the different physical layer configurations and for the different terminal types. The availability results are derived by matching the fade margin obtained in each of the link budgets with the fade margin required for a given multipath factor shown in figure 11.2.

Table 11.26: Availability for reception condition A and 63 dBW of EIRP

EIRP=63 dBW SH-A	QPSK 1/2		QPSK 1/3		QPSK 1/5	
	Fade margin available	**Availability**	**Fade margin available**	**Availability**	**Fade margin available**	**Availability**
-4,7 dB		Out of Coverage	-2,2 dB	Out of Coverage	0,6 dB	70 % (if *K*>15 dB)
Terminal Category 2b (see note)	-1,7 dB	Out of Coverage	0,8 dB	70 % (*K*>12 dB)	3,6 dB	80 % 90 %(if *K*>10 dB)
Terminal Category 2a	2,5 dB	80 % 90 % (if *K*>10 dB)	5,0 dB	90 % 95 % (if *K*>5 dB)	7,8 dB	95 % 99 % (if *K*>6 dB)
NOTE: Also representative of terminal category 3 (handheld) with circular polarized antenna.						

Table 11.27: Availability for reception condition A and 68 dBW of EIRP

EIRP=68 dBW SH-A	QPSK 1/2		QPSK 1/3		QPSK 1/5	
	Fade margin available	**Availability**	**Fade margin available**	**Availability**	**Fade margin available**	**Availability**
Terminal Category 3	0,3 dB	Coverage only in AWGN	2,8 dB	80 % 90 % (if K>10 dB)	5,6 dB	90 % 95 % (if K>6 dB)
Terminal Category 2b	3,3 dB	80 % 90 % (K>6 dB)	5,75 dB	90 % 95 %(K>5 dB)	8,6 dB	95 % 99 %(if K>5 dB)
Terminal Category 2a	7,5 dB	95 % 99 % (if K>7 dB)	10 dB	95 % 99 % (if K>4 dB)	12,8 dB	95 % 99 % (if K>3 dB)

11.7.4.4 Example of availability calculation for reception condition C

The link budgets shown above have been used in the physical layer simulations presented in clause A. These simulations give quality in terms of ESR(5) obtained in a set of scenarios representative of DVB-SH satellite to mobile channel. Tables 11.28 to 11.31 summarizes the percentage of ESR(5) fulfillment for the Intermediate Tree Shadowing (ITS) and SubUrban (SU) environment. The terminal considered is a vehicular terminal (category 1) with a long physical layer interleaver.

Table 11.28: ESR(5) fulfilment for SH-A for the category 1 (vehicular) terminal in reception condition C

SH-A Long physical layer interleaver 63 dBW EIRP Terminal category 1 (Vehicular)			**Physical Layer**			
			QPSK 1/3		16QAM 1/5	
			Available Fade Margin = 11,9 dB (After implementation losses)		Available Fade Margin = 9,3 dB (After implementation losses)	
			Uniform	Uniform-Late	Uniform	Uniform-Late
Mobile Channel	ITS	50 kmph	100 %	99 %	95,5 %	86 %
	SU	50 kmph	100 %	100 %	100 %	99 %

Table 11.29: ESR(5) fulfilment for SH-A in reception condition A with terminal category 2b with circular polarized antenna

SH-A Long physical layer interleaver 68 dBW EIRP Terminal category 2b with linear polarized antenna and 3 with circular polarized antenna			Physical Layer			
			QPSK 1/3		16QAM 1/5	
			Available Fade Margin = 5,0 dB (After implementation losses)		Available Fade Margin = 2,2 dB (After implementation losses)	
			Uniform	Uniform-Late	Uniform	Uniform-Late
Mobile Channel	SU	3 kmph	84 %	75 %	51 %	42 %

Table 11.30: ESR(5) fulfilment for SH-B for Category 1 (vehicular) terminal in reception condition C

SH-B Long physical layer interleaver 63 dBW EIRP Terminal category 1 (Vehicular)			Physical Layer			
			QPSK 1/3		8PSK 2/9	
			Available Fade Margin = 13,3 dB (After implementation losses)		Available Fade Margin = 11,2 dB (After implementation losses)	
			Uniform	Uniform-Late	Uniform	Uniform-Late
Mobile Channel	ITS	50 kmph	97 %	96 %	95,3 %	93,7 %
	SU	50 kmph	98,2 %	97,3 %	99 %	96,8 %

Table 11.31: ESR(5) fulfilment for SH-B in reception condition A with terminal category 2b with circular polarized antenna

SH-A Long physical layer interleaver 68 dBW EIRP Terminal category 2b with linear polarized antenna and 3 with circular polarized antenna			Physical Layer			
			QPSK 1/3		8PSK 2/9	
			Available Fade Margin = 6,5 dB (After implementation losses)		Available Fade Margin = 4,4 dB (After implementation losses)	
			Uniform	Uniform-Late	Uniform	Uniform-Late
Mobile Channel	SU	3 kmph	77,9 %	72,6 %	75,2 %	73,5 %

11.7.5 Example of satellite coverage calculation

Previous clauses have explained the link budget calculations and the margins required for different reception conditions and different terminals. It has to be noted that physical layer simulations refer to one specific environment (Intermediate tree shadowing or suburban) and characterized the quality of DVB-SH physical layer under that environment. However, the satellite coverage is composed of a patchwork of different environments. Once the quality of DVB-SH signal in each environment has been characterized, the overall quality obtained in the satellite coverage can be calculated by a weighted average of the qualities obtained in each individual environment. The weighting factor depends on the percentage of the coverage area occupied by each scenario. This methodology has been explained in clause 11.7.1. Table 11.32 exemplifies this coverage calculation for terminal category 1 and 63 dBW of satellite EIRP.

Table 11.32: Example of satellite coverage calculation

Coverage calculation for vehicular terminal (category 1) and 63 dBW of satellite EIRP			
Environment	suburban	rural (open field)	ITS
Percentage of the satellite coverage per environment $P_E(i)$	15 %	50 %	35 %
Representative terminal speed	50 kmph	50 kmph	50 kmph
Target ESR(5)	90 %	90 %	90 %
Satisfaction index	1 (from simulations in clause A)	1	1 (from simulations in clause A)
Achieved coverage Csat(%)	100 %		

11.8 Hybrid network planning

11.8.1 Coverage improvements

Hybrid coverage provides improvement due to the signal combining.

11.8.2 Exclusion zones in multibeam hybrid networks

In mutibeam satellites, there is the possibility of large overlaps between adjacent beams (see for example clause 4, figure 4.2). In these overlapping areas a terrestrial transmitter in a beam may be interfered by the satellite signal of the adjacent beam, if it reuses the frequency of that adjacent beam. The "exclusion zone" is defined as:

Case (a), for the Local content only: the fringe coverage area of the transmitter carrying the Local content that would be covered if the interfering satellite signal were removed.

Case (b), for the Common content in an MFN (and only MFN): the fringe area of the transmitter carrying the Common content where the wanted satellite signal is not usable (e.g. blocked) and that would be covered if the interfering satellite signal were removed.

Annex A (informative): Data and methodology for DVB-SH assessment

This clause presents the data and methodology used by the DVB TM-SSP group to assess the efficiency of the specified DVB-SH Physical Layer, separately, and together with the selected Upper Layer complementary FEC.

TM-SSP members have agreed on a common simulation framework for a first-step performance assessment of the different DVB-SH configurations. This first-step focuses on the performance of the FEC (both physical layer and upper layer) and interleavers over the reference propagation channels that have been identified and agreed for benchmarking the DVB-SH system.

A.1 Geostationary Satellite Payloadcharacteristics (S-band)

A.1.1 Payload architectures

The satellite payload distortions are of two types: linear distortions caused by the filters amplitude and group delay and the nonlinear distortions (see note) caused by TWTA(s) operated near its saturation point. The general analysis is complicated by the fact that linear and nonlinear distorting elements are typically cascaded. Nevertheless, the linear distortions are of second order compared to the nonlinear distortions.

The target EIRP of a mobile TV satellite in S-band is in the order of 63 dBW to 72 dBW, with the 68 dBW being a possible limit between "medium" and "high" power satellites broadcasting to handheld devices. For comparison, the EIRP of satellites broadcasting to fixed reception in the Ku-band are in the region of 50 dBW to 55 dBW.

Antenna and power amplification architectures play an equally central role whose final objective is the achievement of the target EIRP and the desired coverage areas. Antenna gain is upper-bounded by the theoretical limit relevant to the solid angular view of the coverage area.

EIRP requirements and antenna performance limit lead in most cases to a demanding RF power that exceeds the generation capabilities of single amplifiers available for space applications and "power combining" must be adopted.

Two main payload classes can be distinguished depending on the power amplification and combining architecture: "Linear" and "Non-linear" classes.

For a single-beam satellite (or multibeam satellites that do not require flexible power-sharing between beams), the required EIRP is typically obtained by using parallel TWTAs through network or polarization combining. Each TWTA amplifies only one signal (but several TWTAs amplify the same signal). The TWTA can be driven close to saturation if needed. The satellite is then said to belong to the "non-linear" class, providing a power advantage for the TDM waveform, which does not require the same TWTA back-off as the OFDM waveform.

When there is a need to *reallocate available power to the different beams after the satellite is launched*, a different amplification architecture is adopted. In this case, the amplification is performed through a low-power BFN followed by a shared TWTA network exploiting a stack of multiport amplifiers (MPA) connected to the transmit antenna feeds. Each TWTA now amplifies a weighted sum of several signals. This architecture implements spatial power combining that provides the flexible power allocation to the various beams. This flexibility is achieved at the expense of operating the TWTAs in multicarrier mode (i.e. with sufficient back-off) and therefore the power advantage of TDM over OFDM is diminished.

NOTE: The multiplicative phase noise impact is discussed in clause. A.11.1.2.

A.1.2 TWTA characteristics and Operating Point Optimization

For further study.

A.2 Void

A.3 Simulations with ideal channel estimator in demodulator

A.3.1 System parameters

The first focus is to optimize the interleaving scheme according to the channel impairments. The imperfect channel estimator in the demodulator is neglected at this stage.

To limit the number of cases, the system parameters are those considered relevant to hybrid S-Band systems that are the first to make use of DVB-SH. Waveform spectrum efficiencies resulting in bitrates at MPEG-TS level ranging between 2,2 Mbps and 2,8 Mbps, for a 5 MHz channelization, are considered. The number of services chosen ranges between 8 and 11.

A reference value of roughly 10 s has been considered for the end-to-end delay introduced by the DVB-SH processing. This allowance is to be shared between the physical layer and the LL-FEC, Turbo FEC delays being negligible (one Turbo FEC block is in the order of several msec).

Different configurations selected are shown in tables A.1 to A.4. Direct comparison of power efficiency is possible when the bit rate is strictly the same, otherwise the bit rate should be kept in mind when comparing different cases. The interest to increase the modulation order while keeping the net bit rate is to take advantage of lower Turbo code rates that can bring higher channel erasure resilience. Also different sharing between PL and LL-FEC for class 1 receivers for the same overall spectral efficiency is investigated.

Finally, note that TDM and OFDM configurations are designed so that they can be synchronous in an SH-B system, so that code combining can be performed.

Table A.1: OFDM Reference configurations for Class 1 receivers (5 MHz)

Parameter/Case name	16QAM_1/4_S	16QAM_2/7_S	QPSK1/2_S	QPSK2/3_S
FFT Mode	2K+GI 1/4	2K+GI ¼	2K+GI 1/4	2K+GI 1/4
Modulation	16 QAM	16 QAM	QPSK	QPSK
PHY FEC rate	1/4	2/7	1/2	2/3
LL-FEC rate (recommended)	2/3	7/12	2/3	1/2
OFDM Symbols/ coded FEC	8,00	7,00	8,00	6,00
Services	8	8	8	8
Bit rate/service (at TS level)	279,8 kbps	273,6 kbps	279,8 kbps	277,7 kbps
MPEG TS total bit rate	3,357 Mbps	3,752 Mbps	3,357 Mbps	4,443 Mbps
Early interleaver duration	0 ms	0 ms	0 ms	0 ms
Late interleaver duration	211 ms	211 ms	211 ms	211 ms

Table A.2: OFDM Reference configurations for Class 2 receivers (5 MHz)

Parameter/Case name	QPSK1/3_U	QPSK_1/3_UL	16QAM_1/5_U	16QAM_1/5_UL
FFT Mode	2K+GI 1/4	2K+GI 1/4	2K+GI 1/4	2K+GI 1/4
Modulation	QPSK	QPSK	16 QAM	16 QAM
PHY FEC rate	1/3	1/3	1/5	1/5
LL-FEC rate (recommended)	1	1	1	1
OFDM Symbols/ coded FEC	12	12	10	10
# Services	8	8	9	9
Bit rate/service (at TS level)	277,7 kbps	277,7 kbps	296,2 kbps	296,2 kbps
MPEG TS Total bit rate	2.222 Mbps	2.222 Mbps	2.666 Mbps	2.666 Mbps
Early interleaver duration	11 000 ms	10 000 ms	11 000 ms	10 000 ms
Late interleaver duration	0 ms	215 ms	0 ms	215 ms

Table A.3: TDM Reference configurations for Class 1 receivers, compatible with OFDM QPSK (GI 1/4)

Parameter/Case name	T-8PSK1/3_S	T-QPSK1/2_S
Bandwidth (including R.O.)	4,89 MHz	4,89 MHz
Modulation	8PSK	QPSK
PHY FEC rate	1/3	1/2
UL FEC rate (recommended)	2/3	2/3
CW per SH-FRAME	79	79
Services	9	9
Bit rate/service (at TS level)	288,9 kbps	288,9 kbps
MPEG TS Total bit rate	3.900 Mbps	3.900 Mbps
Early interleaver duration	0 ms	0 ms
Late interleaver duration	190 ms	180 ms

Table A.4: TDM Reference configurations parameters for Class 2 receivers

Parameter/Case name	T-8PSK2/9_U	T-8PSK2/9_UL	T-QPSK1/3_U	T-QPSK1/3_UL
Bandwidth (including R.O.)	4,89 MHz	4,89 MHz	4,89 MHz	4,89 MHz
Modulation	8PSK	8PSK	QPSK	QPSK
PHY FEC rate	2/9	2/9	1/2	1/3
LL-FEC rate (recommended)	Not used	Not used	Not used	Not used
CW per SH-FRAME	52	52	52	52
Services	9	9	9	9
Bit rate/Service (at TS level)	285,2 kbps	285,2 kbps	285,2 kbps	285,2 kbps
MPEG TS Total bit rate	2.567 Mbps	2.567 Mbps	2.567 Mbps	2.567 Mbps
Early interleaver duration	11 000 ms	10 000 ms	11 000 ms	10 000 ms
Late interleaver duration	0 ms	180 ms	0 ms	180 ms

Table A.5: System definitions for OFDM simulations

Definition	System Parameter	Comment
Bandwidth definition	Noise bandwidth	5-MHz
	Signal bandwidth	Limited to modulated carriers (data + TPS + pilots)
Pilots insertion	Pilots as defined by DVB-T	boosted by 16/9
Total Signal Power	Total power of the received signal	Normalized to 1 All Channels (TU6 inclusive) It Affects the Es/N0 calculation
Definition of C/N	ratio of useful power to noise power in the used bandwidth (= non-zero subcarriers)	
Definition of S/N	$\frac{S}{N}=\frac{C}{N}.\frac{(N_{c-U}+N_{c-TPS}+N_{c-Pilot})}{N_{FFT}}$	According to "Bandwidth definition" and "Total Signal Power"
Definition of Es/N0	$\frac{E_{S-U}}{N_0}=\frac{S}{N}.\frac{N_{FFT}}{(N_{c-U}+N_{c-TPS}+16/9.N_{c-Pilot})}$	ES-U accounts for the useful energy per symbol N_FFT is the total number of carriers (FFT-size) Nc-U is the number of useful carrier Nc-TPS is the number of TPS Nc-Pilots is the number of pilots

Table A.6: System definitions for TDM simulations

Definition	System Parameter	Comment
Bandwidth definition	Noise bandwidth = Symbol rate	Square-root raised cosine
	Signal bandwidth = [Symbol rate]*1,15	Roll-off 15 % is proposed
Pilots insertion	80 symbols every 1 008 data symbols	Higher than DVB-S2: 36/1 440
Total Signal Power	Total power of the received signal	Normalized to 1 All Channels (TU6 inclusive) It Affects the Es/N0 calculation
Definition of C/N	ratio of useful power to noise power in the used bandwidth (=pilot power is taken into account)	
Definition of Es/N0	Ratio of Symbol energy to noise density	To be compared with FEC simulations without pilot symbols
Ratio C/N and Es/N0	$\frac{C}{N} = \frac{E_S}{N0}$	Pilots have the same power as data symbols. So the power is conserved.

A.3.2 Physical layer FEC and De-mapper configuration

Table A.7 details the algorithm and parameters chosen for the simulated decoder and de-mapper. Other implementations may use different methods and parameters that could lead to slightly different results.

Table A.7: Physical layer FEC configuration

Criteria	Simulation Parameters	Comment
Code	Turbo Code 3GGP2	3GPP2 C.S0002-D, Version 1.0, Date: February 13, 2004
Information Block size	12 282 bits	
Decoder engine Implementation	LOG-MAP	
Number of iterations	8	Constant
De-mapper engine Implementation	LOG-MAP	as "Decoder engine implementation" - It Affects the LLR calculation

A.4 Channel interleaver

Several configurations of channel interleaver are considered. They are tested over the different channels (terrestrial or LMS) in order to assess their performances and show which are the required margins depending on the use case.

The channel interleaver configurations intended for Class 1 receivers are used also in a 2-step simulation approach to provide performances at physical layer that can be exploited for further evaluation of link layer FEC improvement. This is achieved via the production of dumps tracing errored MPEG packets at physical layer which are exploited to get a performance at service level provided a redundancy rate and a FEC. This methodology is further explained in clause A.6.

The figures provided hereafter are meant to be used as a reference for translation of TPS or frame header signalled values into actual receiver interleaver configurations.

TPS elements are given in table A.8 and detailed interleavers branches values are given in clause A.10.

Table A.8: Interleaver configurations

	Configuration	Common multiplier	Number of late taps	Number of slices	Slices distance	Non-late increment
OFDM Class 1	16QAM_1/4_S	10	48	1	0	0
	16QAM_2/7_S	10	48	1	0	0
	QPSK1/2_S	5	48	1	0	0
	QPSK2/3_S	5	48	1	0	0
OFDM Class 2	16QAM_1/5_U	40	0	12	8	4
	16QAM_1/5_UL	20	24	9	10	12
	QPSK1/3_U	40	0	12	4	2
	QPSK_1/3_UL	10	24	9	5	12
TDM Class 1	T-8PSK1/3_S	8	48	1	0	0
	T-QPSK1/2_S	5	48	1	0	0
TDM Class 2	T-8PSK2/9_U	60	0	12	8	2
	T-8PSK2/9_UL	15	24	9	10	12
	T-QPSK1/3_U	40	0	12	8	2
	T-QPSK1/3_UL	10	24	9	10	12

The number of branches in the channel interleaver has been selected to be 48, using a constant memory size of roughly 4 Mbits for QPSK and 8 Mbits for 16QAM. All 3 configurations correspond to a single interleaving duration of 210 ms for the longest branch. The reason for selecting 48 branches was that it exhibited the best performance and the simplest adaptation to the other waveforms figures.

Clause 7.2.3.3.4 provides typical interleaver profiles and shows the delay introduced by the modulator in number of IUs for each branch of the transmitter interleaver. When Long Interleavers are used, their total delay covers several time slices. Regarding the processing in the receiver of a single burst of this time slice, several successive bursts are mapped in the various branches of the interleaver, as shown in figures A.1 and A.2. Figure 7.13 of clause 7.2.3.3.4 shows an interleaver spread over 12 slices; the Interleaver in the receiver will process the content of 12 consecutive bursts. Figure 7.14 of clause 7.2.3.3.4 shows an interleaver spread over 9 slices. The Interleaver in the receiver will process the content of 9 consecutive bursts. In figures A.1 and A.2, the delay in each path is computed modulo the number of IU between slices (indicated in the TPS bits by the slice_distance field), so that the figure folds each slice over the next one with different colours. The slices are longer than the number of indicated IUs and the number corresponding to one burst only is represented. Bursts in the consecutive slices are numbered (0 to 11), considering that burst 11 is the later one.

With this representation, it can also be seen that the effective "on time" of the receiver demodulating a given service should cover the duration of the all bursts.

This "on time" directly depends on the interleaver configuration, the mode setting for the trailing part of each burst, and on the maximum burst duration of the service, represented on the figures by the length of each bar.

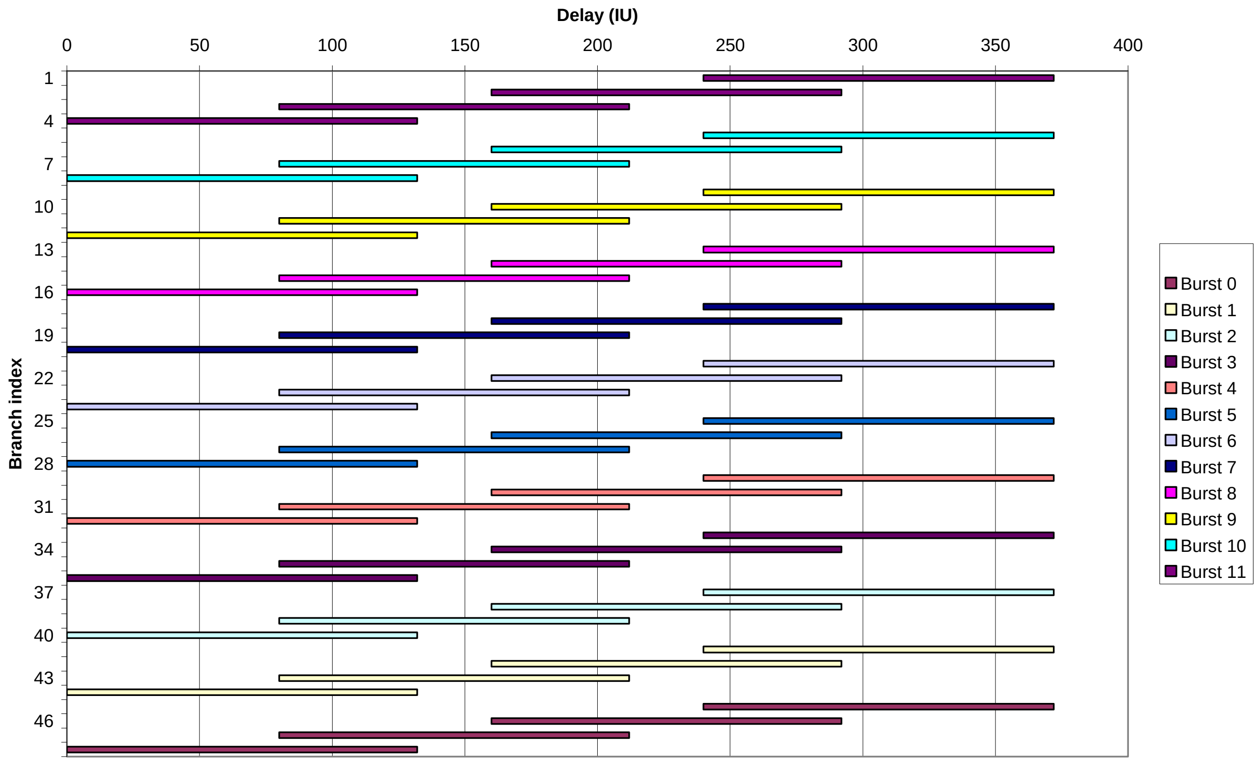


Figure A.1: Branches spreading for long-uniform QPSK interleaver

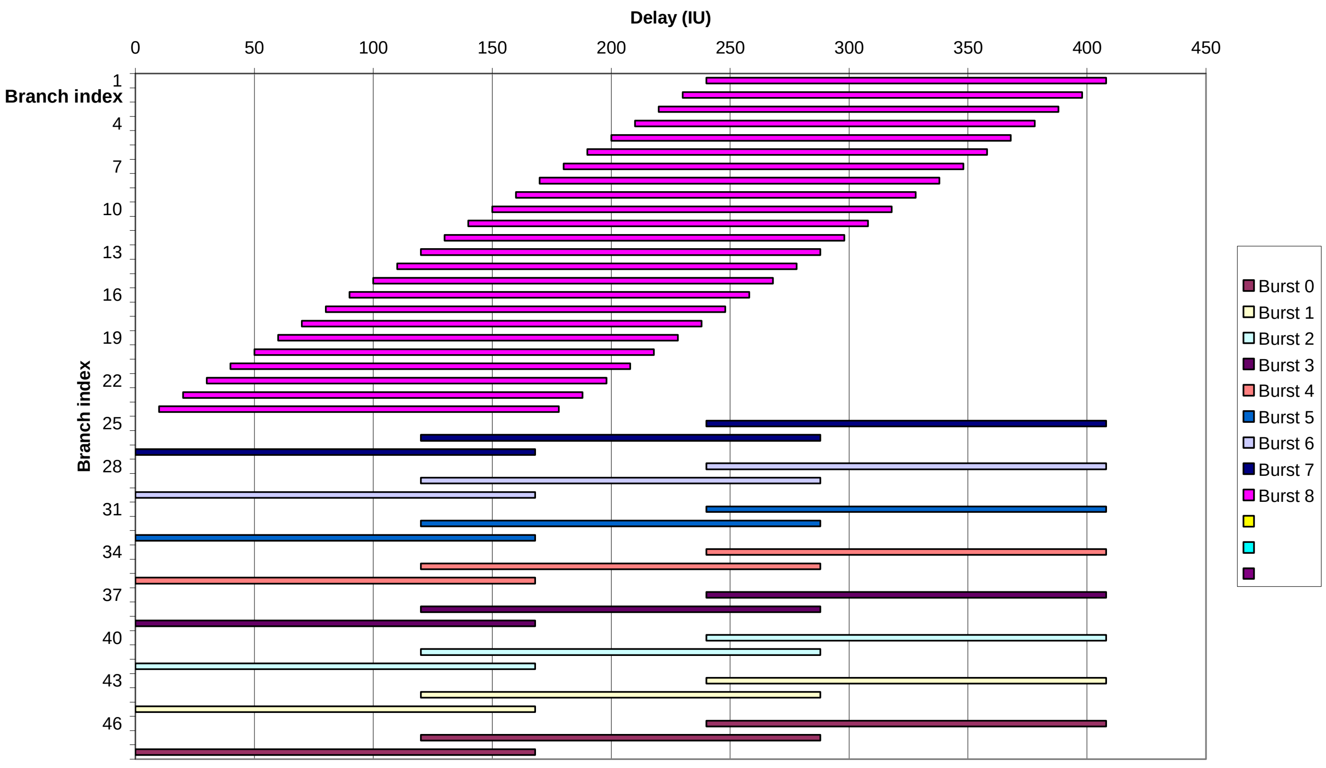


Figure A.2: Branches spreading for uniform-Late QPSK interleaver

A.5 Demodulator state machine principle

A.5.1 States representation of a demodulator

To emulate the demodulator behaviour in fading channels with deep and long fades, a state machine model was defined, based on the fact that demodulation takes place after channel estimation and that channel estimation is an iterative process that requires some time to converge. The overall behaviour is modelled with 4 states, representing the following effects:

- loss of demodulator synchronization because of received C/N falls below operational threshold. Consequently transitions to coasting mode and after some timeout to the re-acquisition mode;
- when the demodulator is operating below threshold, it feeds erased bits (neutral values) to the Turbo decoder;
- demodulator re-acquisition time;
- slicing impact on demodulator operation.

The demodulator threshold is represented with a single value, independently of the channel behaviour, while in real implementation, it will vary according to mobile channel condition (user speed, type of mobile channel).

In the model, the threshold margin is calculated as distance between the estimated C/N of the channel and the nominal demodulator threshold, both with Rayleigh, Rice channels. In the case of the 3-state LMS model, the margin is computed relative to the LOS receive C/N.

All receivers of a multiplexing frame (simulations use a fixed pattern of services distribution) are simulated and the states of the receivers are used to determine whether the LLR estimates of all bits in a received symbol must be erased. As indicated in figure A.3, it should be noted that all states have a transition to the idle state when the end of a burst occurs. While the green states are driven by the modulator behaviour, the red state represents the time slicing mechanism.

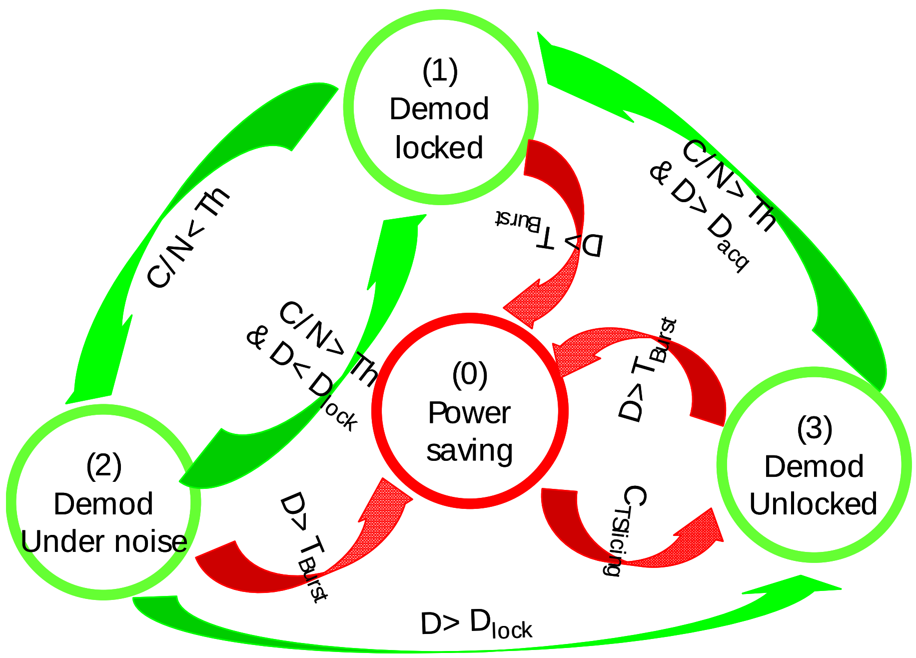


Figure A.3: State machine representing the demodulator process

The transitions parameters are:

- C/N : C/N estimated value at the receiver input;
- Th : receiver C/N Threshold (for staying locked not for successful frame decoding);

- D : duration of a state (origin of the transition);
- D_{lock} : maximum coasting state duration for the receiver operating under noise;
- T_{Burst} : burst duration (assumed constant);
- $C_{TSlicing}$: time interval between slices;
- D_{acq} : demodulator acquisition time (here assumed constant while in reality it will be a random variable).

The start transition stands for the condition required to wake up from the power saving mode. This is dependent on the time slicing mechanism. For example, in the current implementation of DVB-H the start condition ($C_{TSlicing}$) is as follows:

$$C_{TSlicing} = T>(\text{Delta\_T- Delta\_T\_Margin})$$

Where Delta_T stands for the start of the next burst as it is indicated by the header of all clauses of the burst. Delta_T_Margin is the preparation time including some margin for the jitter. It is assumed that Delta_T is known as long as the demodulator has been locked for a certain period of the previous burst. However, if the demodulator has been unlocked or with C/N below the minimum required for frames detection over the whole duration of the previous burst this Delta_T remains unknown. A mechanism to overcome this limitation must then be implemented (i.e. establish a fixed time between bursts).

In any case, the $C_{TSlicing}$ condition is set in accordance with the time slicing strategy. In this way, this method is also suitable for testing the time slicing performance.

In simulations, it is assumed that a fixed and regular time-slicing is used for all programs in a multiplex and therefore, receivers go from idle state to acquisition state at a time corresponding to the time-slice start minus D_{acq}.

To represent faithfully the demodulator estimator, it is necessary to consider that a given number of symbols are affected by the channel state to lead to a channel estimator state transition. This number of symbols is $N_{estimator}$.

A.5.1.1 IDLE state (0)

The idle state represents the power saving mode. The idle state is the starting mode of the state machine. It is also entered each time a time slice ends.

Transition to the unlocked state occurs when a start of slice is anticipated. The start of the slice information is exploited to switch on the demodulator in advance to allow fast acquisition. This time advance is supposed to be D_{acq}.

A.5.1.2 DEMOD unlocked state (3)

The demodulator is trying to acquire the carrier. Neutral value LLRs are fed to the Turbo decoder.

If the channel estimates (A), averaged over $N_{estimator}$ symbols goes below the threshold during the acquisition time, the acquisition is restarted until a correct margin is reached for the whole duration of the acquisition time.

When the channel estimates are above threshold over the whole D_{acq} time, the transition to the locked state occurs.

A.5.1.3 DEMOD locked state (1)

Locked state occurs when the C/N estimate is above the demodulator threshold Th. Actual LLRs are delivered to the decoder.

This state lasts until averaged channel estimates are above the required threshold margin. If the averaged estimated C/N falls below the margin, the demodulator switches to the coasting state.

A.5.1.4 DEMOD coasting state (2)

This state persists as long as the channel C/N estimates averaged over $N_{estimator}$ symbols are under the demodulator threshold Th. Neutral values of LLR are fed to the Turbo decoder.

Each time the demodulator enters this state, a timer starts for a duration of D_{lock}. If the estimated demodulator C/N averaged over $N_{estimator}$ symbols remains under the threshold Th longer then D_{lock}, then the demodulator moves to the acquisition mode (3). Instead if after D_{lock} the averaged C/N estimate is above the demodulator threshold Th, then a transition to the locked mode (1) occurs and the timer is reset.

A.5.2 OFDM simulations

The demodulator state machine model must be activated to provide results on the LMS channel in OFDM mode. It accounts for the degradations introduced by the non-instantaneous channel synchronization in the receiver. As far as channel estimation and symbol phase synchronization are concerned, the ideal channel estimator remains, providing ideal phase and amplitude recovery to any demodulated symbol.

The demodulator state machine parameters are recalled hereafter.

Table A.9: Demodulator state machine parameters definitions

Parameters	Definition
Bandwidth	Bandwidth of the OFDM signal, used to compute the number of subcarriers per symbol.
Mode	OFDM mode value indicating the symbol size (1K, 2K, 4K, 8K).
Constellation	Number of bits per subcarrier. Because input estimates are assumed to be on one subcarrier and output is supposed to correspond to LLR (bit by bit), the state machine generates constellation x number of subcarriers states per channel sample.
Punc_Pattern_ID	Puncturing pattern ID providing the coding rate.
TPS_CommonMultiplier TPS_NLateTaps TPS_NSlices TPS_SliceDistance TPS_NonLateIncrement	TPS parameters specifying the interleaver configuration..
MuxFrameProfile	Profile of the repetition pattern where the services are repeated on a regular pattern, each service being allocated a fixed number of code words during that period. The number of services is deduced from that profile.
ReceiverMarginThreshold_Db	Margin between LOS channel C/N level and demodulator threshold in dB. C/N=-3,5 dB is selected as the receiver threshold value.
LockDurationInSymbols	Number of OFDM symbols during which the demodulator remains locked while the signal averaged values are below threshold. This gives the duration of the "coasting" mode. 68 symbols is selected for this duration value.
AcquisitionDurationInSymbols	Number of OFDM symbols required to switch to locked mode after the demodulator is started or after it has left the locked mode. 68 symbols is selected for this duration value.
LengthOfAveragingInSymbols	Number of OFDM symbols over which the demodulator is supposed to average the received channel estimates to determine its own state. 12 symbols is selected for this duration value.

The set of demodulator state machine values corresponds to typical performance of existing DVB-H receivers.

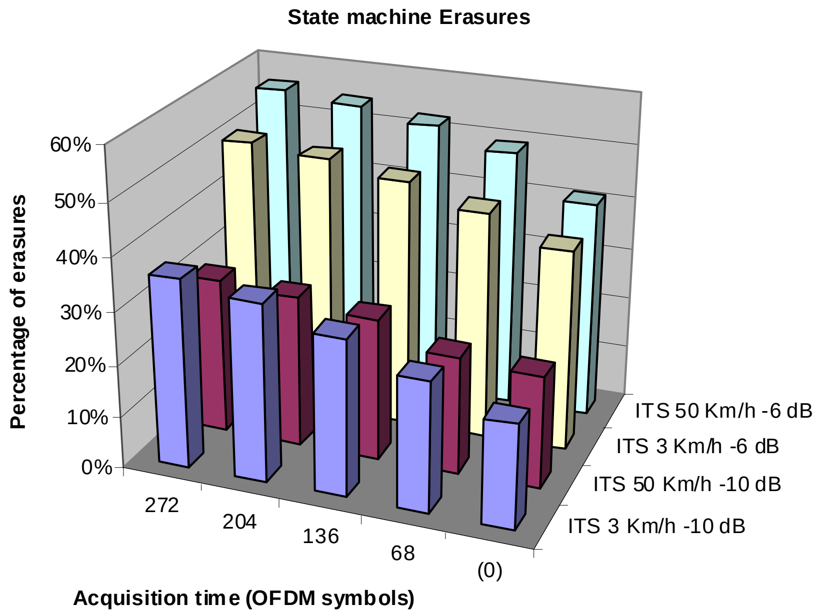
Simulation results without activation of the state machine are also provided as a benchmark.

A.5.3 TDM simulations

The same state machine model and parameters are used TDM simulations. The rationale is that, like for OFDM case, the TDM demodulators will rely on the "data aided" information from the pilots that are present in the TDM frame (equivalent to an OFDM symbol). Therefore, the acquisition and channel estimation algorithms will provide similar performances. This hypothesis is possibly slightly pessimistic for TDM which do not require the two dimensional (time/frequency) channel estimation of OFDM.

A.5.4 Analysis of acquisition time for LMS channels

The statistical analysis uses the symbols generated by the LMS channel model and feeds them into the receivers state machines model. Then, the number of occurrences of states are counted.



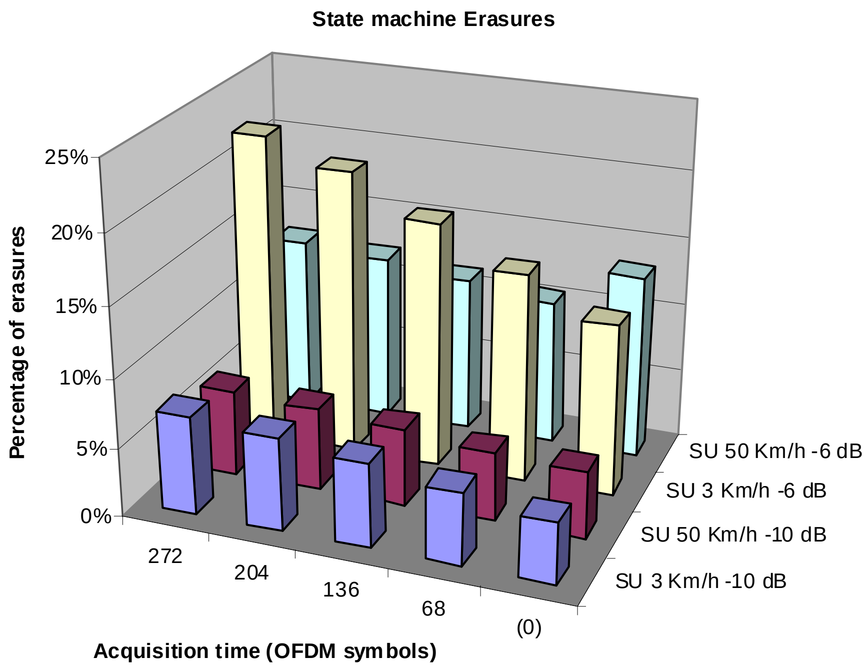


Figure A.4: Erasure rate for ITS and suburban channels

The charts show that the LMS-ITS channel generates much more erasures than the LMS-SU. Also, having the acquisition time in the order of 68 OFDM symbols (one DVB-SH frame) does not increase significantly the percentage of erasures with respect to the ideal case (instantaneous acquisition).

Note that those results also depend on the averaging duration which can even decrease the number of erasures compared to the ideal acquisition case.

A.6 MPE-IFEC Simulation conditions

A.6.1 Definitions and interfaces

The following interfaces are defined and represented in figure A.5:

- on sender side:
  - E1 represents the IP flow coming into the IP encapsulator; these IP flows are of variable bit rates since they represent individual services;
  - E2 represents the TS interface at the output of the IP encapsulator; the TS bit rate is fixed and is a function of the waveform parameters;
  - E3 is the output of the DVB-SH modulator;
- on the receiver side:
  - R3 is the reception of the DVB-SH signal, at the input of the SH baseband decoder;
  - R2 is the output of the SH baseband receiver delivering a TS stream at continuous bit rate after all physical deinterleaving and decoding processes;
  - R1 is the output of the SH link layer receiver delivering IP flows at variable bit rates after all MPE-IFEC deinterleaving and decoding processes;
  - finally R0 is the output of the H264 video buffer ready for rendering.

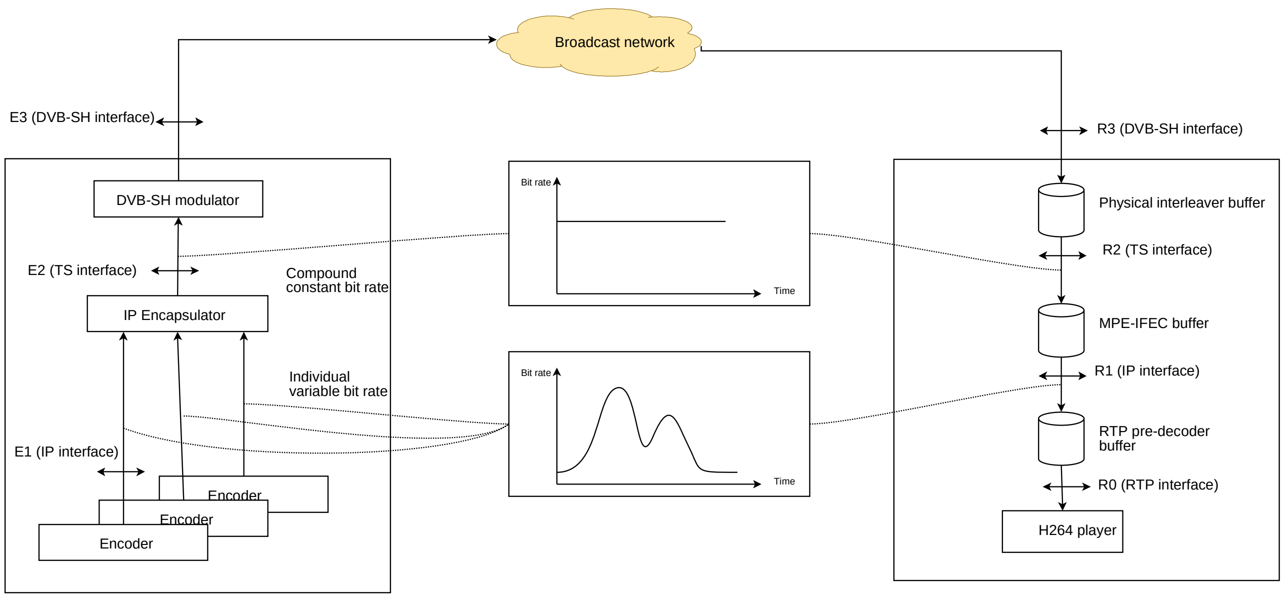


Figure A.5: receiver interfaces

All simulation results have been measured at interface R1.

Some of the processes in the receiver may introduce jitter during transitions like zapping. The following performance results have been produced assuming a jitter-free interface, so assuming a constant delay in the physical and link layer buffers. Performance during transitions like zapping at non-jitter free interfaces are not scope of this implementation guidelines release. Performance including RTP buffer are found in the IPDC over DVB-SH implementation guidelines. However some aspects of zapping time improvements techniques, without performance figures, can be found in clause 6.2.5.2.

A.6.2 Decoder architecture

A simple decoder architecture is presented in figure A.6. It implements the following features:

- event driven approach:
  - time outs for the reception of a time slice burst with simple timing heuristics based on delta-t and max_burst_durations;
  - timing are based on ADST: one ADST is output at a deterministic time; this enables to output an ADST even in the case of a "starvation" due to a long erasure event;
- support of erased bursts detection:
  - when erased bursts are detected, they are sequentially mapped over the ADT to avoid having ADT with outdated columns when decoding is requested;
- "brute force approach" for the decoding:
  - the decoding is initiated every time a new information concerning the Encoding Matrix (ADT and FDT) is detected within a specific range;
  - this enables to output the ADST based on output time without requiring any new decoding since, when the ADST needs to be output, all the relevant information has already been processed by previous "brute force decoding"; the ADST is the best one with received information;
  - this enables also to support the decoding in different configurations by profiling the range of EM to be decoded; one typical configuration is to decode only at the jitter-free interface (no more information will be received so it is time to decode); another one is to decode by anticipation, especially during the zapping period without having received all parity;
  - this also enables to support a D>0;
- typical configurations:
  - anticipated decoding:
    - Min_ADT = Min(ifdt_index(k,S-1); adt_index(k-D, 0));
    - Max_ADT = Min(ifdt_index(k,0); adt_index(k-D,B-1));
  - without anticipated decoding:
    - Min_ADT = Min(ifdt_index(k,S-1); adt_index(k-D, 0));
    - Max_ADT = Min(ifdt_index(k,0); adt_index(k-D,0)).

This decoder can be used for providing performance at the jitter free interface by using the non anticipated decoding range and D=0.

- Min_ADT = Min(ifdt_index(k,S-1); adt_index(k, 0)) [M] = Min (($k - S$– 2 + M)[M], k[M])
- Max_ADT = Min(ifdt_index(k,0); adt_index(k,0)) [M] = Min ((k-1+M)[M], k[M])

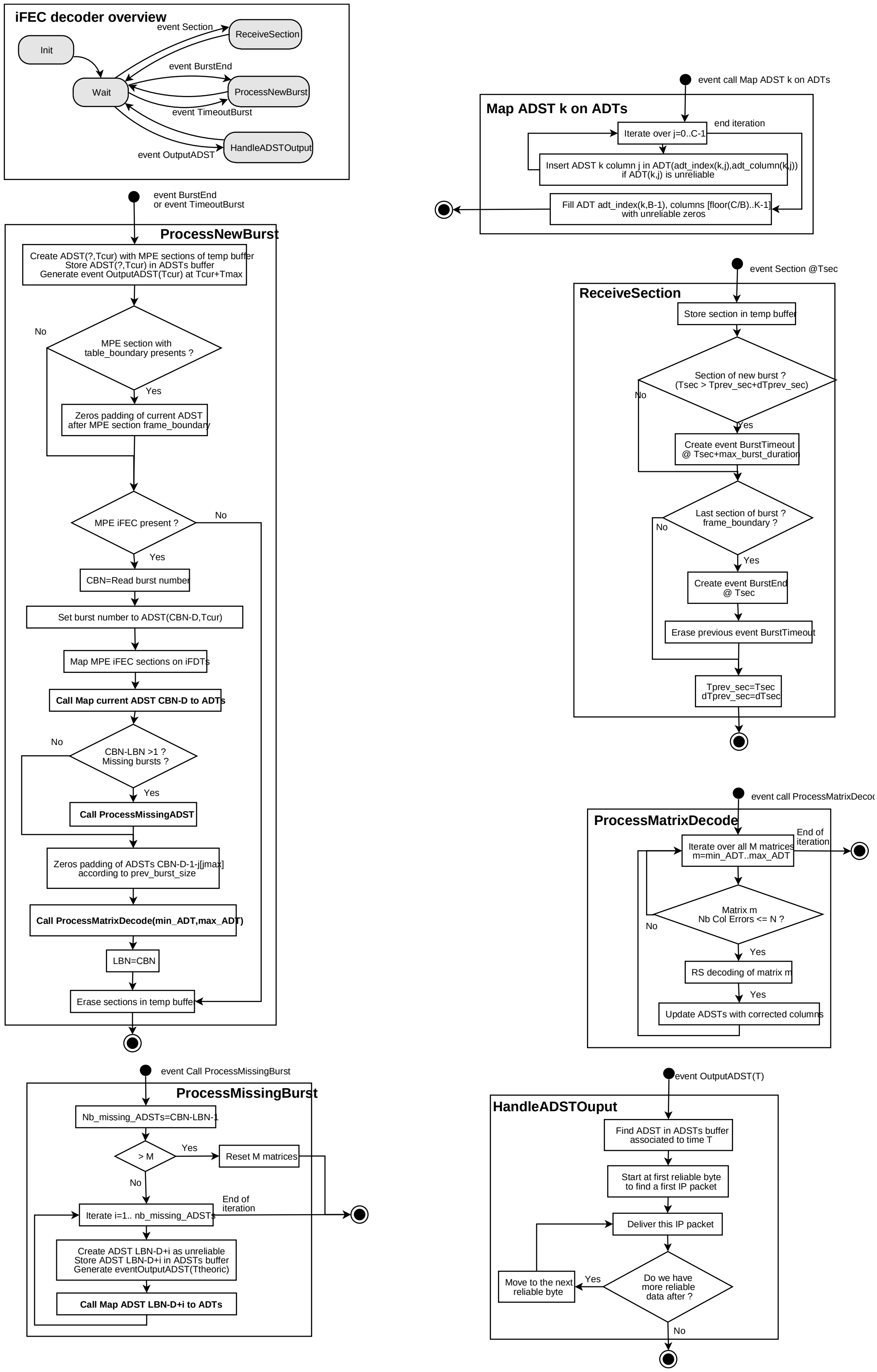


Figure A.6: simple MPE-IFEC decoder architecture

A.6.3 Other parameters

The following parameters and options have been sought during simulation:

- IP size: set to 1 000 bytes;
- encapsulation: MPE;
- section packing: on;
- errors processing: when a TS of a section is lost: the full MPE section is considered as lost;
- the MPE-IFEC makes usage of TS level dumps which have a constant size at TP level; this fixed number of TP gives a capacity that is used to convey the IP flows using MPE sections and the RS parity using the MPE-IFEC. If the TS capacity is not sufficient to transport an IP packet or an MPE-IFEC section, the corresponding TS are set to null and not used.

A.7 Propagation channels

A.7.1 Clarification on Rice and Rayleigh channels

It should be remarked that Rayleigh and Rice channel models definitions are also used in the context of DVB-T/H but they refer to two different contexts:

- The stationary Rice/Rayleigh channel models defined in EN 300 744 [28], clause B. The models represent **stationary** reception with antennas on the roof of a house (model "F1") or reception of signal from very high power repeater (as typical for DVB-T networks) with small antennas (model "P1"). The models are also (improperly) called RICE and RAYLEIGH in the context of DVB-T/H measurement. In reality the models represent a particular realization of TDL with stochastic fading process and TDL complex taps are fixed. These models play a minor role for the evaluation of DVB-SH and have been only used in an intermediate phase of the project;
- The uncorrelated Rice/Rayleigh channel models used in clauses 7.2.2.6.2 and 7.2.2.6.3 of the present document. The performance of the system depends highly on the speed and the interleaver length. For the evaluation of the FEC decoding a simple channel model representing an ideal interleaving (upper bound in performance compared to finite length interleaving) has been selected. A symbol-by-symbol independent Rayleigh amplitude sample distribution is used to represent typical terrestrial FEC operating conditions. In a similar way a symbol-by-symbol independent Rice amplitude sample distribution is used to represent typical satellite FEC operating conditions.

A.7.2 Propagation channels

The following propagation channels have been selected for simulation.

Table A.10

	Terrestrial	Satellite
URBAN	TU6 (3 kmph) TU6 (50 kmph)	
SUBURBAN		LMS-SU (50 kmph) LMS-SU (3 kmph)
RURAL		LMS-ITS (50 kmph) Quasi-stationary Ricean, K=5 dB; 7 dB;10 dB

The C/N and C/I levels to be used are derived from link budgets (see clause 11), with terminal characteristics given in clause 10. The used values are listed in clause A.9.

Terrestrial simulations use the TU6 propagation model, defined in clause C of TS 145.005 [29]. For the terrestrial channel the state machine is not activated as it is assumed that the receiver is always in lock. Indeed, only the short term fading (TU6 model) is simulated, not the macro attenuation due to user large displacement within the cell (lognormal shadowing). For this channel, 10 to 20 minutes duration were used for statistics.

For the LMS channels the simulations were performed with and without receiver state machine. The parameters for the LMS channels are obtained from [16] and [17] and are specified in table A.21. *The applicability of the selected values is not claimed to be universal but only as a starting point to differentiate between waveform configurations.* For a discussion on the applicability of such models, refer to clause 4. From experiments, it was found that the simulation length when the 3-state Markov chain is activated should be at least 1 hour of simulated time for 50 kmph and at least 3 hours for 3 kmph.

Table A.11: LMS model states based on measurements parameters for 40° elevation

Environment	State 1: LOS			State 2: Shadowing			State 3: Deep shadow		
	α (dB)	ψ (dB)	MP (dB)	α (dB)	ψ (dB)	MP (dB)	α (dB)	ψ (dB)	MP (dB)
Open (*)	0,1	0,37	-22,0	-1,0	0,5	-22,0	-2,25	0,13	-21,2
Suburban	-1,0	0,5	-13,0	-3,7	0,98	-12,2	-15,0	5,9	-13,0
Intermediate Tree-Shadow	-0,4	1,5	-13,2	-8,2	3,9	-12,7	-17,0	3,14	-10,0
Heavy Tree-Shadow (2)	-	-	-	-10,1	2,25	-10,0	-19,0	4,0	-10,0

Table A.12

Environment	[P]			[W]	d_{corr} (m)	L_{frame} (m)	L_{trans} (m)
Open (2)	0,9530	0,0431	0,0039	0,5	2,5	8,9	12,4
	0,0515	0,9347	0,0138	0,375		7,5	
	0,0334	0,0238	0,9428	0,125		4,0 (1)	
Suburban	0,8177	0,1715	0,0108	0,4545	1,7	5,2	2,2
	0,1544	0,7997	0,0459	0,4545		3,7	
	0,1400	0,1433	0,7167	0,091		3,0 (1)	
Intermediate Tree-Shadow	0,7193	0,1865	0,0942	0,3929	1,5	6,3	2,6
	0,1848	0,7269	0,0883	0,3571		6,3	
	0,1771	0,0971	0,7258	0,25		4,5	
Heavy Tree-Shadow (2)	0,7792	0,0452	0,1756	0	1,7	-	3,5
	0	0,9259	0,0741	0,5		4,8	
	0	0,0741	0,9259	0,5		4,5	
NOTE 1: These values have been extrapolated since they are not given in [17]. NOTE 2: Not simulated, for information only.							

The parameters values are:

- α : average value of the attenuation on the LOS link for a state;
- ψ : standard deviation of the attenuation on the LOS link for a state;
- [P] : probability of occurrence of a transition (3x3 matrix);
- [W] : total probability of having a given state;
- dcorr: correlation distance of the channel;
- L_{Frame}: minimum state frame length as defined in [17];
- L_{Trans}: transition region length as defined in [16].

A.7.3 Satellite quasi stationary channel

For portable satellite reception it was considered important to also analyse the case of LOS with multipath but very slow user movement (e.g. 0,5 kmph or less). This channel is called quasi-stationary satellite. The results of this analysis are exploited in clause 11.5.3.1. The channel is modelled as AWGN over the FEC block size with a block-by-block slight C/N change due to the very slow fading. Hence, we can analytically compute for a given LOS C/N the probability to be below the demodulator threshold (see clause 11).

Recalling the PDF of the power for a Ricean faded signal:

$$p_D(d) = \frac{1}{2\sigma^2} \exp\left\{ -\left(\frac{s^2 + d}{2\sigma^2} \right) \right\} I_o\left(\frac{s\sqrt{d}}{\sigma^2} \right)$$

where *d* is the LOS signal power;

σ^2 the variance of the in-phase and quadrature multipath fading components; and

$I_o(\cdot)$ is the modified zero-th order Bessel function.

The carrier to multipath ratio is defined as the ratio between the power of the signal complex envelope LOS component *s* and the power of the multipath (random) power component $2\sigma^2$:

$$K = \frac{s^2}{2\sigma^2}.$$

A.8 Evaluation criteria

In case of long time interleaving and block fading, BER and WER alone are not representative of actual perceived quality. A long mute may be better than repetitive short error events. To take this consideration into account the following criteria are considered relevant. The ESR(5) is used, whenever available, as the planning criterion.

Table A.13: Evaluation criteria definitions

Parameter	Description	Remarks
BER	average bit error rate.	
WER	Word Error Rate. A word is a FEC decoded word.	At 2 Mbps, 1E4 FEC blocks correspond roughly to 1,5 min. Estimates to be provided over, at least 25 errored word or 1E4 simulated word
PER	MPEG Packet error rate	8 packets per FEC word
FER(x)	FER: Frame Error Rate, corresponding to the ratio of erroneous Frame including at least one erroneous bit to the total number of Frames in the observation period. A Frame is an MPE Frame, without any IFEC or MPE-FEC error correction being applied. FER(x): Frame Error Rate of x %.	Remark: at MPEG TS interface, a Frame may be defined in number of MPEG-TS packets.
MFER(x)	MFER(x): MPE-Frame Error Rate of x %	An MPE-Frame is an MPE Frame, following IFEC and MPE-FEC (if any) decoding
"ESR(5) fulfillment" or ESR(5) for short	ESR(5) is the ratio of windows for which ESR5 is fulfilled, over the total number of windows	See equation below
ESR5	ESR5 criterion is fulfilled when in a time interval of 20 s there is at most one second in error.	
EFSR	Error free seconds ratio A second may be defined as a fixed number of consecutive Words in function of the service bit rate.	

$$ESR(5)\,fulfillment\_ratio = 1 - \frac{\sum_{window=1}^{window=N} ceil\left(\frac{\max(0; nof\_erroneous\_\sec onds(window) - 5 * window\_size)}{window\_size}\right)}{N}$$

A.9 Summary of simulated cases

Table A.14 hereafter provides all the simulation cases with the associated parameters. The satellite C/N and C/I are in line with the link budgets detailed in clause 11 (rationale with link-budgets is given at the end of this clause). In table A.15 for the satellite case C/N represents the LOS C/N (no channel fading/shadowing) while C/I represents the co-channel useful signal over interference and uplink C/N contribution.

Note that satellite link C/I and C/N are provided separately as the interference is suppose to fade together with the signal and is simulated like an extra noise source modulated by the same fade as the useful signal.

Table A.14: List of simulated cases

Waveform configuration	Channel	Speed	State-machine	C/I	C/N	ID	Comment
O-16QAM1o4_S	TU6	50 kmph	off	20 dB	[2:6] dB	1	Basic terrestrial-only link
O-16QAM1o4_S	TU6	3 kmph	off	20 dB	[2:6] dB	2	Basic terrestrial-only link
O-16QAM1o5_UL	TU6	50 kmph	off	20 dB	[1:3] dB	3	Check Long interleaving on TU6
O-16QAM1o5_UL	TU6	3 kmph	off	20 dB	[2:6] dB	4	Check Long interleaving on TU6
O-16QAM2o7_S	TU6	50 kmph	off	20 dB	[2:6] dB	5	Basic terrestrial-only link
O-16QAM2o7_S	TU6	3 kmph	off	20 dB	[2:6] dB	6	Basic terrestrial-only link
O-QPSK1o2_S	TU6	50 kmph	off	20 dB	[3:5] dB	7	Basic terrestrial-only link
O-QPSK1o2_S	TU6	3 kmph	off	20 dB	[4:6] dB	8	Basic terrestrial-only link
O-QPSK1o3_UL	TU6	50 kmph	off	20 dB	[0:2] dB	9	Check Long interleaving on TU6
O-QPSK1o3_UL	TU6	3 kmph	off	20 dB	[1:5] dB	10	Check Long interleaving on TU6
O-QPSK1o3_UL	TU6	3 kmph	on	20 dB	[1:5] dB	11	Verify contribution from state machine
O-QPSK2o3_S	TU6	50 kmph	off	20 dB	[5:7] dB	12	Basic terrestrial-only link
O-QPSK2o3_S	TU6	3 kmph	off	20 dB	[6:10] dB	13	Basic terrestrial-only link
O-16QAM1o4_S	LMS-ITS	50 kmph	on	11,5 dB	10,8 dB	14	Vehicular
O-16QAM1o5_U	LMS-ITS	50 kmph	on	11,5 dB	10,8 dB	15	Vehicular
O-16QAM1o5_UL	LMS-ITS	50 kmph	on	11,5 dB	10,8 dB	16	Vehicular
O-16QAM2o7_S	LMS-ITS	50 kmph	on	11,5 dB	10,8 dB	17	Vehicular
O-QPSK1o2_S	LMS-ITS	50 kmph	on	11,9 dB	11,2 dB	18	Vehicular
O-QPSK1o3_U	LMS-ITS	50 kmph	on	11,9 dB	11,2 dB	19	Vehicular
O-QPSK1o3_UL	LMS-ITS	50 kmph	on	11,9 dB	11,2 dB	20	Vehicular
T-8PSK1o3_S	LMS-ITS	50 kmph	On	12,0 dB	11,8 dB	21	Vehicular
T-8PSK2o9_U	LMS-ITS	50 kmph	On	12,0 dB	11,8 dB	22	Vehicular
T-8PSK2o9_UL	LMS-ITS	50 kmph	On	12,0 dB	11,8 dB	23	Vehicular
T-QPSK1o2_S	LMS-ITS	50 kmph	on	12,5 dB	12,3 dB	24	Vehicular
T-QPSK1o3_U	LMS-ITS	50 kmph	on	12,5 dB	12,3 dB	25	Vehicular
T-QPSK1o3_UL	LMS-ITS	50 kmph	on	12,5 dB	12,3 dB	26	Vehicular
O-16QAM1o4_S	LMS-SU	50 kmph	on	11,5 dB	10,8 dB	27	Vehicular
O-16QAM1o5_U	LMS-SU	50 kmph	on	11,5 dB	10,8 dB	28	Vehicular
O-16QAM1o5_UL	LMS-SU	50 kmph	on	11,5 dB	10,8 dB	29	Vehicular
O-16QAM2o7_S	LMS-SU	50 kmph	on	11,5 dB	10,8 dB	30	Vehicular
O-QPSK1o2_S	LMS-SU	50 kmph	on	11,9 dB	11,2 dB	31	Vehicular
O-QPSK1o3_U	LMS-SU	50 kmph	on	11,9 dB	11,2 dB	32	Vehicular
O-QPSK1o3_UL	LMS-SU	50 kmph	on	11,9 dB	11,2 dB	33	Vehicular
T-8PSK1o3_S	LMS-SU	50 kmph	on	12,0 dB	11,8 dB	34	Vehicular
T-8PSK2o9_U	LMS-SU	50 kmph	on	12,0 dB	11,8 dB	35	Vehicular
T-8PSK2o9_UL	LMS-SU	50 kmph	on	12,0 dB	11,8 dB	36	Vehicular
T-QPSK1o2_S	LMS-SU	50 kmph	on	12,5 dB	12,3 dB	37	Vehicular
T-QPSK1o3_U	LMS-SU	50 kmph	on	12,5 dB	12,3 dB	38	Vehicular
T-QPSK1o3_UL	LMS-SU	50 kmph	on	12,5 dB	12,3 dB	39	Vehicular
O-16QAM1o5_U	LMS-SU	3 kmph	on	11,5 dB	4,7 dB	72	Vehicular
O-16QAM1o5_UL	LMS-SU	3 kmph	on	11,5 dB	4,7 dB	73	Vehicular
O-QPSK1o2_S	LMS-SU	3 kmph	on	11,9 dB	5,1 dB	74	Vehicular
O-QPSK1o3_U	LMS-SU	3 kmph	on	11,9 dB	5,1 dB	75	Vehicular
O-QPSK1o3_UL	LMS-SU	3 kmph	on	11,9 dB	5,1 dB	76	Vehicular
T-8PSK1o3_S	LMS-SU	3 kmph	on	12,0 dB	5,7 dB	77	Vehicular
T-8PSK2o9_U	LMS-SU	3 kmph	on	12,0 dB	5,7 dB	78	Vehicular
T-8PSK2o9_UL	LMS-SU	3 kmph	on	12,0 dB	5,7 dB	79	Vehicular
T-QPSK1o2_S	LMS-SU	3 kmph	on	12,5 dB	6,2 dB	80	Vehicular
T-QPSK1o3_U	LMS-SU	3 kmph	on	12,5 dB	6,2 dB	81	Vehicular
T-QPSK1o3_UL	LMS-SU	3 kmph	on	12,5 dB	6,2 dB	82	Vehicular

Table A.15 shows the number of EFRAMES (or code words) for each burst of the time slicing. Those values correspond to a fix allocation of code words per service on a regular pattern basis.

Table A.15: Time-slicing configurations for the different OFDM system cases

OFDM Case	Number of Services	Service Number								
		0	1	2	3	4	5	6	7	8
16QAM_1/4_S	8	34	34	34	34	34	34	34	34	
16QAM_2/7_S	8	38	38	38	38	38	38	38	38	
QPSK1/2_S	8	34	34	34	34	34	34	34	34	
QPSK2/3_S	8	45	45	45	45	45	45	45	45	
16QAM_1/5_U	9	24	24	24	24	24	24	24	24	24
16QAM_1/5_UL	9	30	30	30	30	30	30	30	30	30
QPSK1/3_U	8	22	23	22	23	22	23	22	23	
QPSK_1/3_UL	8	28	28	28	28	28	28	28	29	

Table A.16: Time-slicing configurations for the different TDM cases

TDM Case	Number of Services	Service Number								
		0	1	2	3	4	5	6	7	8
8PSK_1/3_S	9	35	35	35	35	35	35	35	35	36
QPSK1/2_S	9	34	35	35	34	35	35	34	35	35
8PSK_2/9_U	9	23	23	23	23	23	23	23	23	24
8PSK_2/9_UL	9	28	29	29	29	29	29	29	29	29
QPSK1/3_U	9	23	23	23	23	23	23	23	23	24
QPSK_1/3_UL	9	28	29	29	29	29	29	29	29	29

Rationale between values used in Simulation and values given in clause 11 link-budgets

C/N are provided in link bugets from clause A.11.

As LMS simulations are lossless, C/N provided in link budgets are reduced from the amount of implementations losses identified in clause 10 (as well as in link budgets of clause 11).

Thus the following C/N are used in the LMS simulations:

Vehicular terminal:

Table A.17

Waveform	Sh-A	Sh-A	Sh-B	Sh-B
Modulation	QPSK	16QAM	QPSK	8PSK
Link budget LOS C/N (dB)	12,3	12,3	12,8	12,8
Implementation losses (dB)	1,1	1,5	0,5	1
C/N for simulations	11,2	10,8	12,3	11,8

Handheld (2b) terminal:

Table A.18

Waveform	Sh-A	Sh-A	Sh-B	Sh-B
Modulation	QPSK	16QAM	QPSK	8PSK
Link budget LOS C/N (dB)	6,2	6,2	6,7(see note)	6,7(see note)
Implementation losses (dB)	1,1	1,5	0,5	1
C/N for simulations	5,1	4,7	6,2	5,7
NOTE: Link budget for SH-B and handheld terminal is not provided in clause 11. However C/N may be estimated by similarly to vehicular terminal, where SH-B C/N is 0,5 dB over SH-A C/N.				

The same methodology is used for the calculation of the C/I. In link budgets, two C/I contributions are identified. Thus the following C/I are used in the LMS simulations:

Table A.19

Waveform	Sh-A	Sh-A	Sh-B	Sh-B
Modulation	QPSK	16QAM	QPSK	8PSK
Link budget uplink C/I (dB)	19,5	19,5	20	20
Link budget Satellite C/I (dB)	14	14	14	14
Link budget total C/I (dB)	13	13	13	13
Implementation losses (dB)	1,1	1,5	0,5	1
C/I for simulations	11,9	11,5	12,5	12

A.10 Detailed interleaver profiles

This clause provides the detailed profiles used for each interleaver use case, with the associated memory requirements.

A.10.1 Class 1 cases

These interleaver profiles are used in the case of short interleaving, which corresponds to the application of DVB-SH in terrestrial usage or the application in satellite usage with an additional FEC and a long interleaving applied at an upper layer.

Table A.20: Interleaver profile class1 cases

Interleaver Profile	Interleaver delay table for taps L[0] to L[47]																
QPSK1/2_S **QPSK2/3_S** **T-QPSK1/2_S** (5 / 48 / 1 / 0 / 0) (see note)	0	5	10	15	20	25	30	35	40	45	50	55	60	65	70	75	80
	85	90	95	100	105	110	115	120	125	130	135	140	145	150	155	160	165
	170	175	180	185	190	195	200	205	510	215	220	225	230	235			
T-8PSK1/3_S (8 / 48 / 1 / 0 / 0)(*)	0	8	16	24	32	40	48	56	64	72	80	88	96	104	112	120	128
	136	144	152	160	168	176	184	192	200	208	216	224	232	240	248	256	264
	272	280	288	296	304	312	320	328	336	344	352	360	368	376			
16QAM_1/4_S **16QAM_2/7_S** (10 / 48 / 1 / 0 / 0) (see note)	0	10	20	30	40	50	60	70	80	90	100	110	120	130	140	150	160
	170	180	190	200	210	220	230	240	250	260	270	280	290	300	310	320	330
	340	350	360	370	380	390	400	410	420	430	440	450	460	470			
NOTE: (common_multiplier / nof_late_taps / nof_slices / slice_distance / non_late_increment)																	

A.10.2 Uniform long interleaver - Class 2 cases

Profile used for cases where long physical layer interleaver is considered (U cases), without late part.

Table A.21: Uniform long interleaver profile - Satellite cases

Interleaver Profile	Interleaver delay table for taps L[0] to L[47]								
16QAM_1/5_U	0	160	320	480	2 176	2 336	2 496	2 656	4 352
(40 / 0 / 12 / 8 / 4)	4 512	4 672	4 832	6 528	6 688	6 848	7 008	8 704	8 864
(see note)	9 024	9 184	10 880	11 040	11 200	11 360	13 056	13 216	13 376
	13 536	15 232	15 392	15 552	15 712	17 408	17 568	17 728	17 888
	19 584	19 744	19 904	20 064	21 760	21 920	22 080	22 240	23 936
	24 096	24 256	24 416						
QPSK1/3_U	0	80	160	240	1 088	1 168	1 248	1 328	2 176
(40 / 0 / 12 / 4 / 2)	2 256	2 336	2 416	3 264	3 344	3 424	3 504	4 352	4 432
(see note)	4 512	4 592	5 440	5 520	5 600	5 680	6 528	6 608	6 688
	6 768	7 616	7 696	7 776	7 856	8 704	8 784	8 864	8 944
	9 792	9 872	9 952	10 032	10 880	10 960	11 040	11 120	11 968
	12 048	12 128	12 208						
T-8PSK2/9_U	0	120	240	360	1 872	1 992	2 112	2 232	3 744
(60 / 0 / 12 / 4 / 2)	3 864	3 984	4 104	5 616	5 736	5 856	5 976	7 488	7 608
(see note)	7 728	7 848	9 360	9 480	9 600	9 720	11 232	11 352	11 472
	11 592	13 104	13 224	13 344	13 464	14 976	15 096	15 216	15 336
	16 848	16 968	17 088	17 208	18 720	18 840	18 960	19 080	20 592
	20 712	20 832	20 952						
T-QPSK1/3_U	0	80	160	240	1 248	1 328	1 408	1 488	2 496
(40 / 0 / 12 / 4 / 2)	2 576	2 656	2 736	3 744	3 824	3 904	3 984	4 992	5 072
(see note)	5 152	5 232	6 240	6 320	6 400	6 480	7 488	7 568	7 648
	7 728	8 736	8 816	8 896	8 976	9 984	10 064	10 144	10 224
	11 232	11 312	11 392	11 472	12 480	12 560	12 640	12 720	13 728
	13 808	13 888	13 968						
NOTE: (common_multiplier / nof_late_taps / nof_slices / slice_distance / non_late_increment).									

A.10.3 Uniform Late interleaver - Class 2 cases

The interleaver configurations are split into a late part which fits into the on-chip memory limited to 8 Mbits and a uniform part which fits into the external 512 Mbits memory.

The late part is also the part which is usable for class 1 receivers which cannot store the long-dispersed redundancy. This part corresponds to 1/2 of the 48 branches (24). The corresponding code rate is therefore multiplied by 2, e.g. rate 2/3 for system cases with QPSK 1/3 and rate 2/5 for system cases with 16QAM 1/5.

Table A.22: Late part for Uniform Late profile - Satellite cases

Interleaver Profile	(common_multiplier / nof_late_taps / nof_slices / slice_distance / non_late_increment)								
16QAM_1/5_UL	0	20	40	60	80	100	120	140	160
(20 / 24 / 9 / 10 / 12)(*)	180	200	220	240	260	280	300	320	340
	360	380	400	420	440	460	2 720	2 960	3 200
	5 440	5 680	5 920	8 160	8 400	8 640	10 880	11 120	11 360
	13 600	13 840	14 080	16 320	16 560	16 800	19 040	19 280	19 520
	21 760	22 000	22 240						
QPSK_1/3_UL	0	10	20	30	40	50	60	70	80
(10 / 24 / 9 / 5 / 12)(*)	90	100	110	120	130	140	150	160	170
	180	190	200	210	220	230	1 360	1 480	1 600
	2 720	2 840	2 960	4 080	4 200	4 320	5 440	5 560	5 680
	6 800	6 920	7 040	8 160	8 280	8 400	9 520	9 640	9 760
	10 880	11 000	11 120						
T-8PSK2/9_UL	0	15	30	45	60	75	90	105	120
(15 / 24 / 9 / 5 / 12)(*)	35	150	165	180	195	210	225	240	255
	270	285	300	315	330	345	2 340	2 520	2 700
	4 680	4 860	5 040	7 020	7 200	7 380	9 360	9 540	9 720
	1 700	11 880	12 060	14 040	14 220	14 400	16 380	16 560	16 740
	18 720	18 900	19 080						
T-QPSK1/3_UL	0	10	20	30	40	50	60	70	80
(10 / 24 / 9 / 5 / 12)(*)	90	100	110	120	130	140	150	160	170
	180	190	200	210	220	230	1 560	1 680	1 800
	3 120	3 240	3 360	4 680	4 800	4 920	6 240	6 360	6 480
	7 800	7 920	8 040	9 360	9 480	9 600	10 920	11 040	11 160
	12 480	12 600	12 720						
NOTE: (common_multiplier / nof_late_taps / nof_slices / slice_distance / non_late_increment).									

A.11 Simulations with imperfect channel estimation in demodulator

A further refinement has been performed by defining reference demodulator algorithms for investigating the impact of imperfect channel estimation. The demodulator algorithms that are described below are based on the data-aided approach and are considered representative of practical implementations (enhancements are possible). Carrier frequency and symbol clock extraction have not been included in the model as their impact are considered negligible. The models have been implemented in floating point.

A.11.1 TDM Case

A.11.1.1 TDM Reference Demodulator

The proposed reference demodulator heavily exploits the TDM pilot symbols. The DVB-SH TDM slot structure is depicted in figure A.7. For easing the acquisition and tracking a regular pilot spacing within each PL slot has been included in the DVB-SH standard. Each TDM slot of length L_{TOT} is subdivided in *two* data sub-slots and pilot fields.

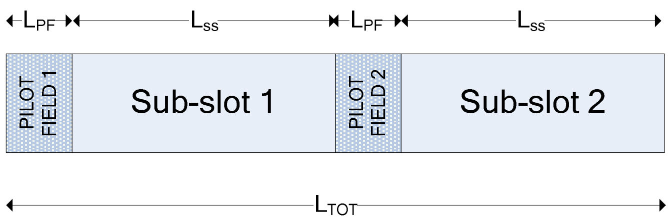


Figure A.7: Slot pilot insertion in DVB-SH TDM frame

Carrier Phase Estimation:

The DVB-S2 carrier phase estimator described in clause 3.6 of [i.16] has been adapted to the DVB-SH case. Following [i.16], the phase estimator consists of a ML feed-forward phase estimator operating on each pilot block belonging to the slot. The ML phase estimator algorithm adapted from [i.16] for the pilot group #n is given by:

$$z_n = \sum_{k=1}^{L_{PF}} \left[\vec{C}\left[k+(n-1)(L_{SS}+L_{PF})\right] \right]^* \vec{r}\left[k+(n-1)(L_{SS}+L_{PF})\right],\ n=1,..\,N_p$$

$$\hat{\theta}_{PF}\left(n\right) = \arg\left\{z_n\right\},\ n=1,..\,N_p$$

where *r* (*k*) represents the k-th on-time baseband demodulator square-root raised-cosine chip matched filter output complex samples after frequency offset removal.

The phase unwrapping allows to implement a simple linear phase interpolator between consecutive pilot estimations.

The final unwrapped pilot estimates $\hat{\theta}_{PFU}\left(n\right)$ are computed from $\hat{\theta}_{PF}\left(n\right)$ as:

$$\hat{\theta}_{PFU}\left(n\right) = \hat{\theta}_{PFU}\left(n-1\right) + SAW\left[\hat{\theta}_{PF}\left(n\right) - \hat{\theta}_{PFU}\left(n-1\right)\right]$$

where $SAW[\Phi] \equiv [\Phi]_{-\pi}^{+\pi}$ is a saw tooth nonlinearity that reduces Φ to the interval (-π, π). Linear interpolation between consecutive pilot blocks *n* and *n*+1 for sub-slice carrier phase removal then follows trough the following equation:

$$\hat{\theta}_{PFUI}\left(l,n\right) = \hat{\theta}_{PFU}\left(n\right) + \frac{l}{L_{SS}}\left[\hat{\theta}_{PFU}\left(n+1\right) - \hat{\theta}_{PFU}\left(n\right)\right]$$

where $l = 1, 2, \ldots, L_{ss}$, is the symbol index within the sub-slice *n*.

Carrier amplitude estimation:

By reusing the pilot-aided estimated signal phasor z_n one can compute the signal amplitude simply as:

$$A\left(n\right) = \left|z_n\right|,\ n=1,..\,N_p$$

Linear amplitude interpolation between consecutive pilot blocks *n* and *n*+1 for sub-slice then follows trough the following equation:

$$A\left(l,n\right) = A\left(n\right) + \frac{l}{L_{SS}}\left[A\left(n+1\right) - A\left(n\right)\right]$$

where $l = 1, 2, \ldots, L_{ss}$, is the symbol index within the sub-slice *n*.

Signal-to-Noise-Ratio Estimation:

The signal to noise ratio estimation is required for the correct turbo decoder operation. For the SNR estimator the data-aided version of the SNORE [SNORE] algorithm based on pilot symbols have been adopted. The algorithmic details can be found in [i.16].

A.11.1.2 TDM Reference Demodulator Simulation Results

To assess the performance of the DVB-SH TDM option a physical layer simulator inclusive of the above described channel estimation algorithms has been implemented.

First a Ricean fading channel with K = 5 dB, mobile speed of v = 50 kmph and worst-case SNR = -3,5 dB has been considered. Channel estimation results are plotted in figures A.8 to A.10. Some high error peaks are present corresponding to deep fading conditions. In these conditions, the channel estimation has no impact on BER as the Turbo decoder will not be able to decode even with perfect channel estimation. The BER curve confirms this analysis.

Another set of channel estimation results have been obtained for the LMS intermediate tree shadowed channel with mobile speed of v = 70 kmph and a line-of-sight SNR = 10 dB (considered to be a representative demodulator operating point in this channel). The pilot-aided phase estimation show to perform well also in this difficult type of channel. Again large phase error peaks correspond to deep channel fading conditions.

Simulated TDM reference demodulator channel estimation for Rice K = 5 dB, mobile v = 50 kmph channel, SNR =-3,5 dB.

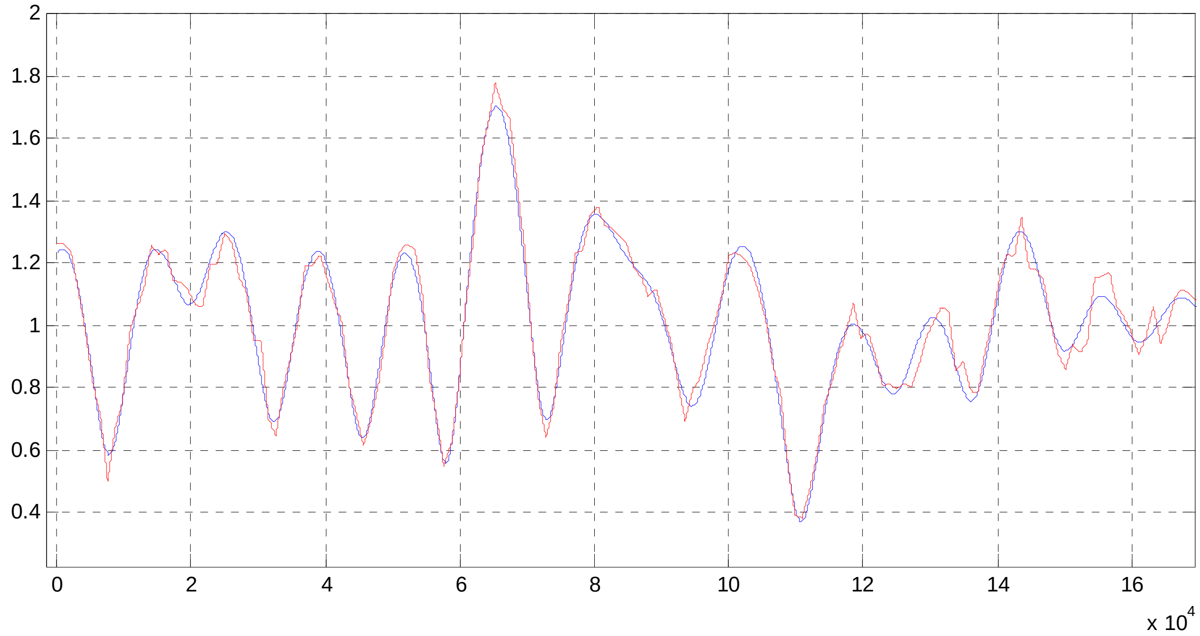


Figure A.8: Amplitude

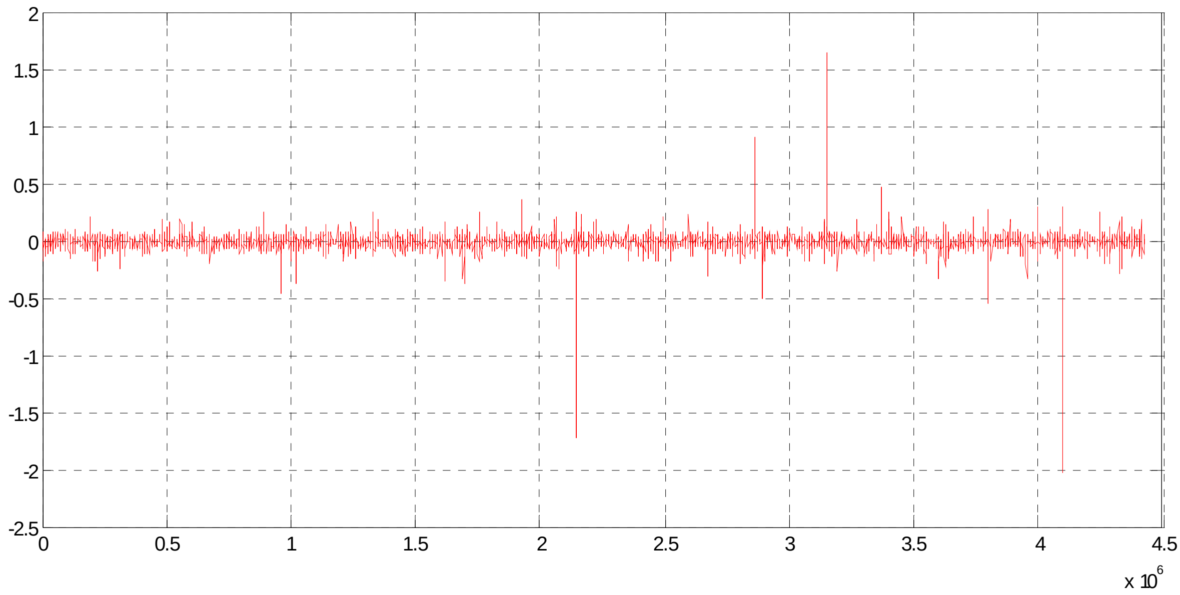


Figure A.9: Phase error in radians

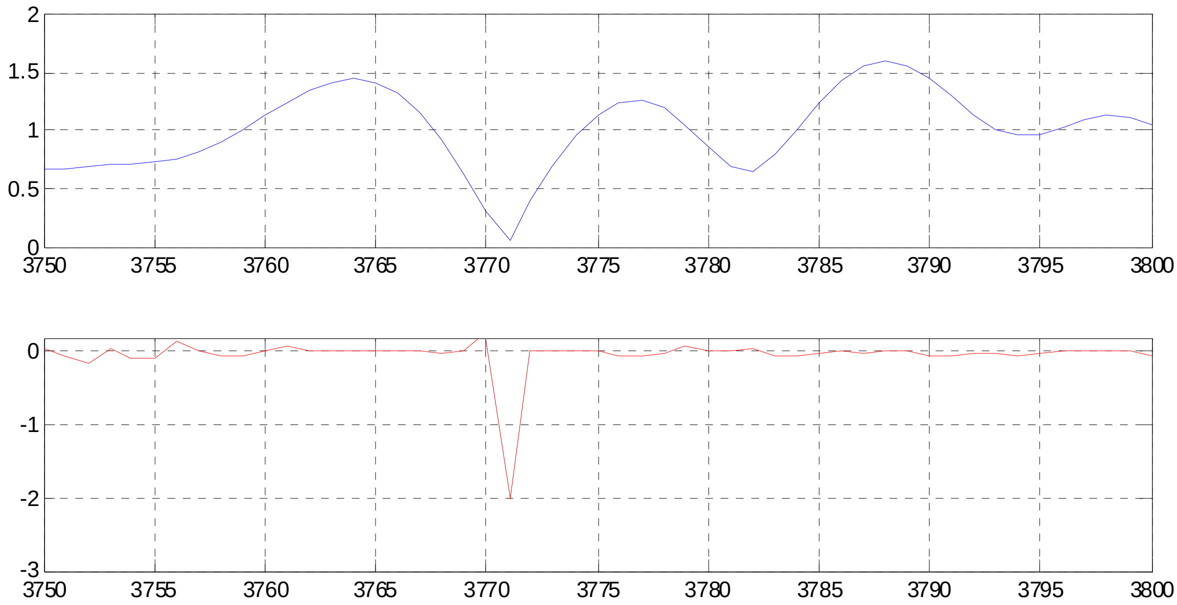


Figure A.10: Zoom of fading channel amplitude and estimated carrier phase error relation

Simulated TDM reference demodulator channel estimation for *v* = 70 kmph, SNR = 10 dB and a LMS intermediate state of the intermediate tree shadowing environment.

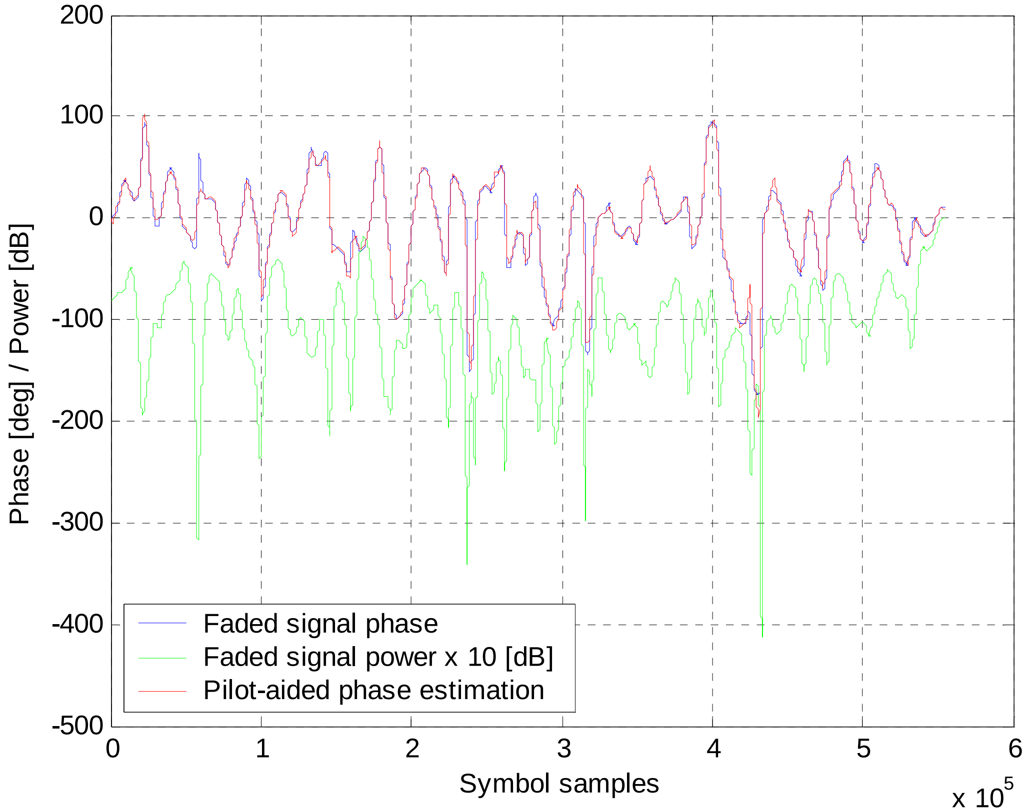


Figure A.11: Real and estimated carrier phase

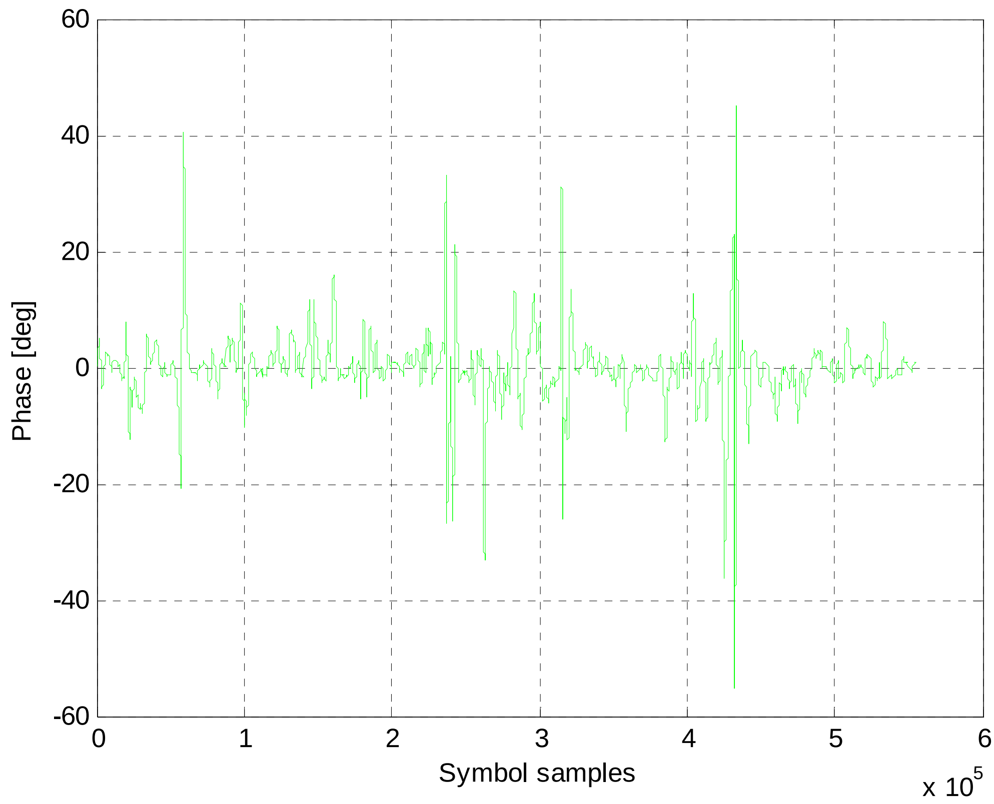


Figure A.12: Estimated carrier phase error

A more comprehensive analysis of the channel estimation impact has been performed looking at the BER performance versus E_s / N_0 in Ricean *K*=5 dB channels. Figure A.13 shows the corresponding results for a short PL interleaving configuration. AWGN and Rice reference performance with ideal channel estimation (for short 200 ms interleaver and infinite length interleaver (uncorrelated fading)) are also reported for convenience. It can be concluded that the proposed reference demodulator channel estimation has negligible impact with Ricean channel *K*=5 dB.

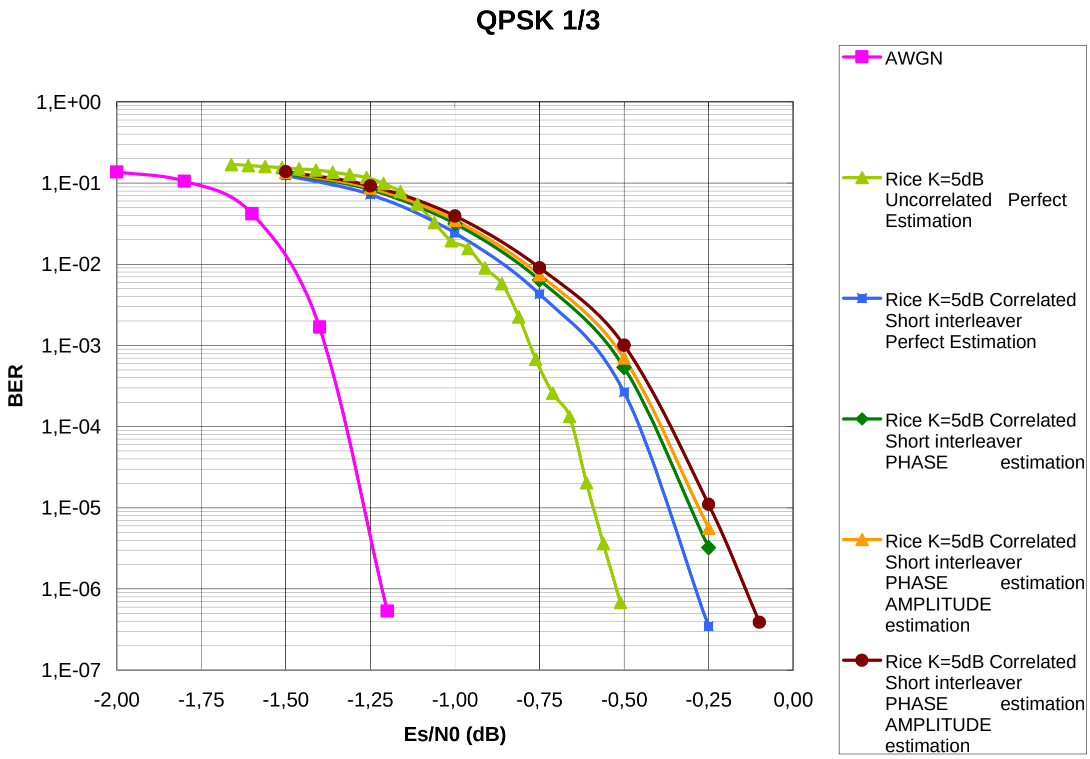


Figure A.13: Simulated BER TDM reference demodulator BER performance for Rice *K*=5 dB, mobile *v*= 50 kmph channel, as a function of E_s / N_0

Reference demodulator performance have also been assessed for the LMS 3-state channel case with the demodulator state machine on. The results at C/N = 10 dB are summarized in the following Table A.23. It is apparent that the TDM reference demodulator has no practical impact on the ESR5 performance.

The final verification performed on the TDM reference demodulator is related to the carrier phase noise impact. For this a synthetic phase noise nuisance following the mask of clause has been included in the simulation chain. We remark that the low frequency component of the DVB-SSP phase noise is less than the DVB-S2 one (about 30 dB at 100 Hz) but higher-frequency component (above 1E5 Hz) is larger. The mild slope of the DVB-SSP phase noise PSD makes the higher frequency component contribution dominating. This makes the phase noise time series much faster than the fading/shadowing ones. The simulated BER performance results are reported in figure A.14. It can be concluded that for Rice channel with *K* = 5 dB and mobile speed of 50 kmph the carrier phase noise impact for QPSK r = 1/3 is less than 0,05 dB at BER = 1E-5.

Table A.23: Summary of ESR5 results at C/N = 10 dB with TDM state machine on and ideal or reference demodulator channel estimation

Reference case	FEC/Interleaver	ESR5 satisfaction % ideal channel estimation	ESR5 satisfaction % reference demodulator channel estimation
SubUrban (SU) V = 3 kmph	Short PL interleaver	45,6	48,9
	Non uniform long PL interleaver	88,5	87,6
	Uniform long PL interleaver	89,3	88,5
	Short PL interleaver plus upper layer FEC	87,6	Not available
Intermediate Tree Shadowing (ITS) V = 50 kmph	Short PL interleaver	0,0	0,0
	Non uniform long PL interleaver	25,2	25,6
	Uniform long PL interleaver	80,4	78,2
	Short PL interleaver plus upper layer FEC	19,5	Not available

Table A.24: S-band DVB-SH phase noise mask (Satellite contribution only)

	10 Hz	100 Hz	1 000 Hz	10 kHz	100 kHz	1 MHz	10 MHz
Phase noise (dBc/Hz)	-29	-59	-69	-74	-83	-95	-101

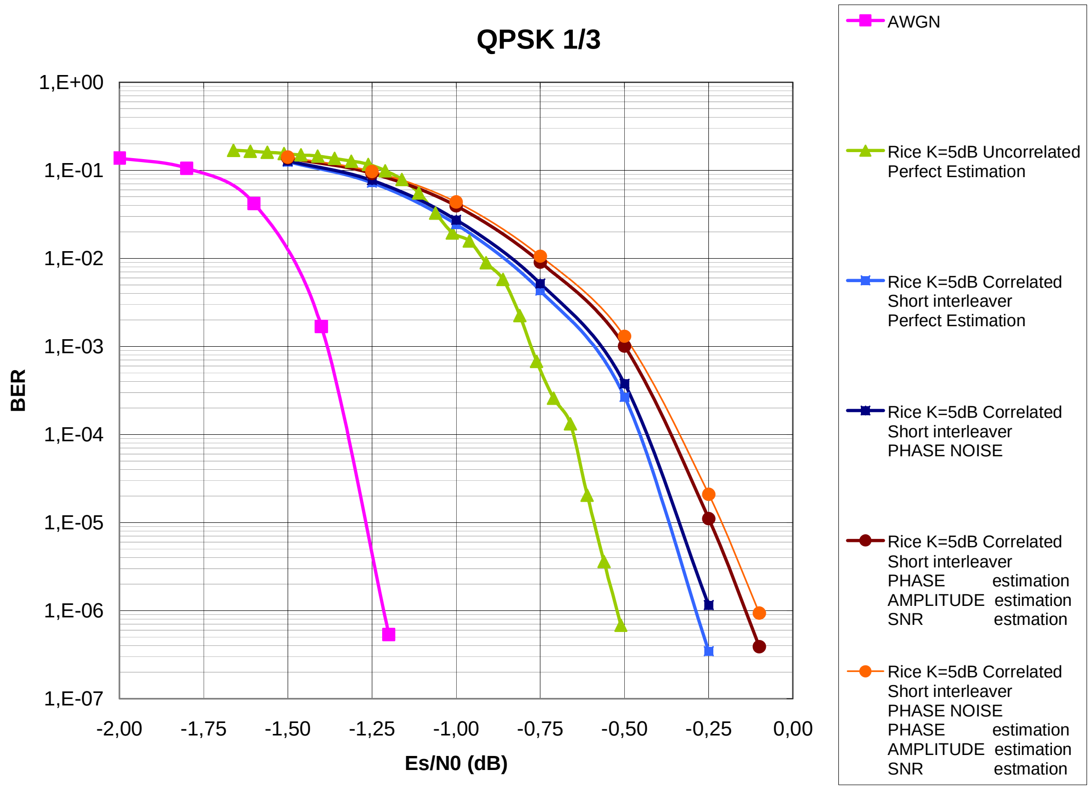


Figure A.14: Simulated BER TDM reference demodulator BER performance for Rice *K* = 5 dB, mobile *v* = 50 kmph channel, as a function of E_s / N_0 with and without phase noise

A.11.2 OFDM Case

A.11.2.1 OFDM Reference Demodulator (see note)

Channel transfer estimation is performed through adaptive Least Mean Square (LMS) interpolation both on time and frequency domain thanks to the knowledge of the channel at pilot positions. Firstly time-domain adaptive interpolation is performed to calculate the channel transfer function at scattered pilot subcarriers. Based on these estimates, frequency-domain adaptive interpolation is then performed at the non-pilot subcarriers. The following scheme is based on the assumption of a DVB-SH compliant OFDM transmission.

At the pilot symbols location, the transfer function of the channel is calculated as:

$$H(n,k) = \frac{Y(n,k)}{X(n,k)} \tag{A.1}$$

where $Y(n,k)$ represents the received complex symbol at for the *k*-th subcarrier of the *n*-th OFDM symbol and $X(n,k)$ represents the transmitted pilot known at the demodulator corresponding to the same position in time and frequency.

Time domain interpolation

a) Filter coefficients update

The interpolator coefficient update exploits the information carried by the continual pilot subcarriers, where the channel transfer function is known at each OFDM symbol.

The result from (Eq.A.1) is then compared to the estimate of the channel in the same position (n,k) and performed through the interpolator filter. The filter coefficients are updated thanks to the estimation error, calculated as the distance between the estimated channel and the calculated one.

The time-domain interpolator filter is designed to be $4(M_1+M_2)$-OFDM symbol long, where $4M1$ and $4M_2$ represent the number of exploited OFDM symbols before and after the instant to be estimated.

The procedure is divided into three separate steps.

Step 1: Channel transfer function estimation

The channel estimation is performed filtering the pilots of the tone under analysis, using the set of coefficients calculated thanks to the previous *(n-1)*-th OFDM symbol.

The estimation is repeated three times employing three different set of coefficients and pilot symbols belonging to continual subcarrier C_i (a graphical interpretation can be find in figure A.15):

$$H_L(n, C_i) = \sum_{m=-M_1+1}^{M_2} W_L(n-1, m) H(n-L+4m, C_i) \qquad \text{(A.2)}$$

where $W_L(n-1, m)$ represents the *m*-th coefficient of the *L*-th filter (*L*=1,2,3), updated during the previous *(n-1)*-th OFDM symbol.

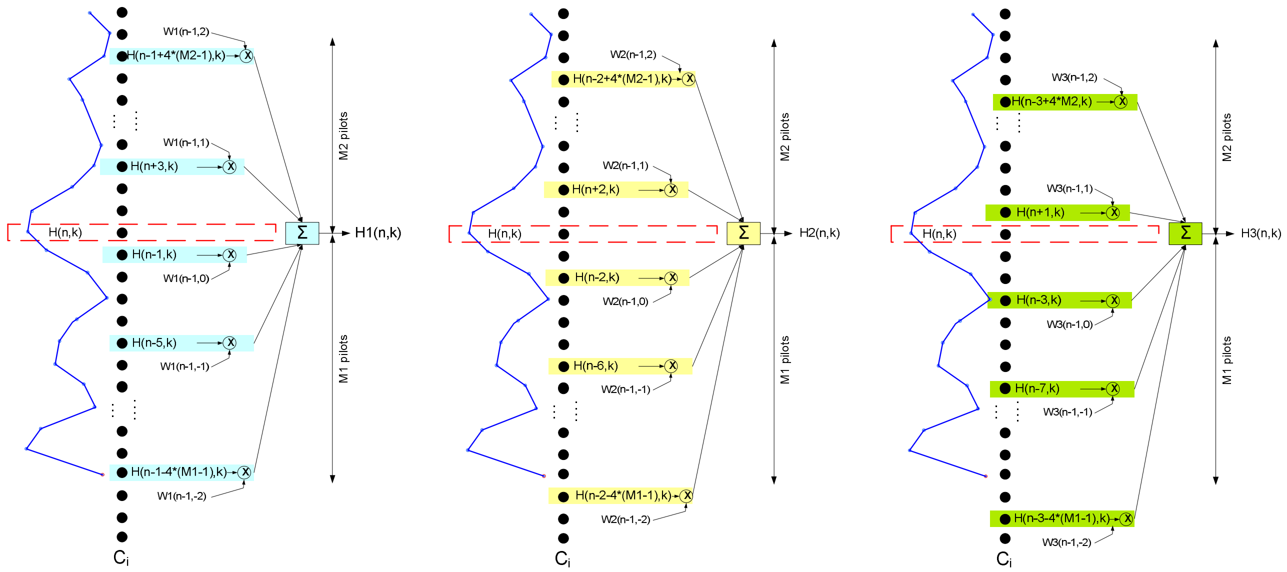


Figure A.15: Graphical representation of the time domain channel estimation algorithm on continual pilot C_i

Each filter performs the channel estimation for the same cell of the *n*-th OFDM symbol. Only one pilot every four is exploited and each set consists of a different pilot group. This redundancy is justified by the regular distribution of the scattered-pilot cells (see figure A.16). The column C_i represents the (time) continual pilots while the column S_j represents the (time) scattered pilots.

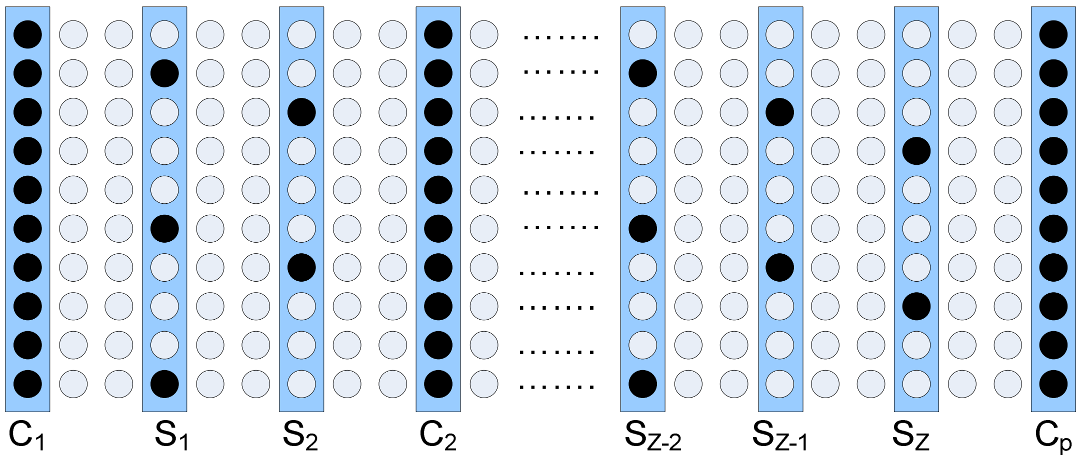


Figure A.16: Graphical representation of the DVB-SH OFDM pilot distribution in time (vertical axis) and frequency (horizontal axis)

In general, the parameters M_1 and M_2 can be independently defined for each filter. Furthermore, when M_2 is set to zero the filter is behaving as a predictor instead as an interpolator.

This procedure is applied to each continual pilot tone (C_i) and repeated for each received OFDM symbol (n).

Step 2: Estimation error calculation

The estimation error for the filter updating is then derived from the estimates performed in Step1 (Eq.2) and the knowledge of the channel thanks to the pilot symbol transmission, according to the following equations for L=1,2,3:

$$E_L(n,C_i) = H(n,C_i) - H_L(n,C_i) \tag{A.3}$$

The error calculation is performed on each continual pilot tone (C_i) and repeated for each received OFDM symbol (n).

Step 3: Update of the time domain interpolation filter coefficients

The interpolator filter coefficients are then updated using the errors calculated in Step2. The following summation are over all continual pilot tones (1..P) and repeated for every m ranging in [$-M_1$+1:M_2] for L=1,2,3:

$$W_L(n,m) = W_L(n-1,m) + \mu \sum_{i=1}^{P} E_L^{*}(n,C_i)\, H(n-L+4m, C_i) \tag{A.4}$$

where μ represents the time interpolator filter LMS adaptation step.

b) Scattered pilot interpolation

The filter coefficients calculated exploiting the subcarriers containing continual pilot symbols are then employed to estimate the channel in non-pilot symbols position for the subcarriers containing scattered pilot symbols.

The regular distribution of the scattered pilots ensures pilots to be transmitted every 4 OFDM symbols (then the factor 4 in Eq.A.2). The distance between a non-pilot carrier-symbol and the previously transmitted pilot is not higher then 3 OFDM symbols. According to which OFDM symbol is transmitting the pilot (n-1, n-2, n-3), the proper filter coefficient set W_1, W_2 or W_3 must be used for the interpolation. More specifically the interpolated channel estimate for symbol n-L belonging to the scattered pilot subcarrier S_i is computed as:

$$H(n,S_i) = \sum_{m=-M_1+1}^{M_2} W_L(n,m)\, H(n-L+4m, S_i) \tag{A.5}$$

The channel estimation $H\left(n,S_i\right)$ is based on the knowledge of the channel at $H\left(n-L+4m,S_i\right)$ with L=1, 2, 3 where scattered pilot symbols are transmitted. This clarifies the necessity of three distinct sets of filter coefficients. This procedure is applied to each scattered-pilot subcarrier (S_i) and repeated for each received OFDM symbol (n).

Frequency domain interpolation

The interpolation in the frequency domain is then performed to derive the channel estimation for the OFDM subcarriers not containing any pilot symbol. The channel transfer function estimation is based on the knowledge of the channel at the continual and scattered subcarriers derived during the previous time-domain interpolation phase. These stones are highlighted in figure A.16. It clearly appears that two subcarriers are not carrying at all pilot symbols.

The frequency-domain interpolator filter is designed to be $3(J_1+J_2)$ subcarrier long, where $3*J_1$ and $3*J_2$ represent the number of exploited OFDM subcarriers before and after the subcarrier to be estimated.

a) Channel interpolation for pilot-less subcarriers

Following an approach similar to the one applied for the time-domain interpolation, two distinct interpolation filters V_I with I=1, 2 are defined in the frequency-domain:

$$H\left(n,k\right)=\sum_{m=-J_1+1}^{J_2} V_I\left(n-1,j\right)H\left(n,k-I+3j\right),\text{ for } k=3K+I \quad \text{(A.6)}$$

where K is an integer and V_I (n-1,j) represents the j -th coefficient of the L-th filter (L=1,2), updated during the previous *(n-1)*-th OFDM symbol.

Each filter performs the channel estimation for the k-th subcarrier. The indices relation $k=3K+I$ ensures that only the not-continual and not-scattered subcarriers are considered. The channel estimation $H\left(n,k\right)$ is based on the knowledge of the channel cells $H\left(n,k-I+3j\right)$ *I=1, 2,* calculated during the time-domain interpolation.

b) Filter Coefficients Update

Step 1: Estimation error calculation

In the absence of pilot symbols, the error signals for the coefficient update are calculated based on a decision-direct approach. After the computation of the channel estimate $H(n,k)$, the estimated transmitted symbol $X\left(n,k\right)$ can be derived as:

$$X\left(n,k\right)=H_D\left\{\frac{Y\left(n,k\right)}{H\left(n,k\right)}\right\} \quad \text{(A.7)}$$

being $H_D\{\ \cdot\ \}$ the function representing the complex symbol hard decision. For QSPK modulation the function $H_D\{\ \cdot\ \}$ is simply given by $H_D\{z_n\}=\frac{1}{\sqrt{2}}\{\text{sign}[\text{Re}(z_n)]+j\ \text{sign}[\text{Im}(z_n)]\}$.

Assuming the symbol hard decision reliable, it can be used as a pilot symbol so that the channel transfer function can be re-calculated according to:

$$H\left(n,k\right)=\frac{Y\left(n,k\right)}{X\left(n,k\right)} \quad \text{(A.8)}$$

The Estimation Error is defined as:

$$D(n,k)=H\left(n,k\right)-H\left(n,k\right) \quad \text{(A.9)}$$

Step 2: Update of the Frequency Domain Interpolation Filter Coefficients

The interpolator filter coefficients are then updated using the errors calculated in Step 1. The following summations are over all possible tones (k) to be updated (not-scattered and not-continual pilot subcarriers) i.e. for k = 3K+I, I=1, 2:

$$V_I(n,j) = V_I(n-1,j) + \gamma \sum D^*(n,k) H(n,k-L+3j) \tag{A.10}$$

where γ represents the frequency interpolator filter LMS adaptation step.

NOTE: The algorithms described here are inferred from the descriptions given in US patent US 2006/0269016 A1, Nov. 2006.

A.11.2.2 OFDM Reference Demodulator Simulation Results

A.11.2.2.1 Terrestrial Channel

First the OFDM demodulator has been tested over the TU6 terrestrial reference channel, as described in TS 145.005 [29]. A fading Doppler spread range spanning from f_d =50 Hz up to f_d = 400 Hz (roughly equivalent to a mobile speed from v = 25 kmph up to v = 200 kmph at S-band) and an SNR=-2,0 dB have been used. Demodulator channel estimation simulation results are plotted in figures A.17 and A.18 where the OFDM subcarrier estimates are serialized in time (the frequency domain is read first). Figure A.17 represents the signal amplitude estimation (orange curve) versus the real one (black curve) as a function of the received symbols. Figure A.18 shows the correlation between large carrier phase estimation errors and deep channel fading conditions. These large phase errors have no practical impact on the end results. Turbo decoder will be anyway in error due to the very low SNR. Figures A.20 to A.22 confirm this analysis.

Simulated OFDM reference demodulator channel estimation for TU6 with fading Doppler spread fd=50Hz, SNR=-2.0 dB.

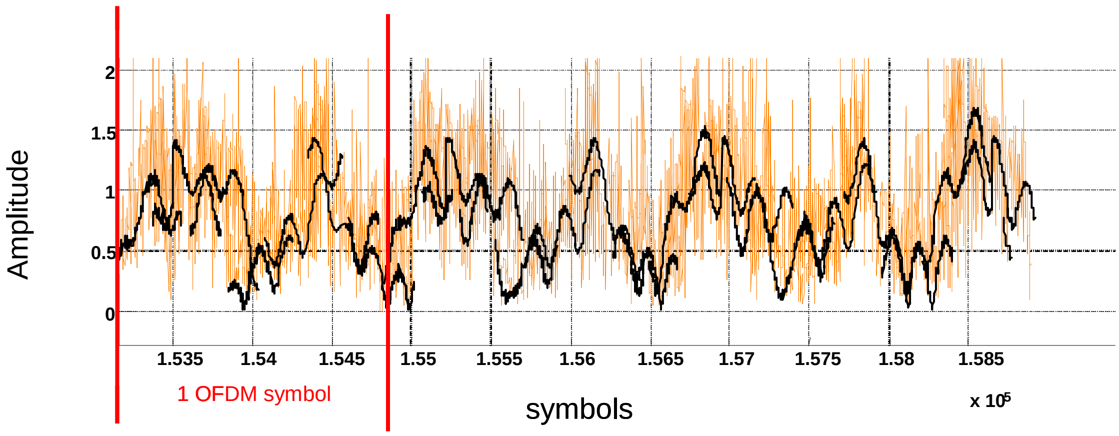


Figure A.17: Channel amplitude (black real, estimate orange)

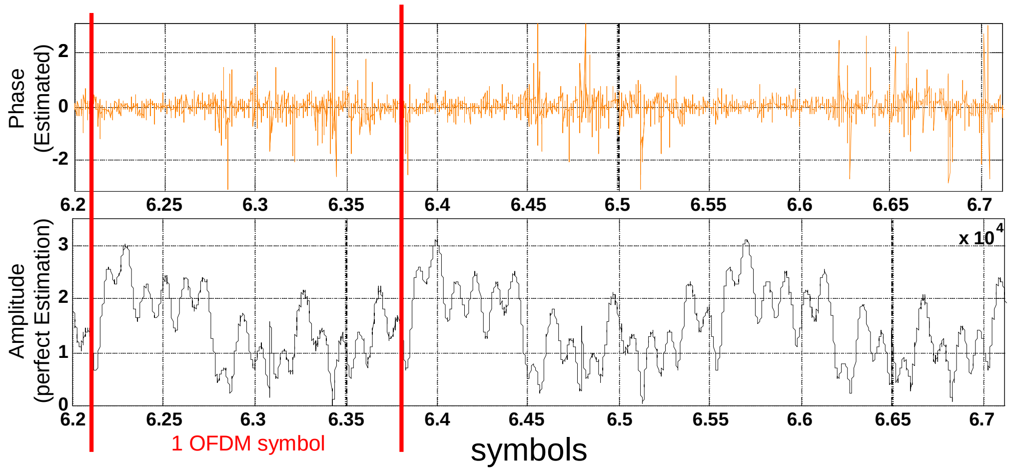


Figure A.18: Fading channel amplitude (black) and estimated carrier phase error (orange)

Another way to analyse the adaptive channel estimation algorithm performance is to look at its capability to estimate the fading bandwidth (Doppler spread). Figure A.19 compares the simulated TU6 fading process bandwidth (blue curve) with the reference demodulator fading bandwidth estimate (red curve). It is remarked the good fading bandwidth estimation performed by the selected LMS adaptive algorithm.

Example of filter adaptation at a continual-pilot-tone position.

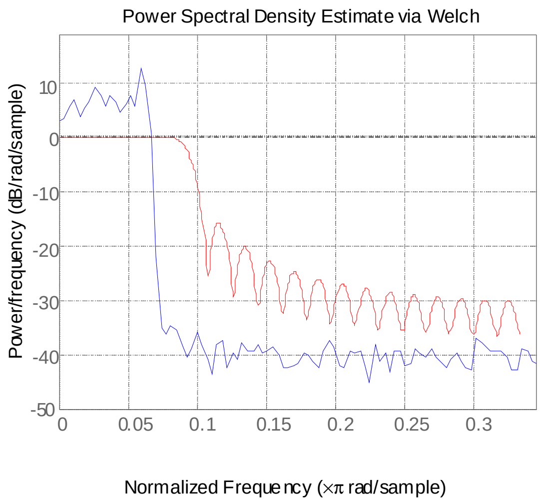


Figure A.19: Channel response of LMS-filter (red plot) to a 50 Hz Doppler TU6 channel (blue plot)

A more comprehensive evaluation of the OFDM reference demodulator channel estimation impact over terrestrial channels has been performed simulating the BER performance versus E_s / N_0 in the TU6 channel. Figures A.20 to A.22 and figures A.23 to A.25 show the corresponding results for a short (±200 ms) PL interleaving configuration at different Doppler spread values. AWGN reference performances with ideal channel estimation are also reported for convenience. Figure A.20 (A.23) and figure A.21 (A.24) show results with perfect and real channel estimation respectively. Figure A.22 (A.25) show the dependency of demodulator losses to the fading Doppler frequency when the target BER is set to 10^{-5}. The losses of the an ideal demodulator w.r.t. AWGN reference performance (blue curve) are decreasing with increasing Doppler frequency as the PL interleaver is increasingly able to decorrelate the fading. Instead losses are increasing with the Doppler spread in case of real channel estimation as the channel estimation errors are growing with the user (and channel) speed (dynamic). Finally the black curve in figure A.18 shows the implementation losses of the reference demodulator compared to ideal channel estimation. In the QPSK-1/3 case, the channel estimation losses are below 1 dB for f_d =50 Hz and goes up to about 4 dB for f_d = 300 Hz. In the QPSK-1/2 case, the losses even lower at low Doppler frequency, below 0,5 dB for f_d = 50 Hz, but remarkably higher, about 15 dB, for f_d = 300. These results are considered representative of a typical OFDM demodulator.

Simulated BER OFDM reference demodulator as a function of Es/N0, for QPSK-1/3 in TU6 channel with several Doppler frequencies.

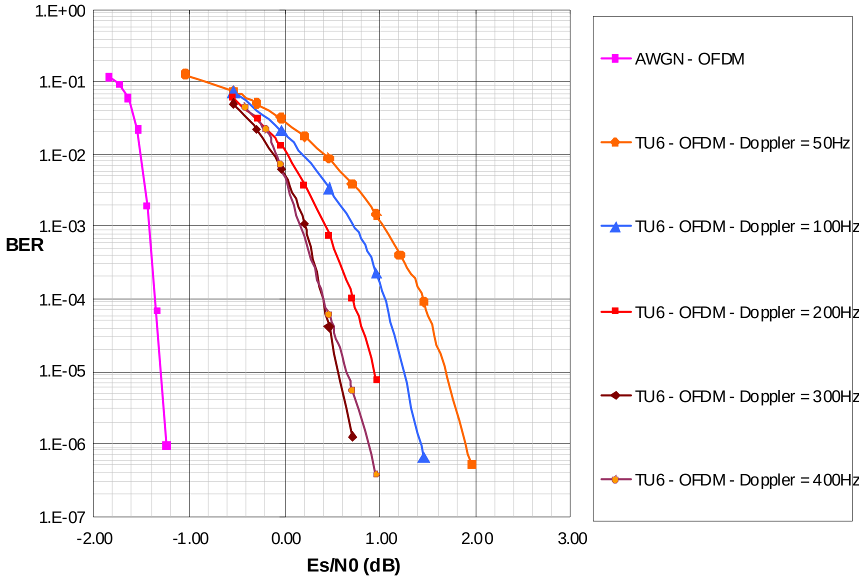


Figure A.20: Perfect channel estimation

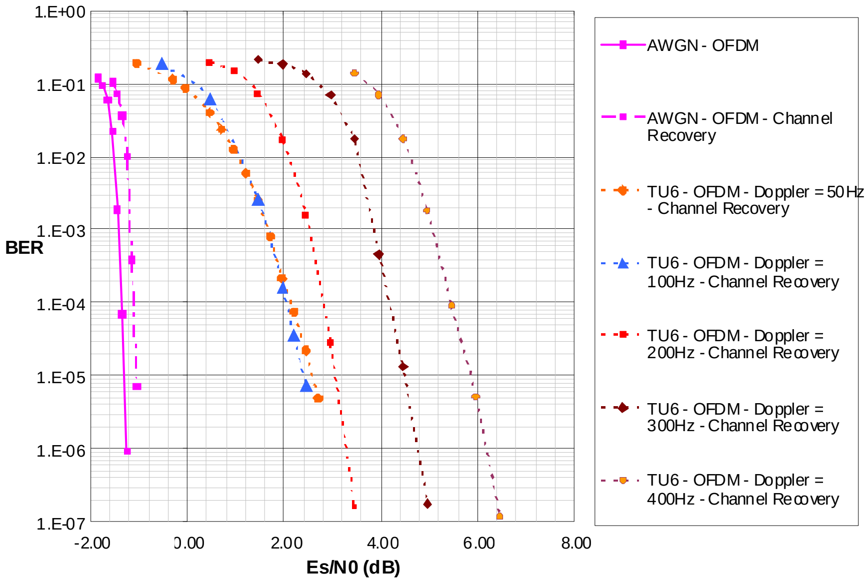


Figure A.21: Real channel estimation

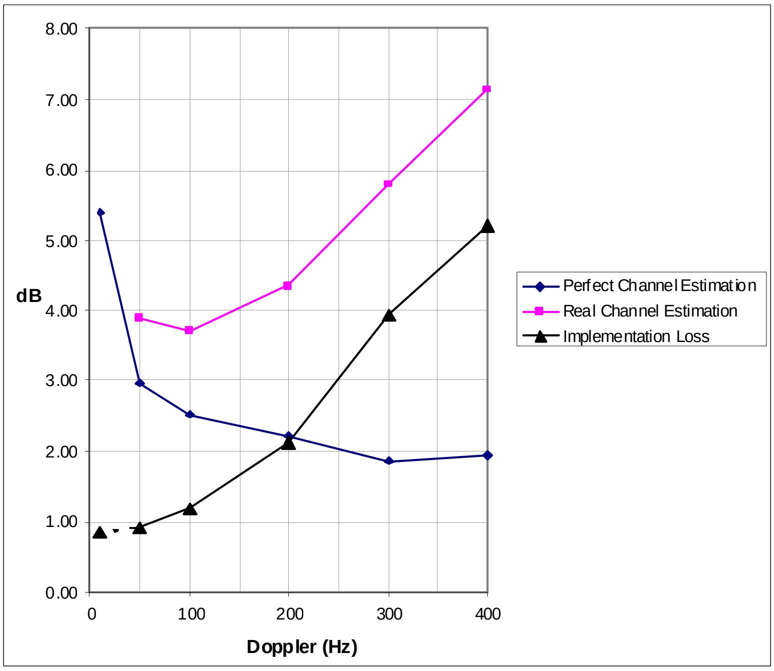


Figure A.22: Losses w.r.t. AWGN reference (Perfect and Real channel estimation) and Implementation losses of a real w.r.t. perfect channel estimation (at BER = 10-5)

Simulated BER OFDM reference demodulator as a function of Es/N0, for QPSK-1/2 in TU6 channel with several Doppler frequencies.

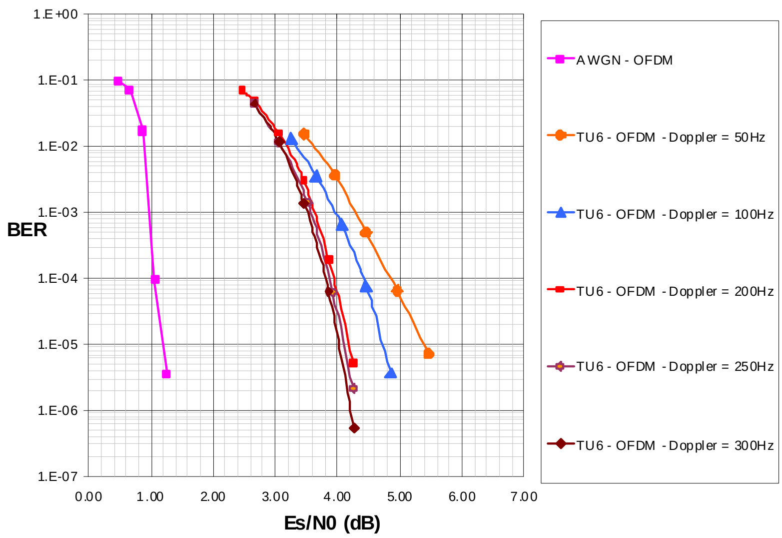


Figure A.23: Perfect channel estimation

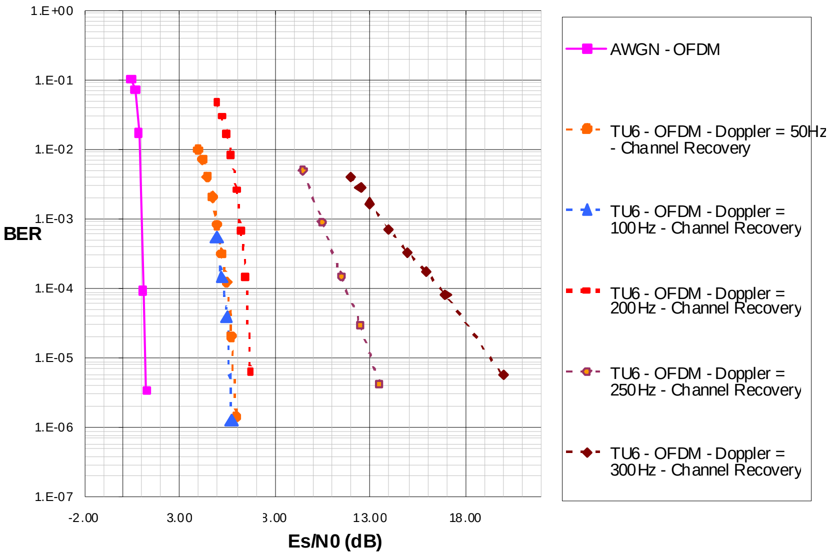


Figure A.24: Real channel estimation

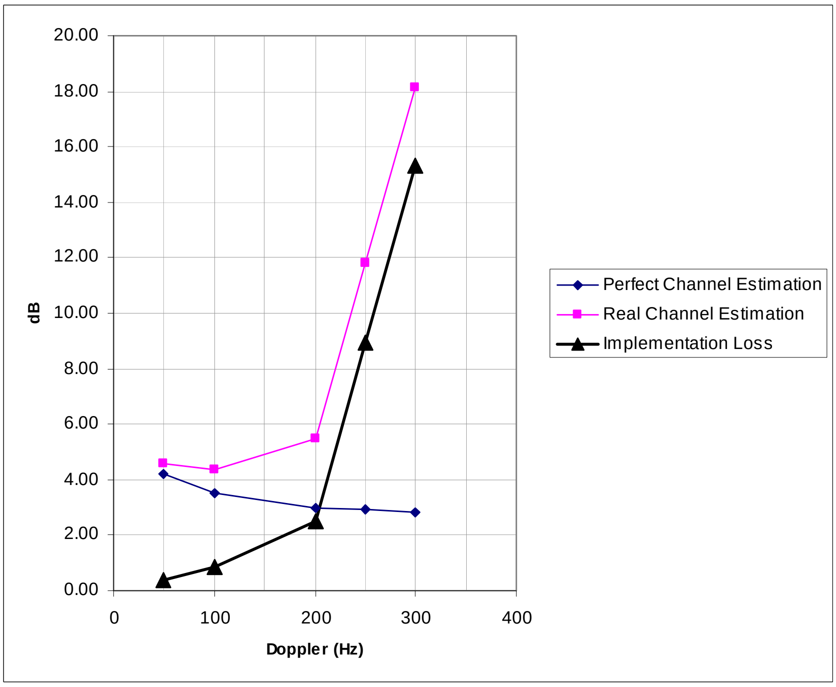


Figure A.25: Losses w.r.t. AWGN reference (Perfect and Real channel estimation) and Implementation losses of a real w.r.t. perfect channel estimation (at BER = 10-5)

A.11.2.2.2 Satellite Channel

Another set of channel estimation results have been obtained for a satellite flat Ricean fading channel with Rice factor K = 5 dB, mobile speed of v = 50 kmph and SNR = -2,0 dB (the same algorithm with the same parameters as above is used here). As described in clause A.10, the reference demodulator algorithm are the same for the terrestrial and satellite case. In case of satellite only operations, being the fading channel typically flat, there is no need for frequency interpolation of the channel estimates. No attempt has been done to optimize the channel estimation for the satellite-only type of operations.

The reference demodulator channel estimation results are plotted in figures A.26 to A.27: BER performances versus E_s/N_0 are plotted in figure A.28 for both perfect and real channel estimation. The impact of the proposed reference demodulator amount to 0,7 dB with Ricean channel and a mobile speed of 100 kmph compared to ideal channel estimation.

Simulated OFDM reference demodulator channel estimation for Ricean, mobile speed 50 kmph, SNR=-2,0 dB

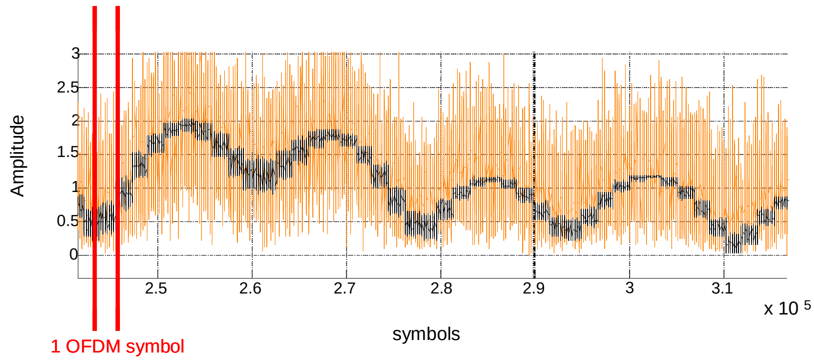


Figure A.26: Amplitude

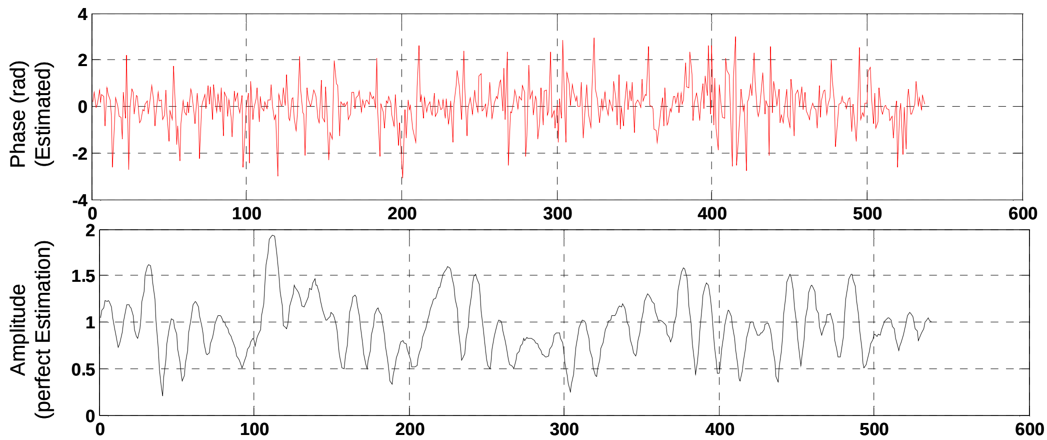


Figure A.27: zoom of amplitude, zoom of fading channel amplitude and estimated carrier phase error relation and zoom of fading channel amplitude and estimated carrier phase error relation for a single FFT-tone

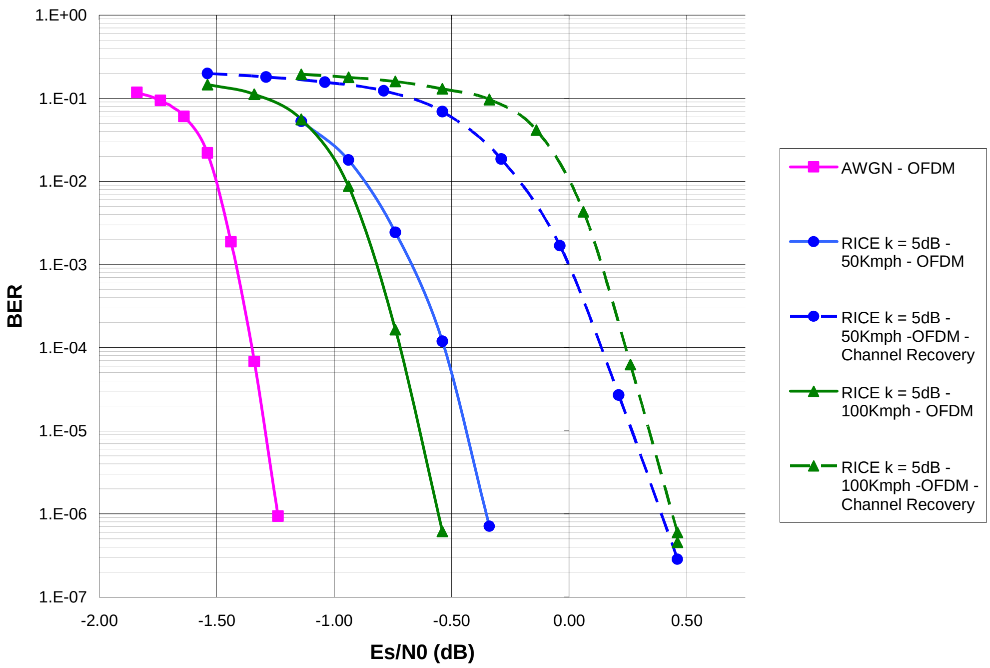


Figure A.28: Simulated BER OFDM reference demodulator as a function of E_s/N_0, for Ricean channel with mobile speed 50 kmph and 100 kmph: perfect channel estimation versus real channel estimation

A.11.3 TDM Signalling Channel Performance Results

In this clause we report some results obtained by simulation on the DVB-SH TDM signalling channel performance. At the terminal switch on it is important to receive the SH network key PL configuration parameters through correct signalling channel detection.

The performance have been derived under the following conditions:

- dynamic 3-state LMS-ITS, LMS-SU channel;
- 400 000 signalling field simulated;
- distance between two consecutive signalling fields (SH frame payload) is taken into account;
- accumulation based on:
  - equal gain combining;
  - maximal ratio combining.

The computed performances are:

- signalling Block Word Error Rate;
- probability of successful detection versus number of received signalling fields.

Figure A.29 shows the simulated signalling field Word Error Rate as a function of the soft accumulated signalling fields for ITS LMS channel with a mobile speed of 50 kmph and E_s / N_0 =-3,5 dB. Two way of performing soft combining have been considered the maximal ratio (MRC) and the equal gain (EGC) ones. As expected, the MRC provides superior performance compared to the ECG.

Figure A.30 shows the probability of correct TDM signalling field detection as a function of the number of received signalling fields for the 3-state LMS-ITS mobile channel v = 50 kmph, E_s / N_0 =-3,5 dB. The results is of course dependent on how many signalling fields are softly combined, but in the worst case of no field combining after 30 signalling fields the signalling information is successfully detected with 99 % probability. With 5 fields soft combined 15 fields are sufficient to have the same detection probability.

Figure A.31 reports the probability of correct TDM signalling field detection as a function of the number of received signalling fields for the 3-state LMS-SU mobile channel v = 3 kmph, E_s / N_0 =-3,5 dB. The results is of course dependent on how many signalling fields are softly combined, but in the worst case of no field combining after 30 signalling fields the signalling information is successfully detected with 99 % probability. With 5 fields soft combined 10 fields are sufficient to have the same detection probability.

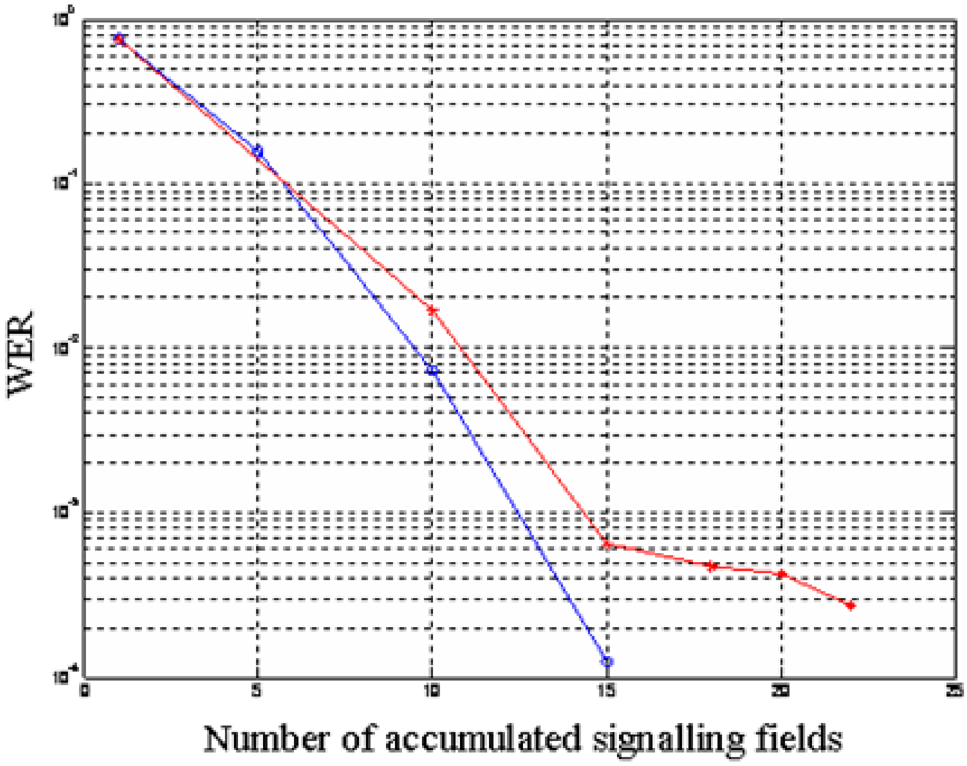


Figure A.29: TDM signalling Word Error Rate as a function of the number of soft accumulated signalling fields for the 3-state LMS-ITS mobile channel *v* = 50 kmph, E_s / N_0 =3,5 dB

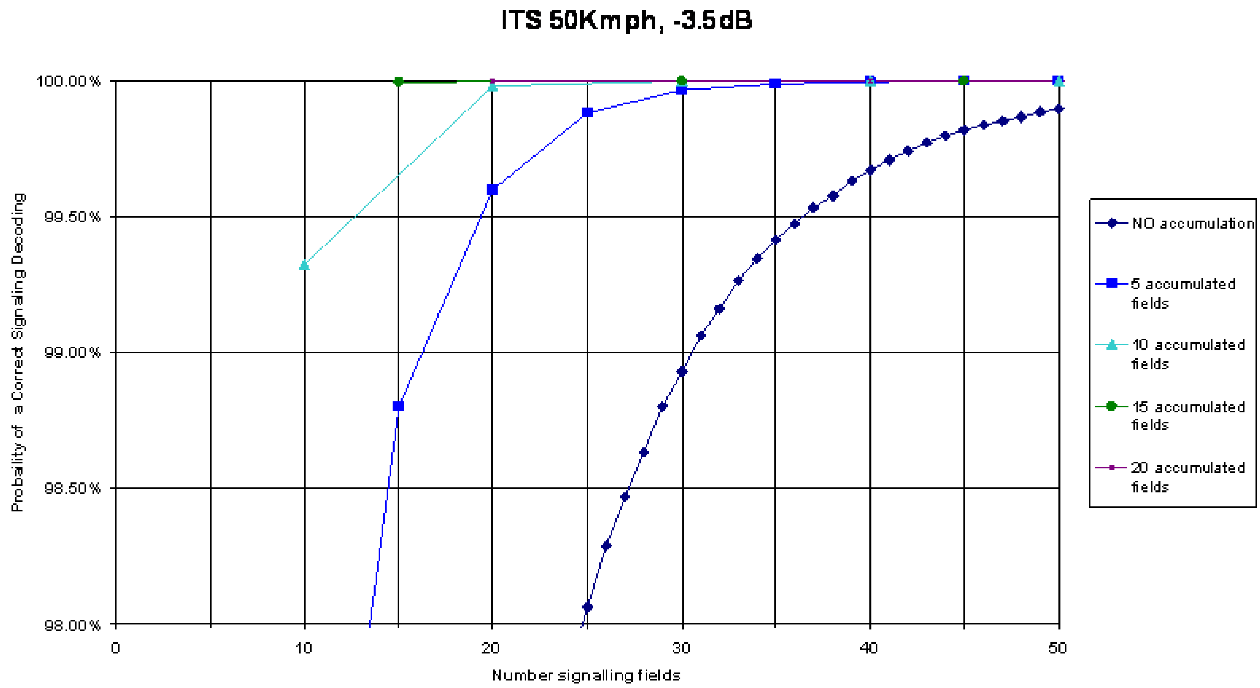


Figure A.30: Probability of correct TDM signalling field detection as a function of the number of received signalling fields for the 3-state LMS-ITS mobile channel *v* = 50 kmph, E_s / N_0 =-3,5 dB

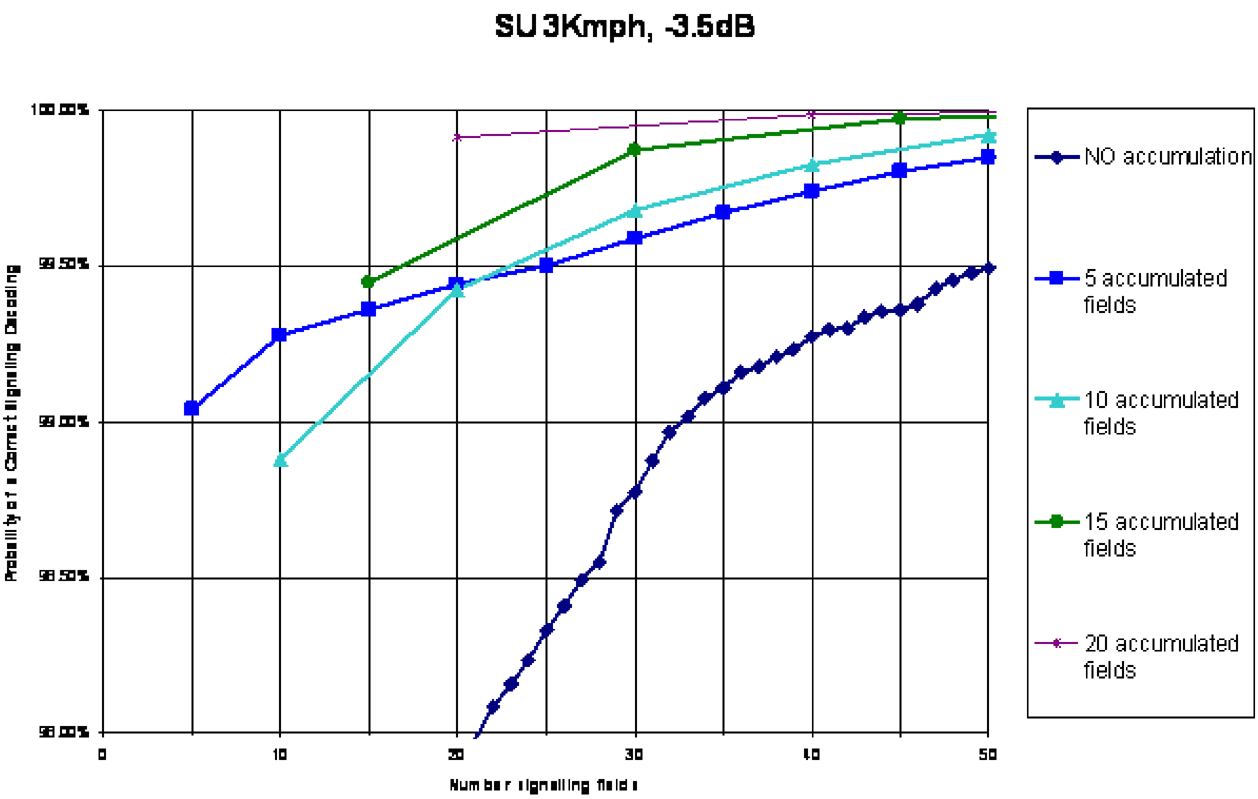


Figure A.31: Probability of correct TDM signalling field detection as a function of the number of received signalling fields for the 3-state LMS-SU mobile channel *v* = 3 kmph, E_s / N_0 =-3,5 dB

A.12 Simulation results

The simulation results reported here are first based on the system configurations described above, referred to as the "Reference Cases" and briefly recalled below. They allow comparison of different physical configurations **under the same high level technical parameter settings (spectrum efficiency, satellite and terminal characteristics or Interleaver length), irrespective of the receiver complexity trade-off, such as memory size**. In addition, supplementary cases have also been performed, referred to as "Sensitivity Analysis", that give complementary information on respective capabilities of different physical configurations, when specific constraints are relaxed/changed. Because the constraints are not relaxed/changed in the same way, direct comparison of these additional results are not straightforward.

Simulated environments

The characteristics of the main terminal Categories are:

Table A.25

	Handheld	**Handheld**	**Portable**	**Vehicular**
	Category 3	category 2b	category 2a	category 1
Antenna polarization	L or C	L or C	L or C	C
G/T (dB/K)	**-32,1**	**-29,1**	**-24,9**	**-21,0**

Service delivery environments/use cases/terminal categories lead to a large combination of simulations targets. The simulations reported here are restricted to the following targets:

1) satellite coverage, Category 1 terminals, LMS-SU model (Suburban);

2) satellite coverage, Category 1 terminals, LMS-ITS model (Rural and forested);

3) satellite coverage, Category 2b terminals, LMS-SU model;

4) terrestrial coverage, any terminal category, TU6 propagation model (fast fading).

Target QoS criteria for the simulations

Two QoS parameters have been computed:

1) for a global approach to QoS the ESR5 fulfilment is used, according to definition in table A.13;

2) for easy reference and reuse of terrestrial network planning efforts conducted on other broadcasting services for terrestrial coverage the FER/MFER5 are also used.

Simulation parameters

The following parameters have been defined to identify uniquely each case:

- waveform configuration, covering all main settings of the DVB-SH waveform:
  - physical layer parameters:
    - transmission Mode: OFDM or TDM for LMS environment, and OFDM only for TU6 environment;
    - physical layer Modulation: QPSK and 16QAM for OFDM, QPSK and 8PSK for TDM;
    - physical layer FEC code rate;
    - physical layer Interleaver Profile: Short (S), 200 ms long and Uniform (U) or Uniform Late (UL) (see clause A.10) with a reference total duration in the range of 10 s;
    - OFDM simulations have been performed using a Guard Interval of 1/4;
  - link Layer parameters (when applicable):
    - link Layer code rate;

    - link Layer Length of the Interleaving, defined as the B+S value (see clauses 6 and A.6 for details) with a reference total duration in the range of 10 s (clause A.3.1 for the delay breakdown);

- channel propagation types as described in clause A.7;
- state Machine Activation (clause A.5), simulating demodulator synchronization behaviour in LMS channel;
- two reference terminal speeds: 3 kmph for Handheld, and 50 kmph for vehicular;
- Carrier-to-Interference ratio (C/I) and Carrier-to-Noise ratio (C/N) according to the link budget calculations (section A.9);
- two types of terminals are considered for LMS simulations, each associated with a specific satellite EIRP:
    - category 1 terminals, with circular polarized antenna, receiving signal from a 63 dBW EIRP satellite;
    - category 2b terminals, with linear polarized antenna, receiving from a 68 dBW EIRP satellite;
- TU6 performance is simulated independently of the terminal category, as it characterizes the required C/N in terrestrial fast fading environment;
- for LMS environments, Class 1 is always used together with MPE-IFEC.

The complete list of Reference Cases is given in table A.14 (at Physical Layer level) for both terrestrial and satellite applications and in tables A.1 to A.4 for Class 1 in LMS.

Simulation results presentation

The following parts of this clause are organized as follows:

- summary of the results for LMS is presented in clause A.12;
- presentation of the simulated performance in LMS environment, including vehicular reception in LMS-Suburban, vehicular reception in LMS-ITS, and handheld reception in LMS Suburban is reported in clause A.12.2;
- presentation of the simulated performance in TU6 environment is reported in clause A.12.3.

A.12.1 Summary of the Results for LMS channel

The results for the Reference Cases show the following main performance differences between the configurations:

1) Physical-Layer (class 2) vs Link-Layer (class 1+ MPE-IFEC) protection

The Reference configuration results show that protection provided exclusively at physical layer (class 2 receivers) performs better than an "equivalent" combination of protections at Physical Layer and Link Layer (class 1 receivers with MPE-IFEC), under the same constraints.

More precisely:

- when the constraint of handheld and portable terminals is removed, keeping only vehicular terminals (Category 1):
    - in LMS-SU, ESR5 fulfillment is above 99 % for Class 2, whereas for Class 1 it is from 88 % to 98 %;
    - in LMS-ITS, ESR5 fulfillment is from 86 % to 100 % for Class 2, whereas Class 1 cannot achieve 90 %;
    - class 2 often exceeds the targeted performance, even for the lowest satellite EIRP considered, both in LMS-ITS and LMS-SU. Therefore, C/N reduction for Class 2 is addressed in the "Sensitivity Analysis" clause below;
    - class 1 with MPE-IFEC, most likely require interleavers longer than 10 s (see clause 10 of the guidelines). Therefore, ESR5 fulfillment improvement versus increased interleaver length are addressed in the "Sensitivity Analysis" clause below;

- for what concerns handheld reception (Category 2b), in LMS-SU and at the highest satellite EIRP considered, the results indicate that ESR5 fulfillment is below target for all physical configurations. However, Class 2 is able to reach 88 % while Class 1+ MPE-IFEC reaches only 62 %.

2) Uniform Late (UL) and Uniform (U) long physical layer interleavers

The Uniform Late interleaver profile is specifically designed to allow fast zapping in good reception condition. Also, it allows coexistence of Class 1 and Class 2 in the same network. However, these advantages entail some performance penalty. For the targeted 90 % of ESR5 fulfillment, the Uniform Late interleaver losses with respect to Uniform Long interleaver are:

- up to 4 dB in the LMS-SU, for Category 1 terminal at 50 kmph;
- below 2 dB in the LMS-ITS, for Category 1 terminal at 50 kmph;
- about 1,5 dB in LMS-SU, for Category 2b terminal at 3 kmph.

3) Usage of Higher Order Modulation with Lower coding rate

When a constant overall spectral efficiency is considered, several configurations (*Modulation* / PHY-*CodeRate* / IFEC-*CodeRate*) can be envisaged.

With reference to the Class 2, the results for the reference scenarios (cases 22/23-25/26 and 35/36-38/39) are inconclusive since these cases show 100 % fulfil of the ESR5.

The results for Class 1 show a slight advantage for the Lower Order Modulation with a less robust FEC protection, in particular for the LMS-ITS (cases 14 vs. 18 and 21 vs. 24). For the analyzed configurations, the results show that the Higher Order Modulation with Lower Coding Rate obtains equivalent performances only when operating in high ESR5 fulfillment regions.

A.12.2 Detailed results in LMS environments: Reference Cases and Sensitivity analysis

A.12.2.1 Reference Cases results

The results for the Reference Cases and the LMS environments are provided in the 3 tables below, corresponding respectively to the following 3 scenarios:

- table A.26, Scenario (a): Category 1 terminal in LMS-SU, speed of 50 kmph and 63 dBW EIRP Satellite;
- table A.27, Scenario (b): Category 1 terminal in LMS-ITS, speed of 50 kmph and 63 dBW EIRP Satellite;
- table A.28, Scenario (c): Category 2b terminal in LMS-SU, speed of 3 kmph and 68 dBW EIRP Satellite.

The tables give the main parameters of the waveform, in particular the type of Physical Layer interleaver, the code rates at Physical Layer and, when used, the code rates at Link Layer level. The total code rate is the product of these two code rates. The total capacity is the equivalent capacity provided after all FEC decoding and computed at the MPEG2 TS interface. The total interleaving length is about 10 s for all cases.

**Table A.26: ESR5 fulfilment in Scenario (a)
LMS-SU environment, about 10 s of interleaving,
63 dBW EIRP Satellite, Category 1 terminal, speed =50 kmph, State Machine = "ON"**

ID	Waveform configuration						TOTAL capacity (Mbps)	ESR5 fulflilment
	transmission Mode	Modulation	PHY code-rate	Link code-rate	TOTAL code-rate	PHY Interleaver Profile		
27	OFDM	16QAM	1/4	2/3	1/6	S	2.24	88.20%
28	OFDM	16QAM	1/5	-	1/5	U	2.67	100.00%
29	OFDM	16QAM	1/5	-	1/5	UL	2.67	99.00%
30	OFDM	16QAM	2/7	7/12	1/6	S	2.19	93.30%
31	OFDM	QPSK	1/2	2/3	1/3	S	2.24	89.70%
32	OFDM	QPSK	1/3	-	1/3	U	2.22	100.00%
33	OFDM	QPSK	1/3	-	1/3	UL	2.22	100.00%
34	TDM	8PSK	1/3	2/3	2/9	S	2.60	96.70%
35	TDM	8PSK	2/9	-	2/9	U	2.57	100.00%
36	TDM	8PSK	2/9	-	2/9	UL	2.57	100.00%
37	TDM	QPSK	1/2	2/3	1/3	S	2.60	95.00%
38	TDM	QPSK	1/3	-	1/3	U	2.57	100.00%
39	TDM	QPSK	1/3	-	1/3	UL	2.57	100.00%

Only two configurations (ID=27 and ID=31), corresponding to Class 1 and MPE-IFEC, have ESR5 fulfillment slightly below 90 %. When Class 2 is used, performance is above 99 %.

**Table A.27: ESR5 fulfillment in Scenario (b)
LMS-ITS environment, about 10 s of interleaving,
63 dBW EIRP Satellite, Category 1 terminals, speed= 50 kmph, State Machine ="ON"**

ID	Waveform configuration						TOTAL capacity (Mbps)	ESR5 fulflilment
	transmission Mode	Modulation	PHY code-rate	Link code-rate	TOTAL code-rate	PHY Interleaver Profile		
14	OFDM	16QAM	1/4	2/3	1/6	S	2.24	6.80%
15	OFDM	16QAM	1/5	-	1/5	U	2.67	95.50%
16	OFDM	16QAM	1/5	-	1/5	UL	2.67	86.00%
17	OFDM	16QAM	2/7	7/12	1/6	S	2.19	6.40%
18	OFDM	QPSK	1/2	2/3	1/3	S	2.24	13.30%
19	OFDM	QPSK	1/3	-	1/3	U	2.22	100.00%
20	OFDM	QPSK	1/3	-	1/3	UL	2.22	99.00%
21	TDM	8PSK	1/3	2/3	2/9	S	2.60	52.80%
22	TDM	8PSK	2/9	-	2/9	U	2.57	100.00%
23	TDM	8PSK	2/9	-	2/9	UL	2.57	100.00%
24	TDM	QPSK	1/2	2/3	1/3	S	2.60	64.70%
25	TDM	QPSK	1/3	-	1/3	U	2.57	100.00%
26	TDM	QPSK	1/3	-	1/3	UL	2.57	100.00%

Most configurations with Class 2 have an ESR5 fulfillment above 90 %. Configurations using 16QAM OFDM and code rate 1/5 have slightly lower performance because of a higher spectrum efficiency. Results for Class 1 and IFEC show that a 10 s IFEC gives an insufficient protection in LMS-ITS. Results for Class 2 show that for most of the cases, performance is above 99 %.

Table A.28: ESR5 fulfillment for Scenario (c)
LMS-SU environment, with about 10 s of interleaving,
68 dBW EIRP Satellite, Category 2b terminals, speed = 3 kmph, State Machine = "ON"

ID	Waveform configuration						TOTAL capacity (Mbps)	ESR5 fulfilment
	transmission Mode	Modulation	PHY code-rate	Link code-rate	TOTAL code-rate	PHY Interleaver Profile		
72	OFDM	16QAM	1/5	-	1/5	U	2.67	51.00%
73	OFDM	16QAM	1/5	-	1/5	UL	2.67	42.00%
75	OFDM	QPSK	1/3	-	1/3	U	2.22	84.00%
76	OFDM	QPSK	1/3	-	1/3	UL	2.22	75.00%
77	TDM	8PSK	1/3	2/3	2/9	S	2.60	48.90%
78	TDM	8PSK	2/9	-	2/9	U	2.57	87.57%
79	TDM	8PSK	2/9	-	2/9	UL	2.57	84.43%
80	TDM	QPSK	1/2	2/3	1/3	S	2.60	62.10%
81	TDM	QPSK	1/3	-	1/3	U	2.57	87.98%
82	TDM	QPSK	1/3	-	1/3	UL	2.57	86.89%

In all configurations, ESR5 fulfillment is below 90 %. QPSK with PHY coding rate of 1/3 provide performance very close to this target (88 %). Possible improvements are either the use of longer interleavers or the increase of the transmission link margin, through the reduction of the capacity, or the increase of the satellite transmit power:

- Interleaver impact is analyzed for IFEC cases in clause 1.2.2.2 (figure 9).
- Link margin impact is analyzed for Long Physical Interleaver case in clause 1.2.2.2 (figure 10).

To the reference cases reported in the 3 tables above:

NOTE 1: When compared to OFDM QPSK, the Spectrum Efficiency for OFDM-16QAM code rate 1/5 (ID# 15-16, 28-29,72-73), is slightly higher (2,7 Mbps capacity instead of 2,2 Mbps). The rationale of this difference is due to the minimum allowed coding rate, equal to 1/5. This reflects both in an higher bit-rate and in a slight performance degradation.

NOTE 2: The total channel capacity refers to the aggregate bit rate of the available services; target net bit-rate per service is roughly 280 kbps, then, depending on the system configuration, the number of services can be either 8 or 9.

NOTE 3: In general it can be noticed that even given the same TOTAL EFFICIENCY, the system configurations can give slightly different net bit rates. The reason is related to a different Physical Layer configuration (modulation and/or coding rate) which could end up in a slightly different waveform configuration (e.g. CU-padding in the SH-Frame composition).

NOTE 4: The exact C/N for each Scenario is reported in the link-budgets in clause 1.7.6. However, the results reported here already take into account the implementation losses, i.e: 1,1 dB for OFDM QPSK, 0,5 dB for TDM QPSK, 1,5 dB for OFDM 16QAM and 1 dB for TDM 8PSK.

A.12.2.2 Sensitivity Analysis

This clause provides with performances in the LMS environments when different system configurations with respect to the Reference Cases are considered, relaxing or changing some constraints as appropriate.

As noted, the performance of Class 1 receiver using MPE-IFEC with interleaving length up to 10 s do not always meet 90 % ESR5 fulfillment. Therefore, a sensitivity analysis to the length of protection period and to the code-rate is performed with the aim of finding the optimal IFEC parameters in each scenario, keeping the same reference C/N. The results are presented in a graphical format, reporting the minimum required (Link-Layer) interleaver length as a function of the supported aggregated net-bit rate. It is important to notice that, for each graphic, the curves represent the boundary conditions: all the system configurations which belong to the left-bottom part of the plane are viable solutions. Indeed, for a given interleaver length, all the bit-rates up to the point reported in the curve are guaranteed with 90 % of ESR5 fulfillment as minimum.

On the contrary, Class 2 receivers in Category 1 terminals generally exceed 90 % ESR5 fulfillment. Therefore, a parametric analysis of the ESR5 as a function of the C/N is performed for a set of cases when Class 2 is considered. The same C/I as reported in the reference cases is kept. Influence of the type of long Physical Layer interleaver (Uniform versus Uniform Late) is also addressed.

The reader is reminded that he should be cautious in comparing OFDM versus TDM performances from the results presented, due to the guard interval value chosen for OFDM, different net bit rates, implementation losses and noise bandwidth assumption made.

A.12.2.3 Vehicular reception in LMS environments

A.12.2.3.1 Vehicular Reception in LMS-SU and SH-A Waveform (OFDM)

The simulation performances reported in this clause have been obtained in LMS-SU environment and for SH-A waveform with a 63 dBW EIRP satellite (reference cases IDs 27 to 33 in table A.26). These cases are considered for vehicular applications; therefore a typical speed of 50 kmph has been analyzed.

Figure A.32 shows the sensitivity analysis for Class 1. The supported bit rate for the reference case is indicated by the dash line. Under these working conditions, it is possible to increase the supported bit rate while satisfying 90 % ESR5 fulfillment. This is achieved by increasing the MPE-IFEC code rate and interleaving depth.

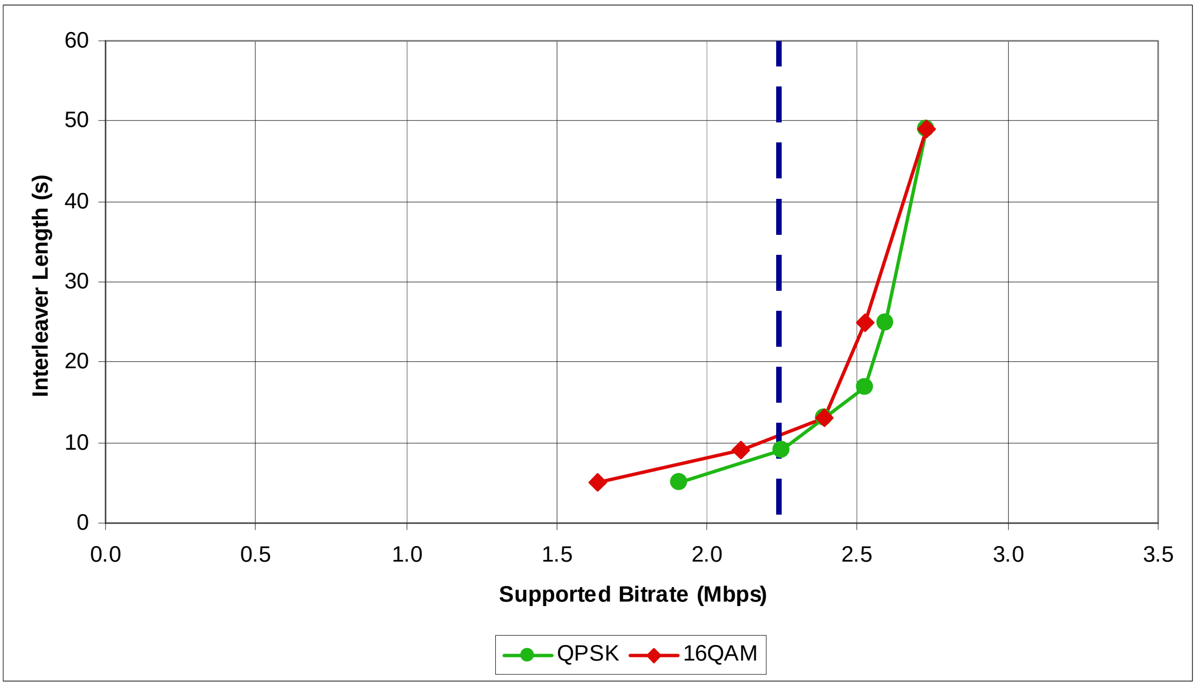


Figure A.32: SH-A, Class 1 – QPSK 1/2 and 16QAM 1/4 – LMS-SU - 50 kmph - 63 dBW EIRP Satellite Different Link-Layer configuration fulfilling the ESR(5) criterion at 90 % (corresponding reference cases IDs 27 to 31)

Figure A.33 shows the sensitivity analysis for Class 2. The C/N setting used for the reference case was 11,2 dB (shown by the vertical dashed line). The required C/N to guarantee the 90 % fulfillment of ESR5 is found to be below 5 dB, i.e. 7 dB below the reference configuration setting.

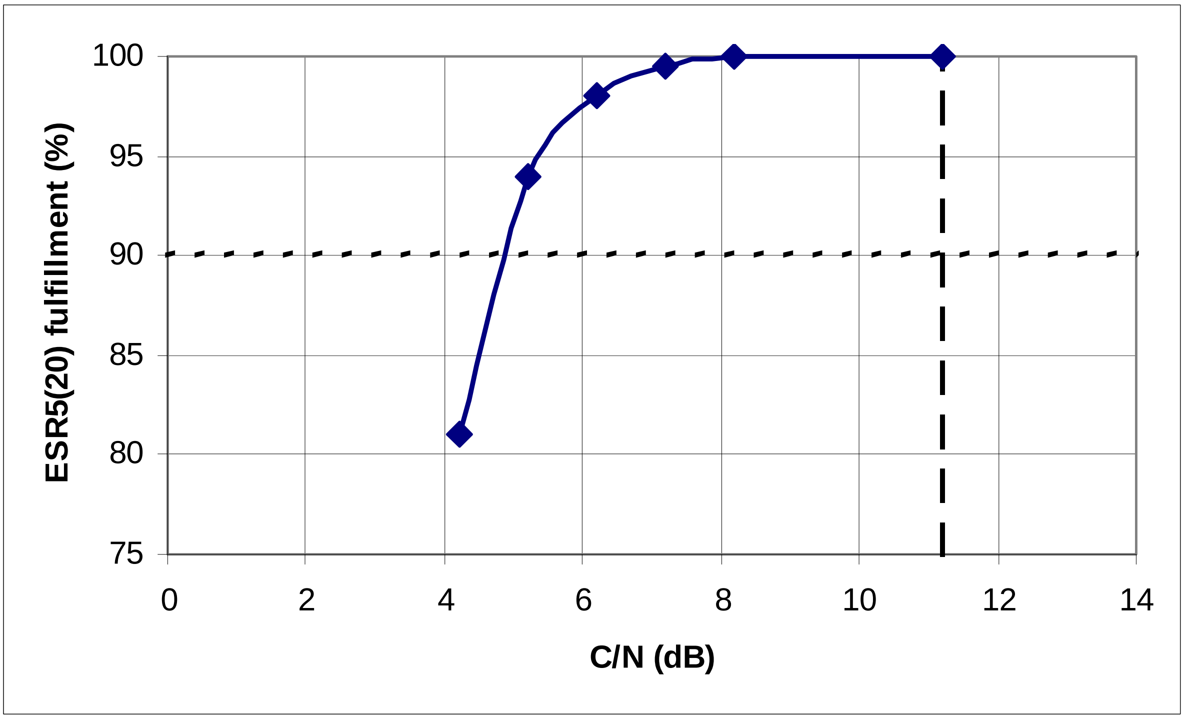


Figure A.33: SH-A, Class 2: Uniform Long Interleaver Profile – QPSK 1/3 – LMS-SU - 50 kmph Sensitivity Analysis to C/N value for an OFDM modulation (corresponding reference cases IDs 31 and 32)

A.12.2.3.2 Vehicular Reception in a LMS-SU and SH-B Waveform (TDM)

The simulation performances reported in this clause have been obtained in an LMS SU channel and for a SH-B waveform with a 63 dBW EIRP satellite (reference cases IDs 34 to 39 in table A.26). These cases are considered for vehicular applications; therefore a typical speed of 50 kmph has been analyzed.

Figure A.34 shows the sensitivity analysis for Class 1. The same conclusions as for the corresponding case of SH-A can be drawn.

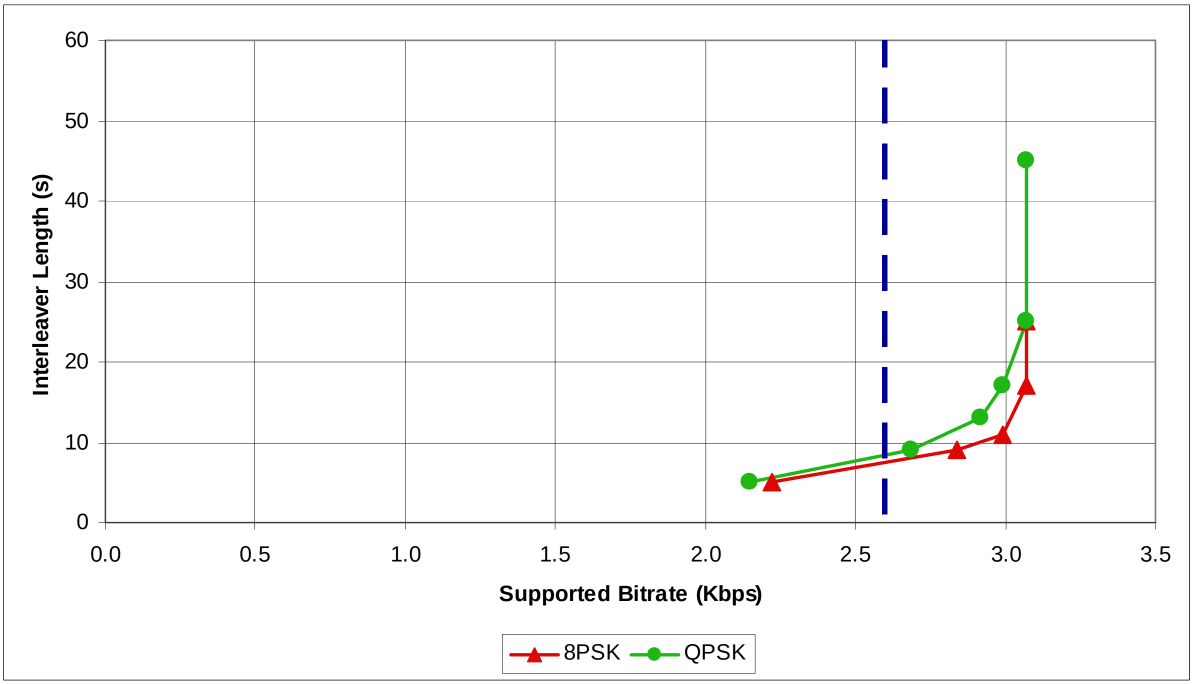


Figure A.34: SH-B, Class 1: – QPSK ½ and 8PSK1/3 - LMS-SU - 50 kmph - 63 dBW EIRP Satellite Different Link-Layer configuration fulfilling the ESR(5) criterion at 90 % (corresponding reference cases IDs 34 to 37)

Figure A.35 shows the sensitivity analysis for Class 2. The C/N setting used for the reference case was 12,3 dB (shown by the vertical dashed line). The required C/N to guarantee the 90 % fulfillment of ESR5 is found to be below 4 dB, i.e. 8 dB below the reference configuration setting.

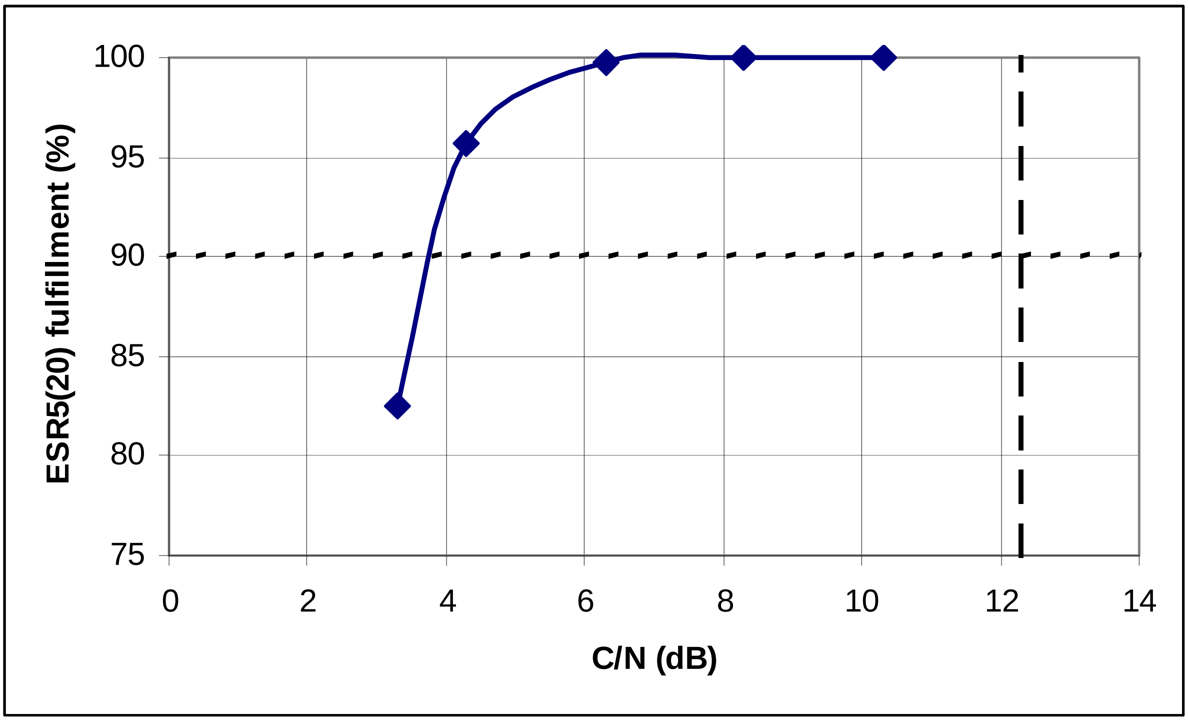


Figure A.35: SH-B, Class 2: with Uniform Long Interleaver Profile – QPSK 1/3 – LMS-SU - 50 kmph Sensitivity Analysis to C/N value for an OFDM modulation (corresponding reference cases IDs 37 and 38)

A.12.2.3.3 Vehicular Reception in a LMS ITS and DVB SH-A Waveform (OFDM)

The simulation performances reported in this clause have been obtained in LMS-ITS environment and for SH-A waveform with a 63 dBW EIRP satellite (reference cases IDs 14 to 20 in table A.27). These cases are considered for vehicular applications; therefore a typical speed of 50 kmph has been analyzed.

As can be seen in figure A.36, in LMS-ITS, an SH-A Class 1 receiver cannot reach the target capacity at 63 dBW satellite EIRP for the reference case (represented as vertical dashed line).

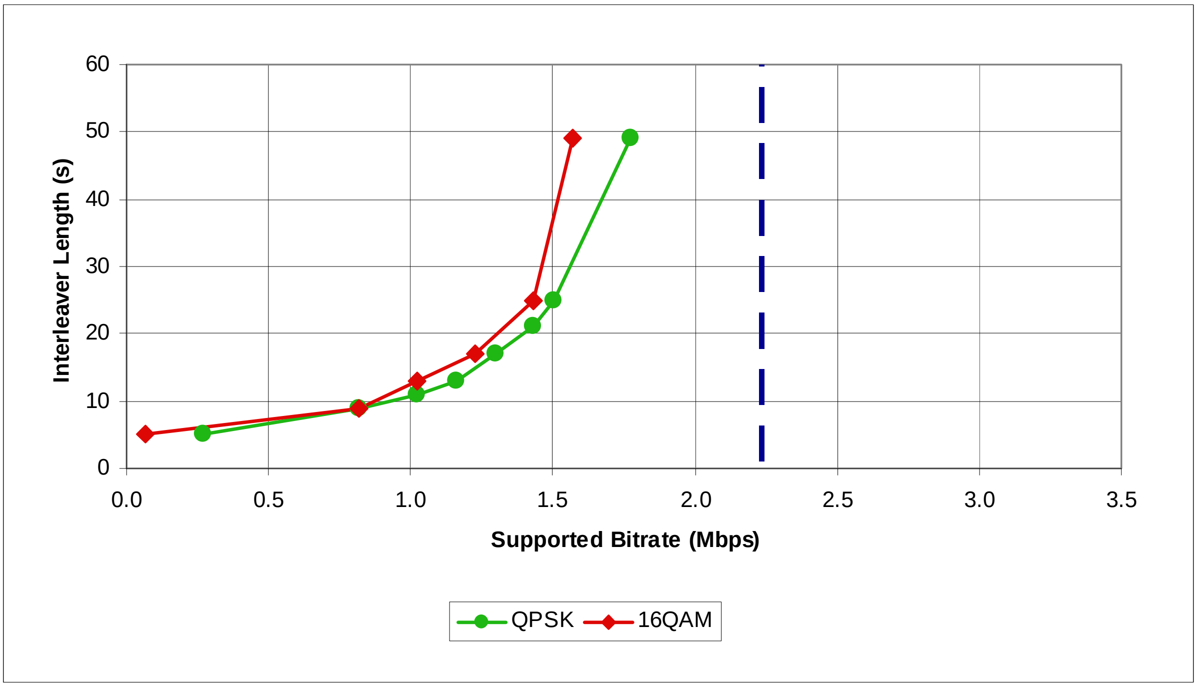


Figure A.36: SH-A, Class 1: - 16QAM 1/4 and QPSK 1/2– LMS-ITS - 50 kmph - 63 dBW EIRP Satellite Different Link-Layer configuration fulfilling the ESR(5) criterion at 90 % (corresponding reference cases IDs 14 to 18)

Figure A.37 shows the sensitivity analysis for Class 2. The C/N setting used for the reference case was 11,2 dB (shown by the vertical dashed line). The required C/N to guarantee the 90 % fulfillment of ESR5 is found to be 9 dB, i.e. 2 dB below the reference configuration setting.

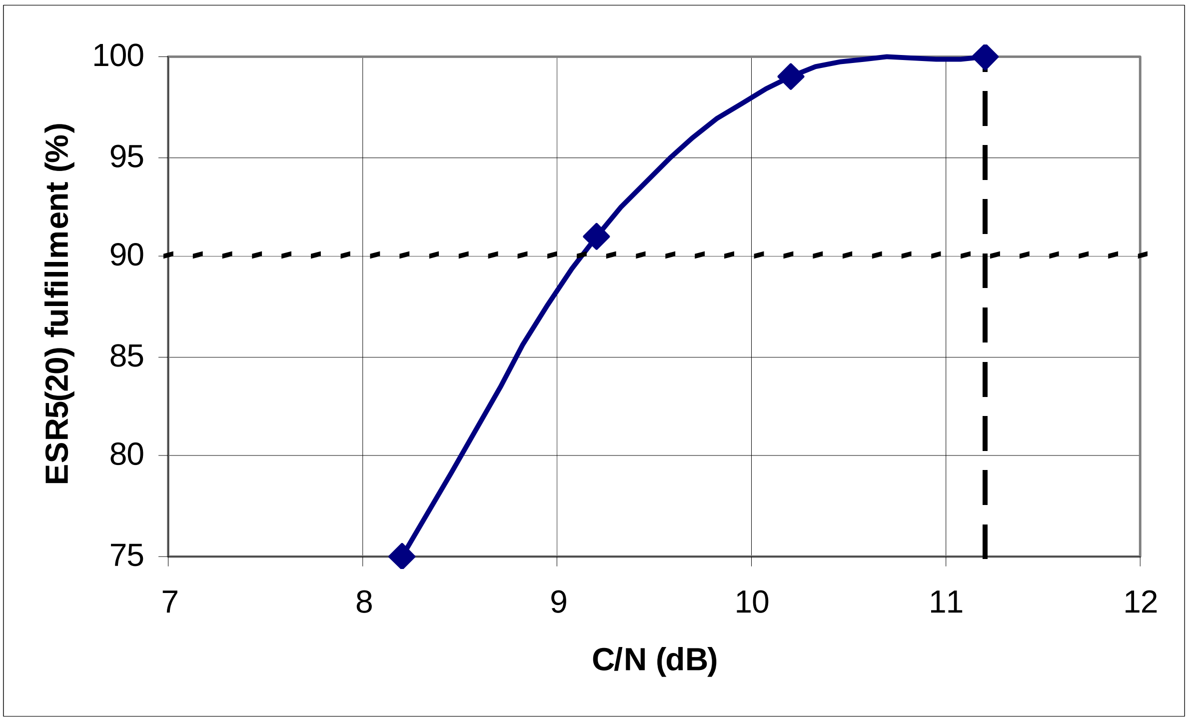


Figure A.37: SH-A, Class 2 with Uniform Long Interleaver Profile- QPSK 1/3 – LMS-ITS - 50 kmph Sensitivity Analysis to C/N value for a OFDM modulation (corresponding reference cases IDs 18 and 19)

A.12.2.3.4 Vehicular Reception in a LMS ITS and DVB SH-B Waveform (TDM)

The simulation performances reported in this clause have been obtained in an LMS ITS channel and for a SH-B waveform with a 63 dBW EIRP satellite (reference cases IDs 21 to 26 in table A.27). These cases are considered characteristic for vehicular applications; in particular a typical speed of 50 kmph has been analyzed.

From figure A.38, it can be noticed that a SH-B Class 1 receiver is able to guarantee an aggregate traffic slightly lower than the target capacity for the reference case (vertical dashed line in the following figure) up to 2,5 Mbps with roughly 25 s of interleaver length at the reference C/N of 12,8 dB. The performances for the QPSK modulation (with a code rate of 1/2 at physical layer) are close to those for an 8PSK modulation (with a code rate of 1/3 at physical layer). Indeed, under this condition, the advantage for the QPSK for the reference 10 s of interleaver length is limited to about 10 % in terms of bit-rate or interleaver-length.

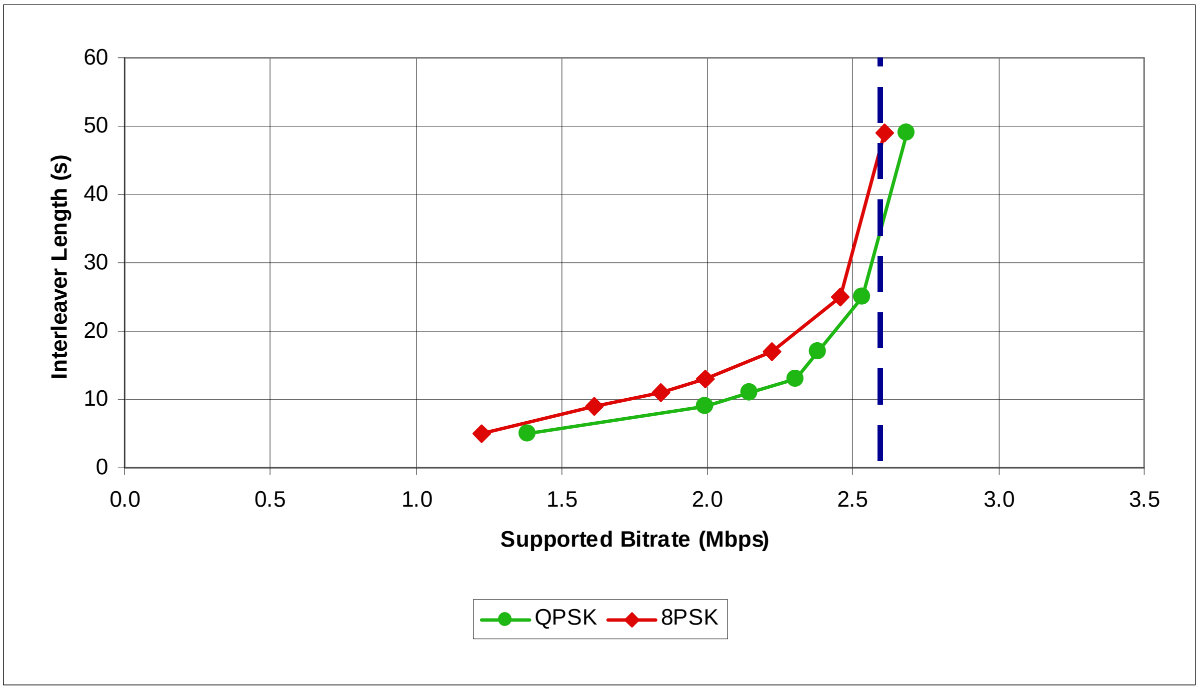


Figure A.38: TDM, Class 1 – QPSK ½ and 8PSK 1/3 – LMS-ITS - 50 kmph - 63 dBW EIRP Satellite Different Link-Layer configuration fulfilling the ESR(5) criterion at 90 % (corresponding reference cases IDs 21 to 24)

Figure A.39 shows the sensitivity analysis for Class 2. The C/N setting used for the reference case was 12,3 dB (shown by the vertical dashed line). The required C/N to guarantee the 90 % fulfillment of ESR5 is found to be 8 dB, i.e. 4 dB below the reference configuration setting.

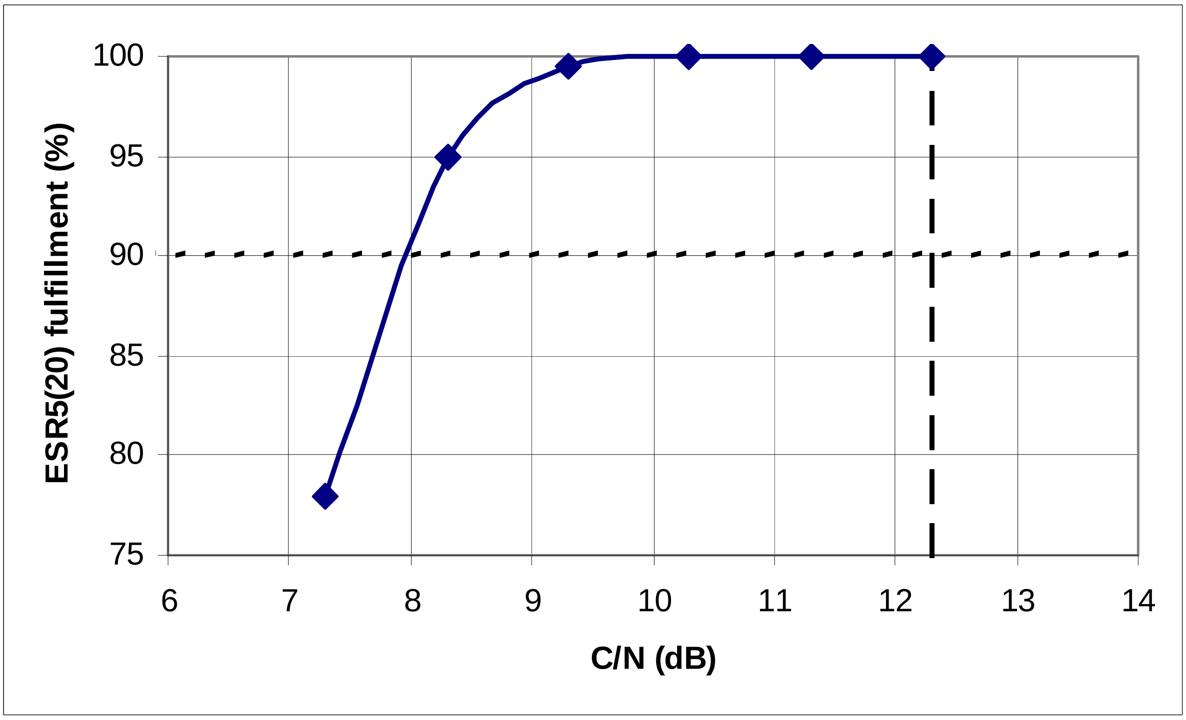


Figure A.39: SH-B, Class 2 with Uniform Long Interleaver Profile - QPSK – LMS-ITS - 50 kmph Sensitivity Analysis to C/N value for a TDM modulation (corresponding reference cases IDs 24 and 25)

A.12.2.3.5 Category 2b reception in LMS-SU at 68 dBW satellite EIRP

The simulation performances reported in this clause have been obtained in an LMS SU channel with a 68 dBW EIRP satellite (reference cases IDs 72 to 82 in table A.28). These cases are considered for handheld applications; therefore a typical speed of 3 kmph has been analyzed.

As can be seen in figure A.40, in LMS-ITS, an SH-A Class 1 receiver cannot reach the target capacity (represented as vertical dashed line) at 68 dBW satellite EIRP.

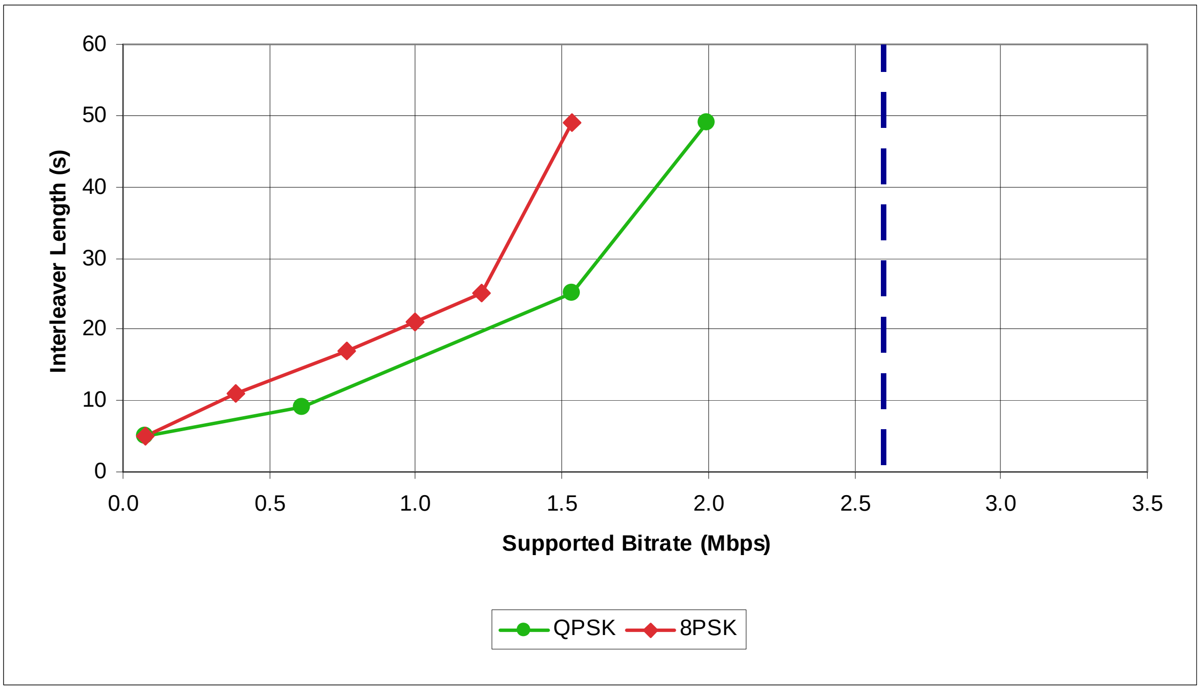


Figure A.40: SH-B, Class 1: – QPSK ½ and 8PSK 1/3 – LMS-SU - 3 kmph - 68 dBW EIRP Satellite Different Link-Layer configuration fulfilling the ESR(5) criterion at 90 % (corresponding reference cases IDs 77 and 78)

Figures A.41 and A.42 give the sensitivity analysis for Class 2. The C/N has been increased with respect to the reference value of 5,1 dB and 6,2 dB for the reference OFDM and TDM cases respectively. The required C/N to guarantee the 90 % fulfillment of ESR5 is increased between 8 dB and 9 dB.

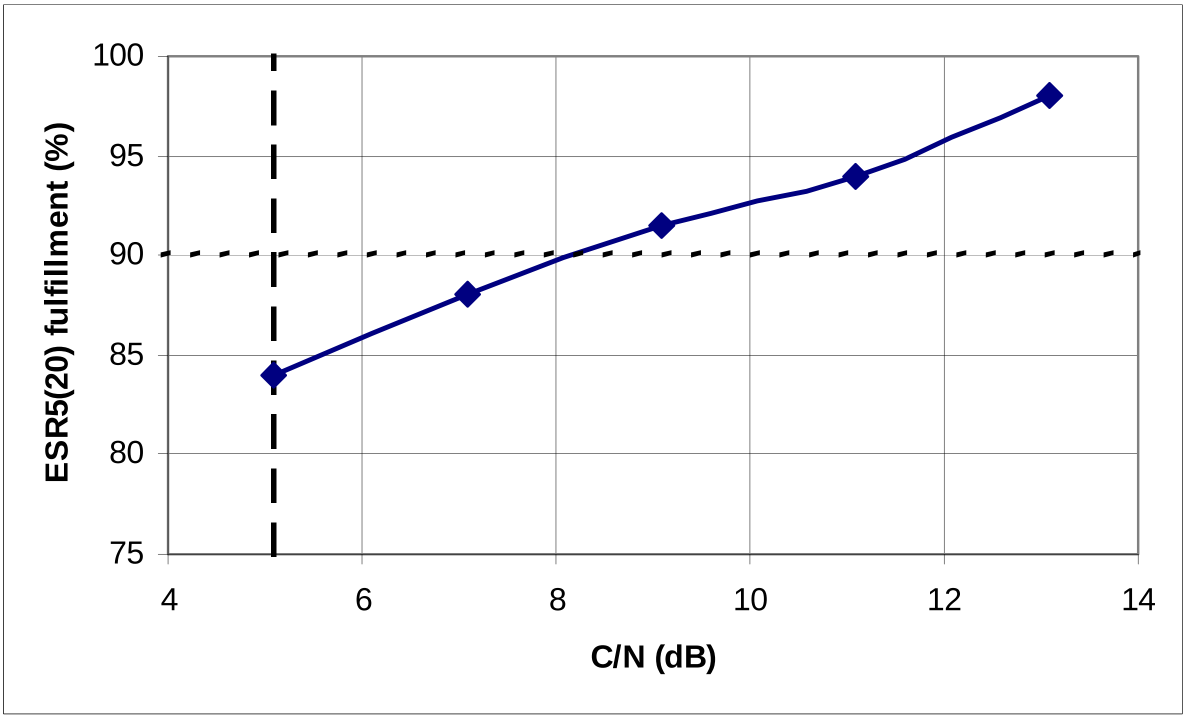


Figure A.41: SH-A, Class 2: Uniform Interleaver Profile – QPSK 1/3 – LMS-ITS - 3 kmph Sensitivity Analysis to C/N value for a TDM modulation (corresponding reference cases ID 75)

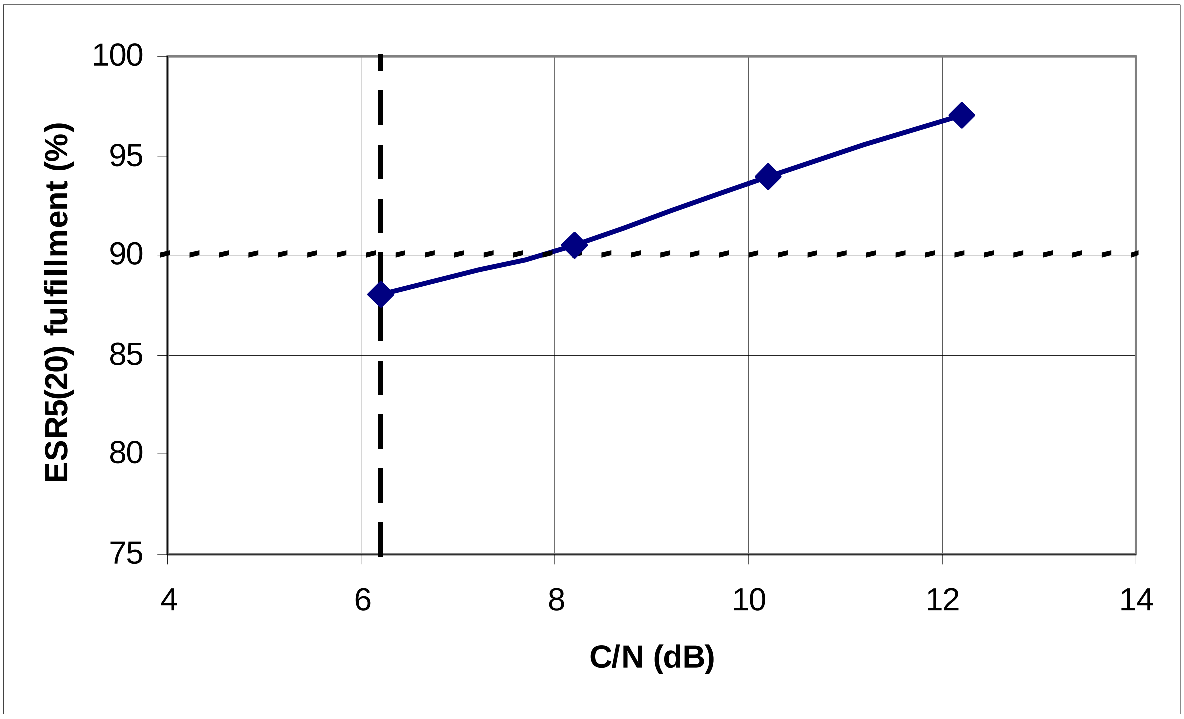


Figure A.42: SH-B, Class 2: with Uniform Interleaver Profile - QPSK – LMS-ITS - 3 kmph Sensitivity Analysis to C/N value for a TDM modulation (corresponding reference cases IDs 80 and 81)

A.12.2.4 Uniform Late (UL) and Uniform (U) long interleavers

This clause provides with performances comparison between the interleaver profiles Uniform Long (U) and Uniform Late (UL) as defined in clause A.4 and used for the simulation campaign for the Class 2 receiver performance evaluation. Only SH-B waveform is considered, but the results are considered of general validity and applicable also to SH-A profile.

Both interleaver profiles show excellent performances in the reference system configurations, if typical satellite EIRP is considered and vehicular terminals Category 1 are addressed (e.g. cases 19 and 20 or 25 and 26 in table A.27). When the system is operating in such high ESR5 fulfillment regions, the performances of the two different profiles are almost equivalent and it is also possible to exploit the fast-zapping feature allowed by the UL profile.

On the contrary, when the target 90 % of ESR5 fulfillment is considered, the Uniform Long interleaver profile is characterized by a lower C/N working point and an optimization in terms of satellite EIRP, terminal RF performances and spectral efficiency is possible. In particular, this clause quantifies this advantage for the reference scenarios.

Three different cases are analyzed in this clause:

a) LMS-SU scenario at 50 kmph for vehicular reception (figure A.43);

b) LMS-ITS scenario at 50 kmph for vehicular reception (figure A.44);

c) LMS-SU scenario at 3 kmph for handheld reception (figure A.45).

With reference to a), the Uniform Long interleaver meets the 90 % ESR5 fulfillment at roughly 4 dB of C/N with a saving of 4 dB with respect the Uniform Late. This represent the worst case within the selected cases, indeed this difference reduces below 1 dB for both b) and c).

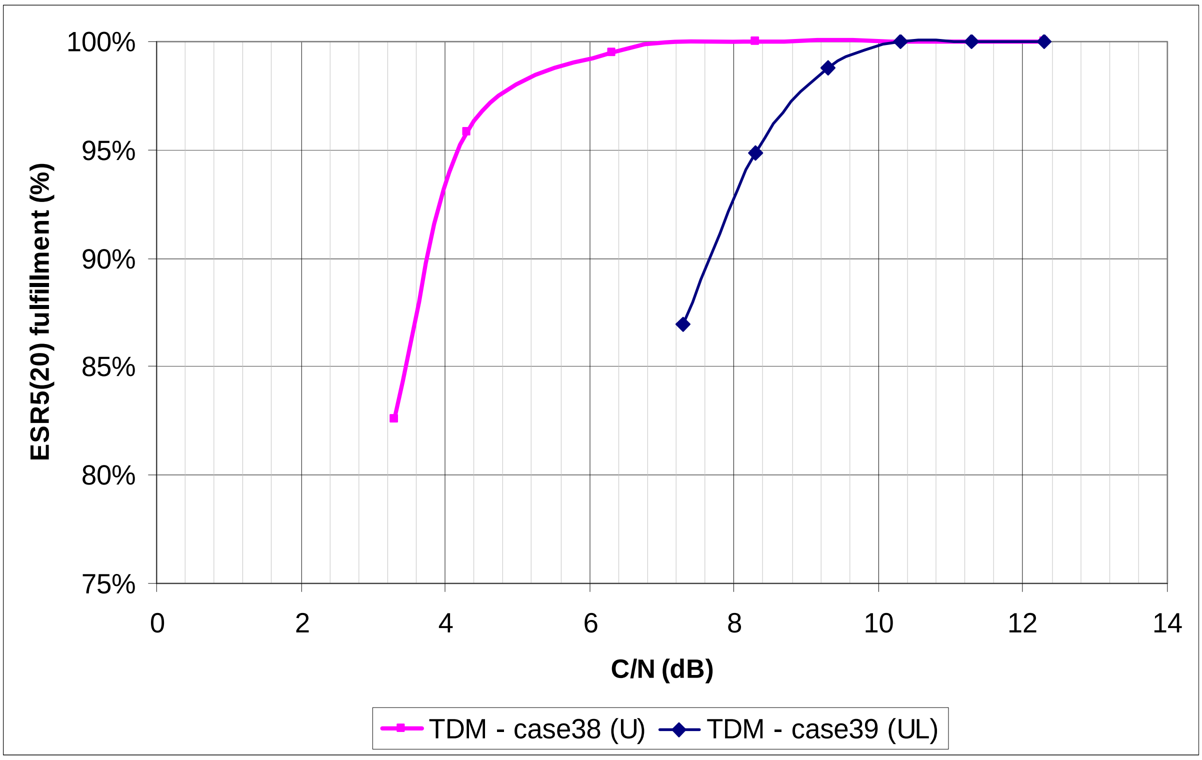


Figure A.43: TDM, Class 2: - QPSK 1/3 – LMS-SU - 50 kmph Sensitivity Analysis to C/N value for a TDM modulation (corresponding reference case IDs 38 and 39)

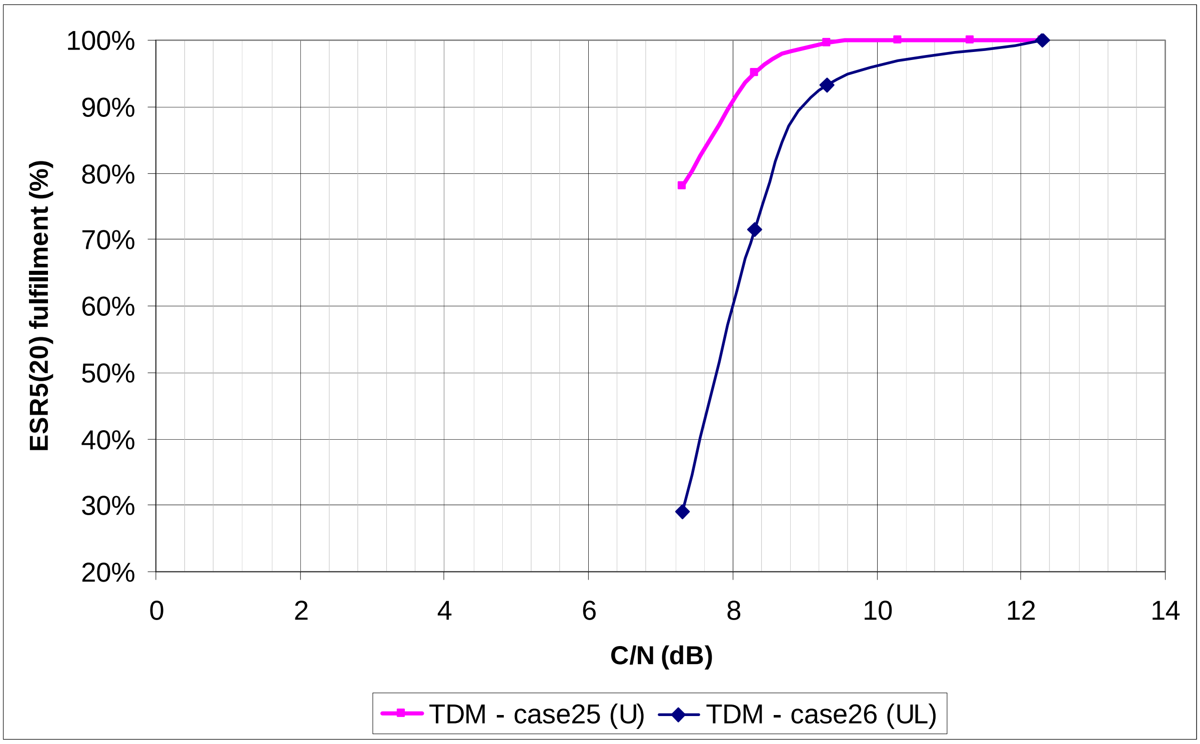


Figure A.44: SH-B, Class 2: - QPSK 1/3 – LMS-ITS - 50 kmph Sensitivity Analysis to C/N value for a TDM modulation (corresponding reference case IDs 25 and 26)

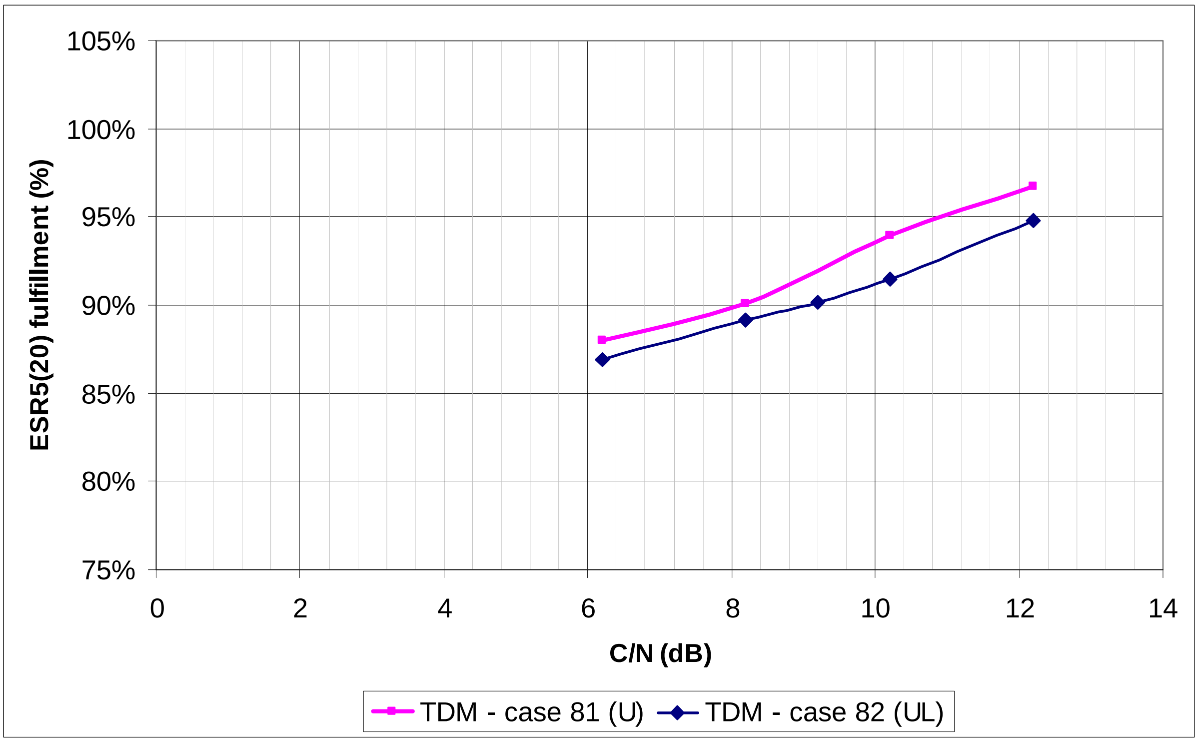


Figure A.45: TDM, Class 2: - QPSK 1/3 – LMS-SU - 3 kmph Sensitivity Analysis to C/N value for a TDM modulation (corresponding reference case IDs 81 and 82)

A.12.3 Reception in TU6 environment

TU6 propagation model is used as reference for characterization of the small scale fading effects encountered in terrestrial environment reception. Thus, it only applies to the OFDM waveform. Performance are defined in terms of required C/N, to be used for Terrestrial Network Planning (clause 11).

Because of the hybrid structure of the DVB-SH system, two cases of terrestrial transmission are possible:

- terrestrial retransmission of the satellite signal of an SH-A (SFN) system, where in particular long interleavers, at Physical Layer level, or at IFEC level, may be implemented;
- dedicated terrestrial transmission, in case of retransmission of the satellite content for SH-A (non-SFN) or SH-B system, or in case of terrestrial-only local content broadcasting.

This clause gives the simulation performance in a TU6 channel, for an OFDM waveform, with a terminal speed of 3 kmph and 50 kmph. They apply to a reference service of 280 kbps. Overall capacity is calculated on the basis of a 1/8 guard interval, as simulation results in TU6 are not sensitive to guard interval larger than 1/8.

A.12.3.1 FER and ESR5 relationship

The receiver threshold performance is a function of the quality criteria. Two criteria are commonly used in broadcasting systems, which are the FER and the ESR5.

From the various simulations performed with the DVB-SH waveform in TU6, the following relationship between ESR5 and FER is observed, whatever the code rate and the speed of the receiver.

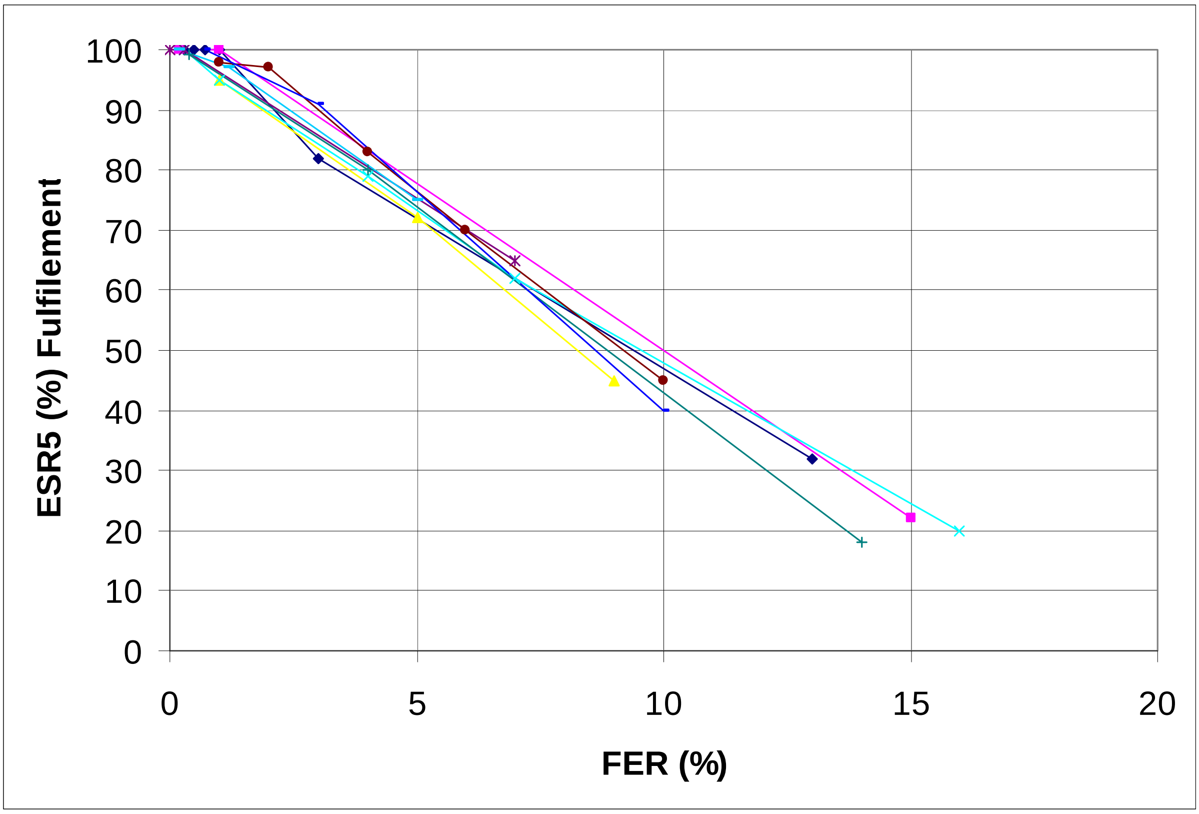


Figure A.46: ESR5 fulfillment performance versus FER performance for TU6 environment, for various speed, code rate and modulation

Each plot on the figure refers to a different order of modulation (QPSK or 16QAM) and code rate (from ½ to 1/5).

From this plot, 1-ESR5 = 99 % is equivalent to an FER = 1 %. In terms of C/N threshold difference between FER = 5 % and 1-ESR5 = 99 % (or 1 % FER), is typically less than 0,5 dB at 50 kmph, and is about 1 dB at 3 kmph.

A.12.3.2 FER performance in TU6

Table A.29 gives the FER=5 % performance of the DVB-SH OFDM waveform, using only a short Physical interleaver of 200 ms, and no IFEC. These results are based on the compiling of several simulation runs and tools.

Table A.29: FER performance in TU6

Code rate PHY	Capacity (Mbps)	C/N @ FER=5 % (dB)
TU6 OFDM QPSK - 50 kmph		
1/3	2,5	1,2
1/2	3,7	4,3
TU6 OFDM 16QAM - 50 kmph		
1/5	3,0	2,5
1/4	3,7	4,1
1/3	4,9	6,2
1/2	7,5	9,5
TU6 OFDM QPSK - 3 kmph		
1/3	2,5	3,3
1/2	3,7	5,8
TU6 OFDM 16QAM - 3 kmph		
1/5	3,0	4,8
1/4	3,7	5,6
1/3	4,9	7,8
1/2	7,5	11

From table A.29, it may be noted that:

- thresholds at 3 kmph are typically 1,5 dB worse than those at 50 kmph;
- for the same spectrum efficiency, 16QAM performance is slightly better than QPSK, thank to the lower available coding rate. For real receivers, this must be tempered by the implementation losses of the receiver, which could be higher in 16 QAM than in QPSK.

A.12.3.3 ESR5 Performance in TU6

Three sets of configurations are considered:

- short Physical layer interleaving, and no IFEC;
- long Physical layer interleaving, and no IFEC;
- short Physical layer interleaving, and IFEC.

In the following performance description, they are arranged in terms of Physical interleaver length, and IFEC definition. The figure at the end of the clause summarizes all these performance.

Performance versus Physical Interleaver length

Short and Long Physical layer interleaving configurations are analyzed first. No IFEC is considered. Short Interleaver is 200 ms long, and the Long Interleaver is 10 s long, with half of the interleaver paths within 200 ms.

The following tables give the ESR5 simulation performance in a TU6 channel, for an OFDM waveform, with a terminal speed of 3 kmph and 50 kmph. In both interleaver configurations, the same capacity is provided, as no IFEC is used.

The improvement on the C/N between Short Physical interleaver and long Physical interleaver is in the range of 0,5 dB at 50 kmph, and 1 dB at 3 kmph, which points to the conclusion that the TU6 channel can already be coped quite well with short interleaving only.

Table A.30: ESR5 performance of UL versus S, in TU6

Code Rate PHY	Interleaver PHY	ID	Capacity (Mbps)	C/N (dB)	Interleaving length (s)	1-ESR5 (%)
TU6 OFDM 16QAM - 3 kmph						
1/3	S	-	4,9	8,5	0	100
1/3	UL	-	4,9	7	10	99
¼	S	2	3,7	6,5	0	100
¼	UL	2'	3,7	5	10	98,5
1/5	S	4'	3,0	5,5	0	99
1/5	UL	4	3,0	4	10	100
TU6 OFDM 16QAM - 50 kmph						
1/3	S	-	4,9	6,5	0	100
1/3	UL	-'	4,9	6	10	100
¼	S	1	3,7	4,5	0	100
¼	UL	1'	3,7	4	10	100
1/5	S	3'	3,0	3	0	100
1/5	UL	3	3,0	3	10	100
TU6 OFDM QPSK - 3 kmph						
½	S	8	3,7	6,5	0	99,00
½	UL	8'	3,7	5,5	10	100,00
1/3	S	10'	2,5	3,5	0	99,00
1/3	UL	10	2,5	2,5	10	100
TU6 OFDM QPSK - 50 kmph						
½	S	7	3,7	4,5	0	100
½	UL	7'	3,7	4	10	99
1/3	S	9'	2,5	1,5	0	100
1/3	UL	9	2,5	1	10	100

Performance versus IFEC definition

Configurations with different IFEC setting are analyzed. In any case, only Short Physical interleaver is considered. Short Interleaver is 200 ms long.

The following tables give the ESR5 simulation performance in a TU6 channel, for an OFDM waveform, with a terminal speed of 3 kmph and 50 kmph. In most configurations, the same capacity after IFEC decoding is provided. Related effect on the required C/N of IFEC code rate together with IFEC interleaving length (B+S) is analyzed. The QoS target is 99 % of ESR5 fulfillment.

At 3 kmph, which is somewhat representative of handheld reception in terrestrial environment, it shows that IFEC configurations provides similar performance to configurations without IFEC:

- in case of QPSK, same capacity is achieved with code rate ½ and IFEC and with code rate 1/3 and no IFEC. The length of the IFEC interleaver allows to improve the C/N threshold;
- in case of 16QAM, slightly higher capacity is achieved with code rate 1/5 and no IFEC, at the expense of a larger required C/N. This is due to the fact that the selected 16QAM 1/5 configuration is deliberately more spectrum efficient. A setting with adequate spectrum efficiency (code rate 2/9 for example) would provide results closer to cases with ¼ code rate at physical layer and IFEC.

At 50 kmph, compared to 3 kmph, configurations without IFEC improve more than configurations with IFEC. However, low speed conditions are the most severe, and make these improvement not a driving parameter in a system definition.

So, in a SH-A SFN configuration, IFEC settings defined for the satellite link will not penalize the reception in terrestrial environment.

In case of dedicated terrestrial transmission, the introduction of IFEC will offer fine tuning system capability. Typically, addition of IFEC will provide a control of the quality of each service based on high flexibility capability.

Table A.31

Code Rate PHY	Interleaver PHY	ID	Code Rate IFEC	Capacity (Mbps)	C/N (dB)	Interleaving length (s)	1-ESR5 (%)
TU6 OFDM QPSK - 3 kmph							
1/3	S	10'	-	2,5	4	0	99,00
1/2	S	8	0,67	2,5	5,5	4 to 10	99,50
1/2	S	8	0,67	2,5	5	10 to 16	99,20
1/2	S	8	0,67	2,5	4,5	20 to 30	99,00
TU6 OFDM 16QAM - 3 kmph							
1/5	S	4'	-	3,0	6	0	99,00
1/4	S	2	0,67	2,5	5	10 to 16	> 99,20
1/4	S	2	0,60	2,2	5	10	99,30
1/4	S	2	0,42	1,5	4	20 to 30	> 99,10

Table A.32

TU6 OFDM QPSK - 50 kmph							
1/3	S	9'	-	2,5	1,5	0	100,00
1/2	S	7	0,67	2,5	4	10	100,00
1/2	S	7	0,43	1,6	3,5	20 to 30	100,00

IFECSynthetic overall Performance

The following figure gives a synthetic overview of the performance achieved at ESR(5) @ 99 % criteria for all above configurations, in TU6 at 3 kmph:

- short Physical Interleaver only configurations are round market points, on the upper right end of the lines;
- IFEC configurations are along the lines;
- long Physical Interleaver configurations are triangle marked points;

- the squared marked point corresponds to a configuration using MPE-FEC.

DVB-SH includes 8 code rate settings at physical layer, from 1/5 to 2/3. Only two of them are presented in figure A.47 with QPSK modulation, and 3 with 16QAM modulation. The full set of code rates, together with the flexibility of IFEC, allows to provide a very fine system capacity optimization capability.

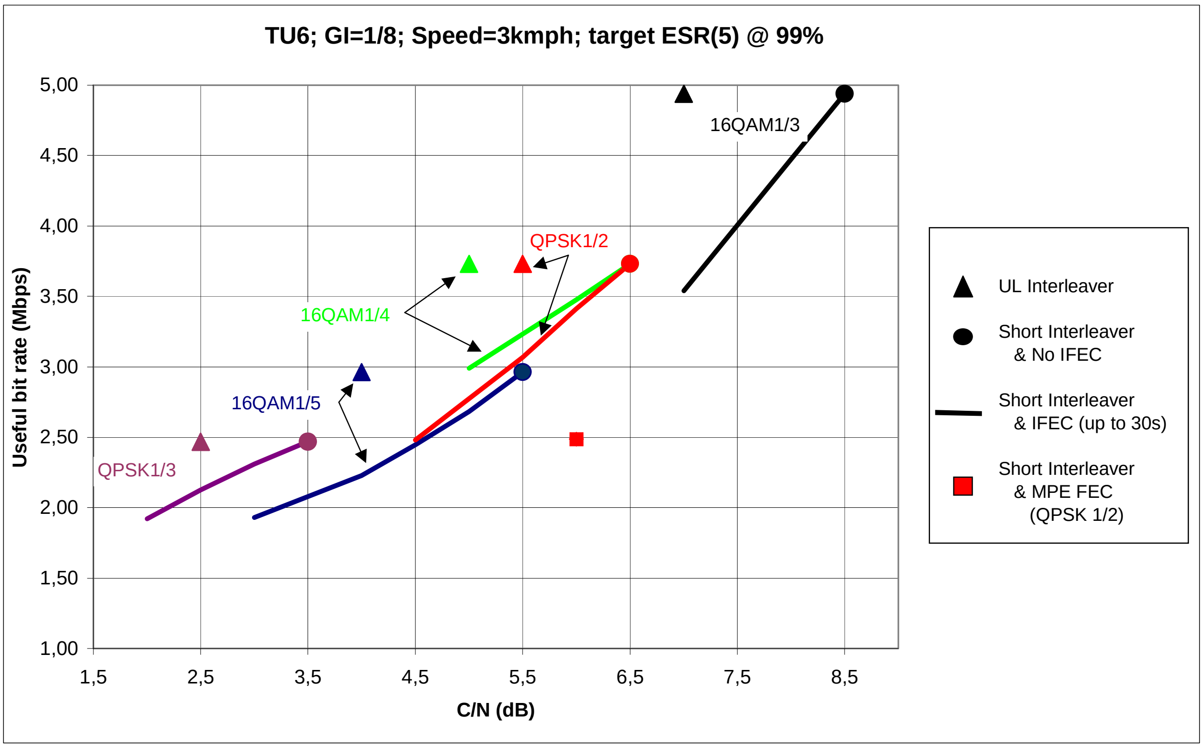


Figure A.47

Figure A.48 gives a synthetic overview of the performance achieved at ESR(5) @99 % criteria for all above configurations, in TU6 at 50 kmph.

- short Physical Interleaver only configurations are round market points;
- long Physical Interleaver configurations are triangle marked points;
- IFEC configuration is along the line.

At 50 kmph, performance at physical layer level changes very quickly with C/N. Available dumps have allowed to only draw the IFEC performance for QPSK 1/2.

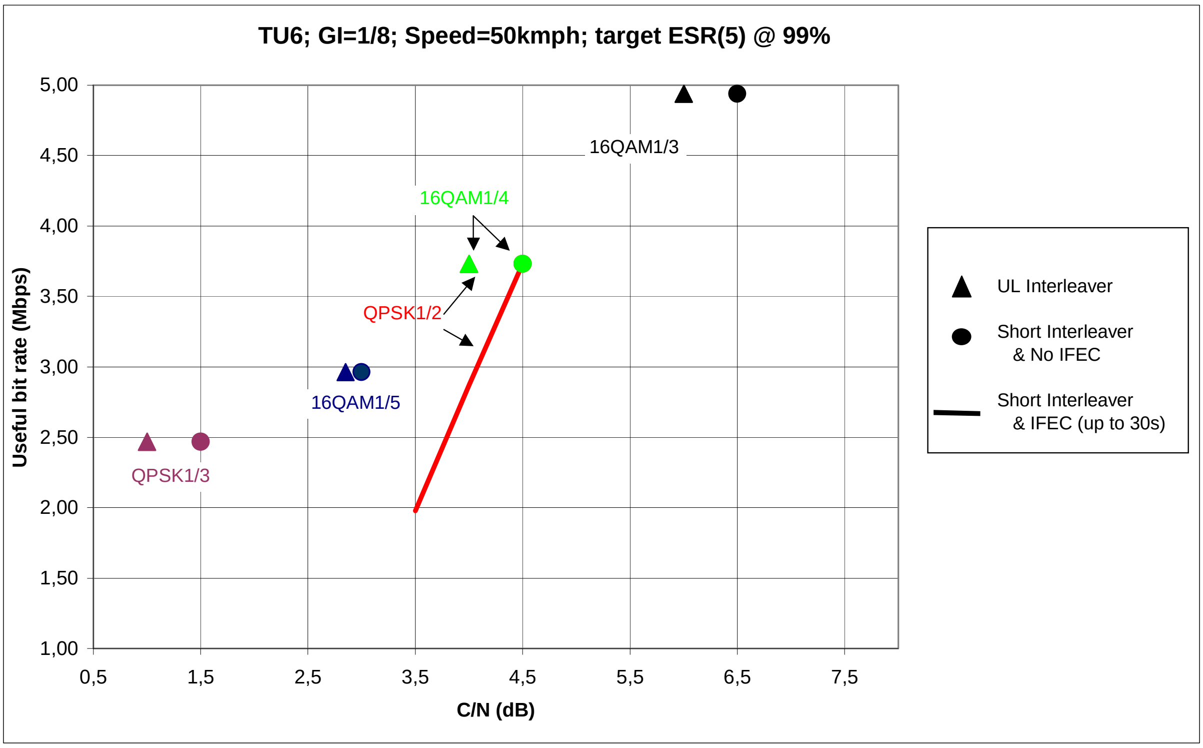


Figure A.48

A.13 MPE-IFEC specifications

The present clause is a reproduction of a DVB-SSP working document and is provided to facilitate the understanding of the clause A results. Therefore, the entire present clause is for information only.

1 Introduction (informative)

MPE-IFEC is introduced to support reception in situations of long duration erasure on the MPE section level spanning several consecutive time slice bursts. Such erasure situations may for example occur on satellite mobile channels (LMS: land mobile satellite) without any terrestrial repeaters in the vicinity: obstacles may hinder direct satellite reception and induce losses of several successive bursts. For example, with an MPE-IFEC protection where about 30 % of TS data are allocated to parity overhead computed over 10 successive bursts, it is possible to compensate up to 3 successive complete burst losses whereas recovery of a complete burst loss with DVB-H MPE-FEC protection would not be possible. Such erasure situation may also occur in terrestrial networks so that MPE-IFEC may be useful in other channels than LMS.

The MPE-IFEC protection is computed over several successive datagram bursts, as opposed to MPE-FEC where the computation is performed on a single datagram burst. This multi-burst protection is enabled by an enlargement of the encoding matrix to sizes greater than one burst (an IFEC matrix is filled not by one burst as in MPE-FEC but by several successive bursts), by a parallelization of the encoding mechanism (instead of using only one matrix, the data are distributed to a number of parallel matrices equal to *B*) or by a combination of both principles. The datagrams themselves are sent in MPE sections without any modification compared to EN 301 192 [9], clause 9.6. The resulting parity may also be spread over several bursts instead of one single burst in the MPE-FEC case: each burst contains parity coming from *S* matrices.

An overview of the link layer operations, especially the MPE-IFEC, in the case of DVB-SH is presented in figure 1. Datagram bursts of variable size (in terms of number of bytes and/or number of datagrams) are used as input and mapped by an ADST (Application Data Sub Table) function on to the ADTs (Application Data Tables) of up to *M* parallel encoding matrices. The IFEC burst collects all MPE-IFEC sections from iFDTs (IFEC Data Tables, the IFEC parity symbols for an ADT) of several encoding matrices. An MPE-IFEC section is comprised of a header, the data from multiple columns from the same iFDT, and a checksum. This IFEC burst is then merged with all MPE sections of an original datagram burst, including its MPE-FEC if present, to generate the time slice burst that is actually sent over the air. This merging is done with the datagram just received when the delay parameter *D* is set to *D=0*, or with a previously received datagram when *D>0*. Note that the original data in MPE sections fully complies with EN 301 192 [9], clause 9.6.

This multi-burst parity computation and spreading is achieved at the expense of some latency for generating and receiving parity data. Nevertheless, as datagram bursts are sent unmodified, many well-known receiver operations are still possible. For example, the received MPE sections may be immediately forwarded to media decoders for fast channel-switching (zapping) support. Only when the parity is needed, the forwarding should be delayed to accumulate sufficient redundancy information.

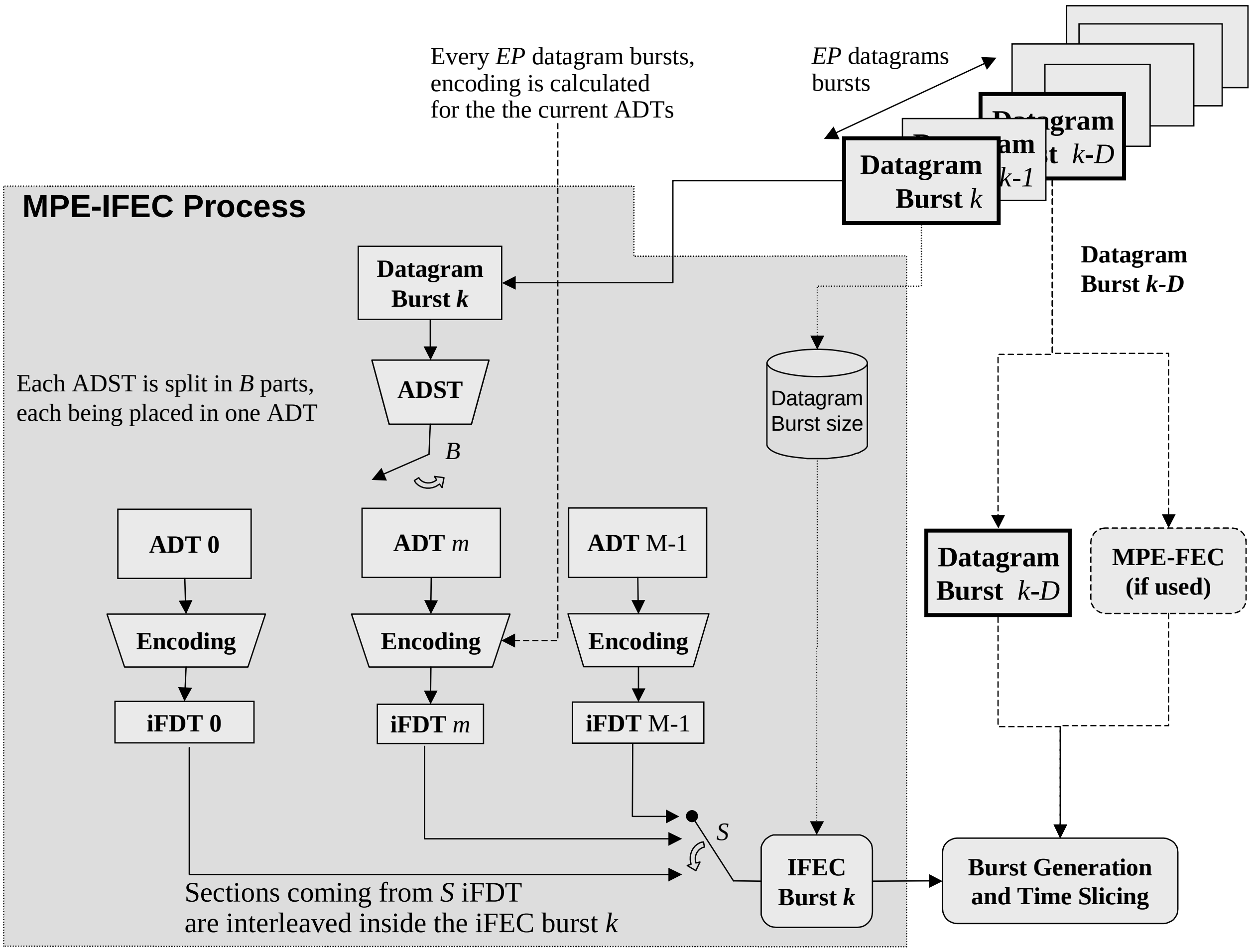


Figure 1: MPE-IFEC encoding process

The MPE-IFEC is introduced in such a way that MPE-IFEC ignorant (but MPE and MPE-FEC capable) DVB receivers will be able to extract the MPE stream in a fully backwards-compatible way. This backwards compatibility holds both when the MPE-IFEC is used with and without Time Slicing. The use of MPE-IFEC is not mandatory and is defined separately for each elementary stream in the transport stream. For each elementary stream it is possible to choose whether or not MPE-IFEC is used, and if it is used, to choose the trade-off between IFEC overhead, extra delay and performance. Time critical services, without MPE-IFEC and therefore minimal delay, could therefore be transmitted together with less time critical services using MPE-IFEC, on the same transport stream but on different elementary streams.

The MPE-IFEC is specified as a generic framework that presents enough flexibility for a variety of applications. For a usage in DVB-SH, its parameters are restricted to some specific values via the "framework mapping". Two of such "mappings" are presented in this document. One is based on MPE-FEC Reed Solomon code
(see EN 301 192 [9], clause 9.5.1) and is normative. The other mapping is based on Raptor code as specified in the Content Delivery Protocols (CDP) specification of IP Datacast over DVB-H [22] and is informative.

The document is intended to be ultimately added as an Annex to EN 301 192 [9], but can be used independently as it is natively self-contained. This document is structured as follows:

- chapter 1 gives this (informative) introduction;
- chapter 2 introduces (normative) definitions and abbreviations;
- chapter 3 introduces the (normative) sender operation;
- chapter 4 describes the (normative) carriage of MPE-IFEC Frame;
- chapter 5 provides (normative) syntax of Time Slice and FEC identifier descriptor and the two mapping examples, sliding encoding with RS code (normative) and generalized encoding with Raptor code (informative);
- chapter 6 introduces (informative) prototype IFEC decoding.

2 Notation, abbreviations and definitions (normative)

2.1 Notations

All symbols and parameters are in *italics*: e.g. *k*, *datagram_burst_size(k)*

A [B] means *A* modulo *B*

All functions are defined according to this template:

output_unit function_name(input_unit *parameter1*, input_unit *parameter2*,…) e.g. *ifdt_index(k)*

All fields that are transmitted are denoted with: `courier new`, e.g. `burst_number`.

2.2 Abreviations

IFEC: MPE-IFEC

NOTE 1: IFEC will be used inside the specification whenever possible.

NOTE 2: "i" is restricted to the short form "IFEC" whereas "I" is mandated for the long form MPE-IFEC.

ADT: Application Data Table
ADST: Application Data Sub Table
iFDT: IFEC Data Table
EM: Encoding Matrix

2.3 Definitions

The definitions are explicitly kept simple and not all parameters used by definitions are specified to avoid overloading explanations. However, all used parameters can be found, either in the generic sender operation description (section 3.2) or the specific mapping over the two parity computation codes (clauses 5.3 and 5.4).

2.3.1 Datagram burst

A datagram burst is a collection of one or several contiguous OSI layer 3 (Network layer) datagrams (e.g. IP datagrams). Datagram bursts are numbered with continuous sequence numbers *k*=0, 1, …

The bytes constituting one datagram burst are made of the succession of the bytes constituting the OSI layer 3 datagram sequence, starting with the first byte of the header of the first datagram of the sequence and ending with the last byte of the payload of the last datagram of the sequence. Each byte in the datagram burst is assigned a position address that indicates the number of bytes separating the start of the burst from this byte. Each datagram within each datagram burst is assigned an address pointing to the first byte of the datagram. Therefore, each datagram is uniquely identified by its datagram burst number *k* and its *address.*

Furthermore, each datagram burst *k* is assigned a datagram burst size referred to by *datagram_burst_size(k)* which is equal to the address of the last byte of the last datagram in the burst plus one byte.

Datagram burst size is limited to 2^{18}-1 = 262143 bytes due to signalling range restrictions of the address field in the real-time parameters in MPE header. However, the maximum size of a datagram burst MAY be restricted by other constraints. It is assumed that each datagram burst does not exceed a certain maximum datagram burst size *max_datagram_burst_size,* whereby an upper limit for *max_datagram_burst_size* is 2^{18}-1 due to the above-mentioned signalling reasons. Furthermore, the number of datagrams in each burst as well as the datagram burst size MAY vary for each datagram burst. The generation of datagram bursts from the sequence of OSI layer 3 (Network layer) datagrams is not further discussed, as it MAY depend on many different aspects.

2.3.2 Encoding Period

The Encoding Period, or *EP*, determines the frequency with which FEC is computed: an *EP* of 1 means that the encoding process occurs at every datagram burst whereas an *EP* greater than 1 means that the encoding process occurs every *EP* bursts. For *EP*>1, the encoding matrix capacity (ADT) is *EP* times greater than with *EP*=1 so that it takes *EP* times longer to fill it with data. *EP* is expressed in datagram burst units.

This parameter normalizes several other parameters:

- *EP* normalizes the encoding process depth *B*, which is the number of ADTs, expressed in units of EPs, over which the datagram bursts are interleaved;
- *EP* also normalizes the spreading process parameter *S*. The IFEC sections in one IFEC burst are interleaved from *S* iFDTs; here again S is expressed in *EP* units.

Finally, *EP* increases the encoding matrix size: since each ADT has *K*=*C***EP* columns, when *EP* is increased, the encoding matrix size is therefore increased. This allows IFEC to use very large encoding matrices where appropriate.

2.3.3 Application Data Sub-Table

The Application Data Sub-Table (`ADST`) is a function that allows mapping of a datagram burst to an ADT: thereby `ADST`(*j*) refers to the j^{th} column of the result of the mapping of the datagram burst on a matrix of *C* columns and *T* rows whereby *j*=0,1,…, *C*-1.

The mapping of a datagram burst with Layer 3 datagrams on such a *C***T* matrix is shown in figure 2: the leftmost columns of the matrix host all datagrams of a datagram burst. The remaining columns are filled with possible padding. The first datagram in the datagram burst starts with its first byte in the upper left corner of the matrix and goes downwards to the first column. The length of the datagrams MAY vary arbitrarily from datagram to datagram. Immediately after the end of one datagram the following datagram starts. If a datagram does not end precisely at the end of a column, it continues at the top of the following column. When all datagrams have entered the matrix, any unfilled byte positions are padded with zero bytes, making all columns completely filled. After the mapping, each position in the matrix hosts an information byte or padding byte.

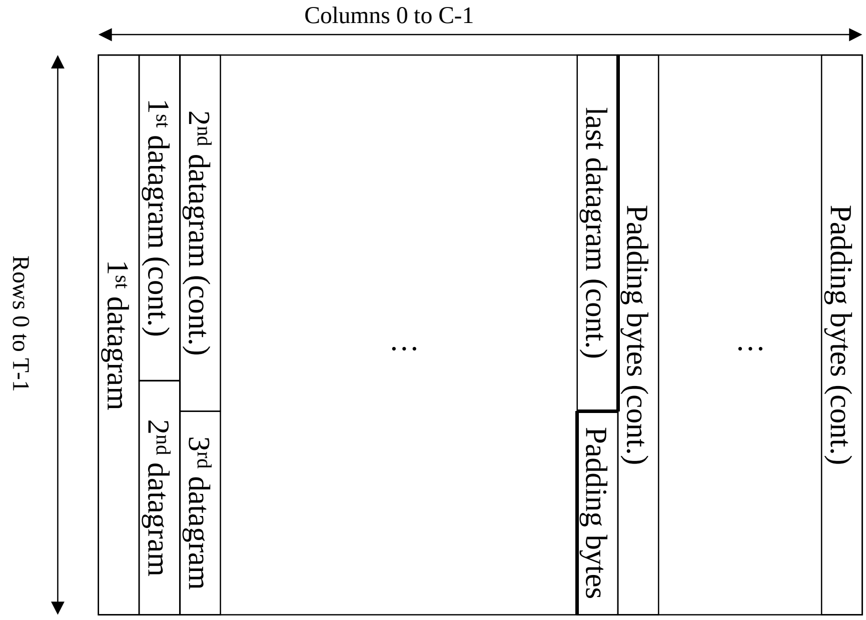


Figure 2: Datagram burst to ADST mapping

Note that maximum size of each datagram burst, *max_datagram_burst_size*, is restricted to $C*T$.

2.3.4 Encoding Matrix

The IFEC encoding process hosts M encoding matrices. Encoding matrices are sequentially numbered by $m= 0,\dots, M-1$. The actual number M depends on several parameters and is specified in section 5. Each encoding matrix m contains exactly one Application Data Table (ADT), referred to by `ADT`(m) and one IFEC Data Table (iFDT), referred to by `iFDT`(m).

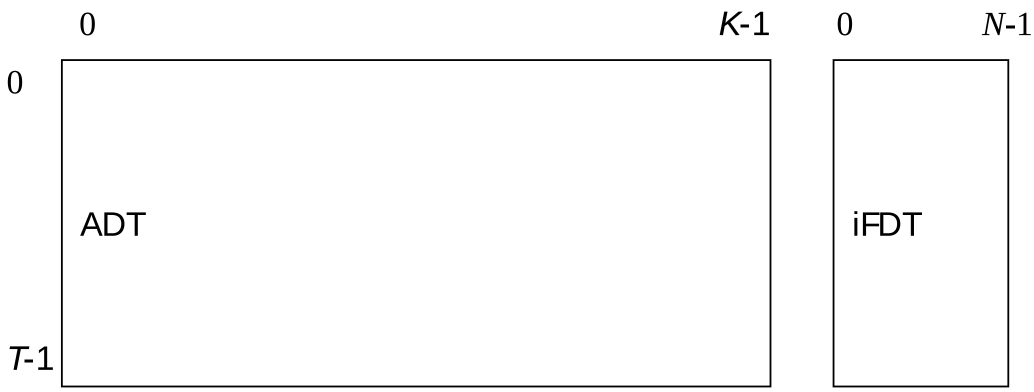


Figure 3: encoding matrix

2.3.4.1 Application Data Table

Each ADT (Application Data Table) is arranged as a matrix of $K=C*EP$ columns and T rows. Each column in each ADT is uniquely identified by its encoding matrix number m and its column number $n=0,\dots, K-1$ by `ADT`(m,n).

2.3.4.2 IFEC Data Table

Each iFDT (IFEC Data Table) is arranged as a matrix of N columns and T rows. Each column in each iFDT is uniquely identified by its encoding matrix number m and its column number $n=0,\dots, N-1$ by `iFDT`(m,n). The iFDT hosts the parity symbols generated for the corresponding ADT of the same encoding matrix.

2.3.5 MPE-IFEC section

An IFEC Section is comprised of a header, the data (parity symbols) from multiple iFDT columns in sequence, and a checksum. Structure of an MPE-IFEC section is specified in section 5 and generation of the IFEC section is covered in section 3.5.

2.3.6 IFEC burst

An IFEC burst is the collection of IFEC sections sent in one time-slice burst. The maximum size of an IFEC burst is R, where R is a number of IFEC sections. A new IFEC burst is generated with the reception of each datagram burst k. Each IFEC section of an IFEC burst is uniquely defined by its burst number $k' = k[k_{max}]$ and its IFEC section index $j=0,\ldots,R-1$. Each such IFEC section is therefore referred to as `IFEC`(k', j). Note that an IFEC burst MAY contain less than R IFEC sections and that the section indices MAY be not consecutive.

2.3.7 Time-slice burst

The time-slice burst is *what is actually sent over the air*. A time-slice burst corresponding to datagram burst k consists of a collection of:

- MPE sections generated from datagram burst k-D.
- MPE-FEC sections generated from datagram burst k-D (if MPE-FEC is also used).
- IFEC burst k containing MPE-IFEC sections generated when datagram burst k was received.

The transmission order of MPE sections, MPE-FEC sections, and MPE-IFEC sections is specified in section 3.6. Note that each of the sections requires settings of relevant real-time parameters, including time slicing. Time-slicing information can only be set after the definition of the transmission order.

3 Sender operation (normative)

3.1 Introduction

According to figure 1, the sender operation takes as its input the datagram bursts and generates at its output the time-slice bursts. The flow diagram of the sender operation is shown in figure 4. The initialization process before receiving the first datagram burst is described in section 3.3. Then following procedure is applied for each newly received datagram burst.

1) **Reception of new datagram burst**: the burst number k is incremented by one. The actual datagram burst k is stored such that it can be delivered after delay D. Its *datagram_burst_size(k)* is stored.

2) **Generation of IFEC burst:** The corresponding IFEC burst is generated taking into account the current datagram burst k, the data in iFDTs, as well as the *datagram_burst_size* of previous datagram bursts. The details of this process are described in section 3.5.

3) **Time-slice burst generation and sending:** The time-slice burst is generated from the MPE sections obtained from datagram burst k-D, possibly from the corresponding MPE-FEC sections included in this burst, and the IFEC sections of the corresponding IFEC burst. The details of this process are described in section 3.6.

4) **Datagram burst to ADTs mapping:** The actual datagram burst k is mapped by the use of the `ADST` to the ADTs. The ADST mapping includes the necessary padding. The details of this process are described in section 3.7.

5) **Generation of iFDT:** If the next burst number k+1 is a multiple of the encoding period EP it is necessary to generate the parity symbols in iFDT(m) from the data symbols in ADT(m), with m= floor(k/EP) [M]. The details of this process are described in section 3.8.

6) **Return to Receive the Next Datagram Burst:** The sender process continues in step 1 once a new datagram burst is received.

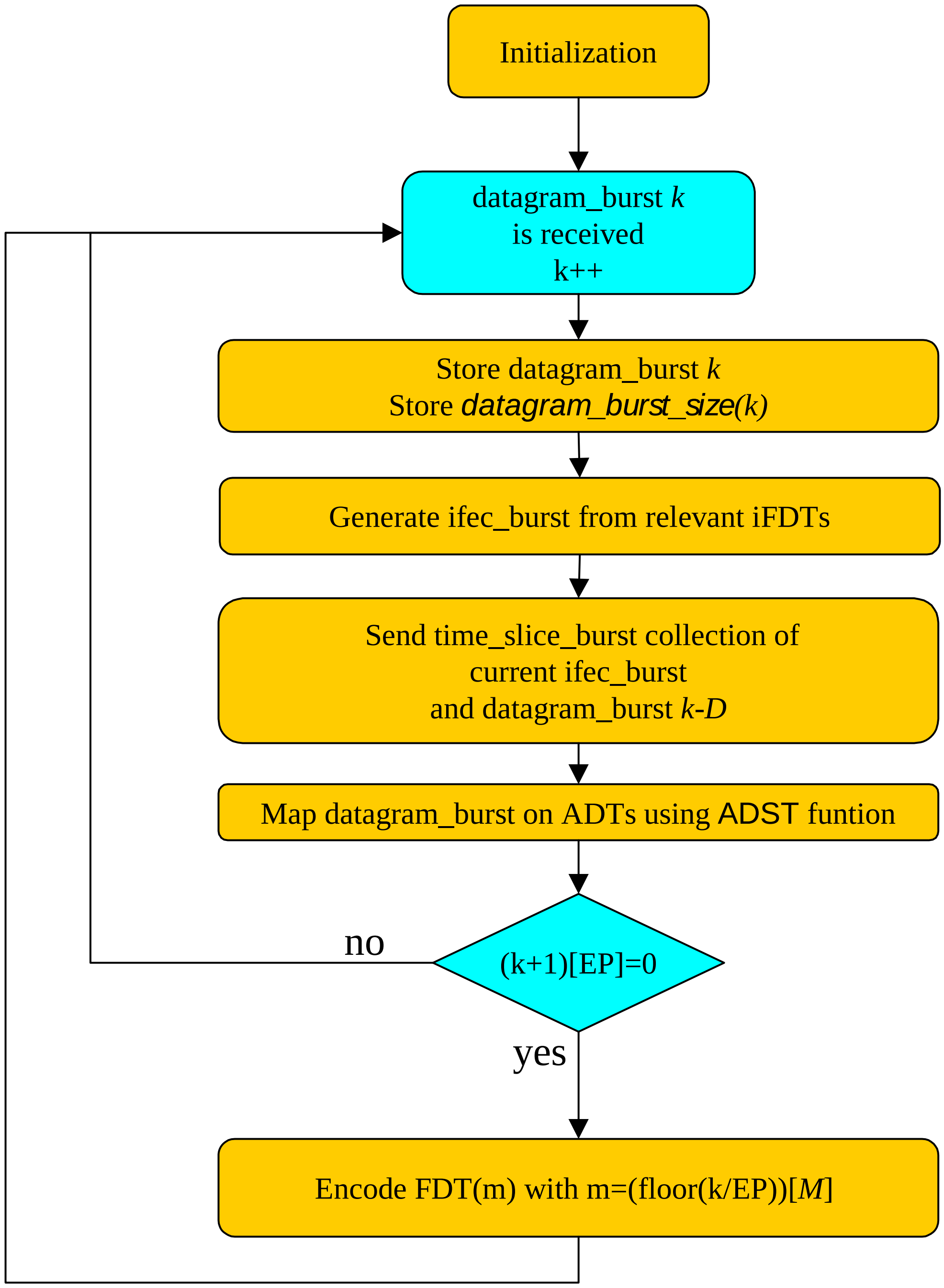


Figure 4: sender operation input/output

The description of the sender operation is the common framework. Functions and variables of this description are either part of the common framework and defined in this section or are specialized according to the code and defined in section 5.

3.2 Parameters

The following parameters are assumed to be available to the sender process and either by direct signalling or by indirect computation. The carriage or derivation of the parameters is detailed in chapter 5. Only those parameters required in chapter 3 are presented; as a consequence those parameters that are used only by the mapping functions `adt_index` and `ifdt_index` are actually given in chapter 5. This is in particular the case for B and S parameters that are defined in the chapter 5.

Table 1: MPE-IFEC generic parameters list

Parameter	Unit	Category	Description	Signalling	Scoping
EP	Datagram burst	Taxonomy	IFEC Encoding Period	Direct via Time_slice_fec_identifier	Time_slice_fec_identifier
D	Datagram burst	Taxonomy	Datagram burst sending delay	Direct via Time_slice_fec_identifier	Time_slice_fec_identifier
T	rows	Table sizing	Number of ADST, ADT, iFDT rows: *T*= MPE-FEC Frame rows /*G*	Indirect via Time_slice_fec_identifier	Time_slice_fec_identifier
C	columns	Table sizing	Number of ADST columns	Direct via Time_slice_fec_identifier	Time_slice_fec_identifier
R	sections	Table sizing	Maximum number of MPE IFEC sections per Time-Slice Burst	Direct via Time_slice_fec_identifier	Time_slice_fec_identifier
K	columns	Table sizing	Number of ADT columns = *EP***C*	Indirect via Time_slice_fec_identifier	Time_slice_fec_identifier
N	columns	Table sizing	Number of iFDT columns = *EP***R***G*	Indirect via Time_slice_fec_identifier	Time_slice_fec_identifier
G	columns	Table sizing	Maximum number of iFDT columns per IFEC section	Direct	Time_slice_fec_identifier
M	ADT	Protocol sizing	Number of concurrent encoding matrices *M*	Indirect (formula dependent on T_code and given in the parameter definition of chapter 5)	Time_slice_fec_identifier
K_{max}	N/A	Protocol sizing	Modulo operator for IFEC burst counter	Indirect (formula dependent on T_code and given in the parameter definition of chapter 5)	Time_slice_fec_identifier
j_{max}	N/A	Protocol sizing	Maximum backward pointing for datagram burst size used in PREV_BURST_SIZE parameter in §3.5	Indirect (formula dependent on T_code and given in the parameter definition of chapter 5)	Time_slice_fec_identifier
K	datagramburst	Index	continuous burst counter internal to sender	N/A	Loop
K'	IFEC burst	field	Burst number	N/A	IFEC section

3.3 Initialization

Before the first datagram burst is received, the following actions are required:

- the internal burst number k is set to any suitable value, e.g. k_{max}-1;
- M concurrent encoding matrices are allocated, each with one ADT of size K times T and one iFDT with size N times T;
- each ADT and iFDT is filled entirely with 0 bytes;
- the datagram burst size of virtual previous bursts k=0,…,k_{max}-1 is set to 0, i.e. for all k=0,…,k_{max}-1 *datagram_burst_size*(k)=0.

3.4 Reception of New Datagram Burst

With the reception of a new datagram burst, the following steps are carried out:

- assign datagram burst number k by incrementing counter k of one unit;
- store the corresponding datagram burst size *datagram_burst_size*(k);

- store the datagram burst k such that it can be delivered after delay D.

3.5 Generation of IFEC-Burst

An IFEC burst is generated taking into account the current burst number k, data in iFDTs, as well as the datagram burst size of previous datagram bursts. In total, at most R IFEC sections are generated for each IFEC burst. Each IFEC section in this burst gets assigned a unique IFEC burst number $k'=k$ [k_{max}]. In addition, each IFEC section gets assigned a unique section index j with j=0,1,…, R-1.

For each MPE-IFEC section `IFEC(k',j)` with IFEC burst number k' and section index j included in the IFEC burst, the following information is generated:

- burst_number = k';
- section_index = j;
- the iFDT index is obtained as m = ifdt_index(k', j); equivalently m is the encoding matrix number as defined in section 2.3.4;
- IFEC_data_bytes are obtained from iFDT(m) and correspond to the sequence of the iFDT columns iFDT(m, i), iFDT(m, $i+1$), …, iFDT(m, $i+g-1$) with i = ifdt_column(k', j) and g any integer between 1 and G. Each column iFDT(m) contributes databytes in order from row 1 to row T;
- max_iFDT_column = floor(255**max_iFDT_column*(m)/N), whereby *max_iFDT_column*(m) represents the largest column number plus one intended to be sent for the iFDT in encoding matrix m. The value *max_iFDT_column(m)* is lower or equal to N. A value of 255 SHALL be used for max_iFDT_column if *max_iFDT_column*(m) is not known and MAY be as large N;
- IFEC_burst_size defines the total number of IFEC sections included in this IFEC burst;
- section_length = $g*T$+13;
- prev_burst_size = *datagram_burst_size*((k-D-j[j_{max}]-1+k_{max})[k_{max}]);
- the remaining real-time parameters are set according to section 3.6.

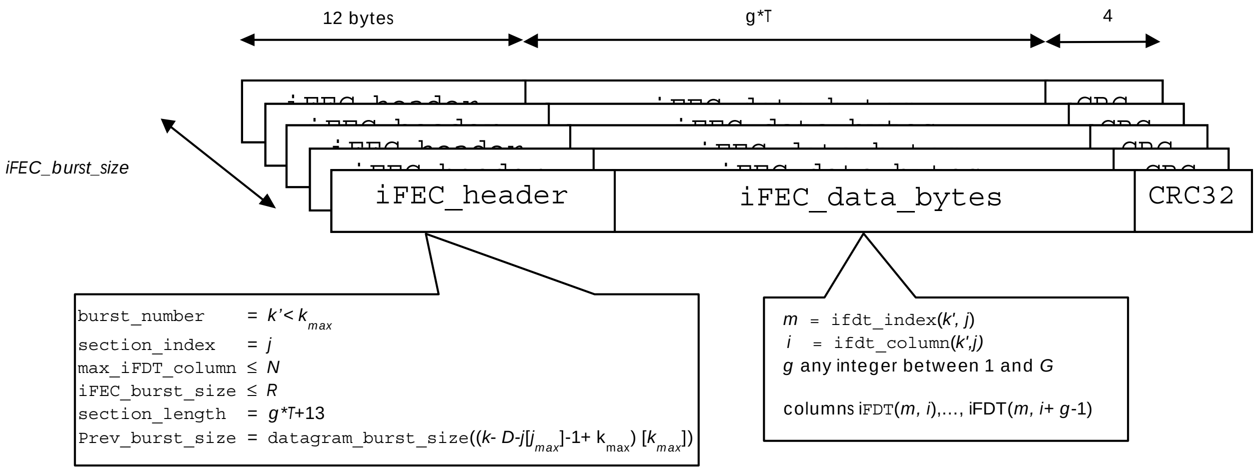


Figure 5: Generation of IFEC burst

The exact mapping of this header and payload information on IFEC sections is provided in section 4.2. The functions `ifdt_index`(k', j) and `ifdt_column`(k', j) are specified in section 5 and depend on `T_code` signalled in `id_selector_bytes` of `time_slicing_fec_descriptor` and some additional parameters in the `id_selector_bytes` and `time_slicing_fec_descriptor`.

3.6 Time-Slice Burst Generation and Sending

The time-slice burst is generated from the MPE sections obtained from datagram burst k-D, possibly from the corresponding MPE-FEC sections included in this burst k-D, and the IFEC sections of the corresponding IFEC burst generated when datagram burst k was received. The rules for generating this time-slice burst are listed below:

- a time-slice burst MUST contain all datagrams in the same order coming from datagram_burst(k-D):
  - datagram encapsulation is identical to the one specified in EN 301 192 [9], clause 9.6;
  - MPE sections within one time-slice burst SHALL be sent with increasing address;
- a time-slice burst MAY contain MPE-FEC sections related to this datagram burst k-D, if present. The generation and encapsulation of such MPE-FEC sections SHALL follow the procedure described in EN 301 192 [9], clause 9;
- a time-slice burst MAY contain IFEC sections of the corresponding IFEC burst. The encapsulation of the IFEC burst into IFEC sections is described in sections 3.5 and 4.2.

The following rules apply to MPE, MPE-FEC, and IFEC sections sending order within time-slice burst corresponding to the received datagram burst k:

- IFEC sections and MPE sections MAY be freely interleaved if their respective orders are respected;
- First IFEC section SHOULD be sent before any MPE section;
- Last IFEC section MAY be sent before last MPE section.

After having specified the sending order, for each of the MPE sections transmitted, the real-time parameters SHALL be set accordingly. For IFEC sections, the real-time parameters `delta_t`, `MPE_boundary`, and `frame_boundary` SHALL be set as specified in clause 4.3.

3.7 Datagram Burst to ADT Mapping

The datagram burst k is mapped by `ADST` function to the ADT, including the necessary padding as specified in section 2.3.

The mapping of the datagram burst to the ADST is specified in section 2.3. Then, for j= all 0,…,C-1 the ADST columns `ADST`(j) are shifted into `ADT`(m), whereby:

- k'=k[$k_{\max}$] and m = `adt_index`(k', j);
- shifting means that column `ADST`(j) is inserted at the right of `ADT`(m) moving all other columns to the left:
  - `ADT`(m,0) = `ADT`(m,1);
  - `ADT`(m,1) = `ADT`(m,2);
  - …;
  - `ADT`(m, K-2) = `ADT`(m,K-1);
  - `ADT`(m, K-1) = `ADST`(j).

m=`adt_index`(k', j) is specified in section 5. Note that the insertion of columns in ADTs via `ADST` function is independent of whether `ADST`(j) is a data or a padding column: they are treated identically.

3.8 Generation of iFDT

If the next burst number k+1 is a multiple of the encoding period EP it is necessary to generate the N parity symbols in `iFDT`(m) from the K data symbols in `ADT`(m), with m= floor(k/EP) [M].

The details of iFDT generation are presented in section 5 and depend on `T_code` signalled in `id_selector_bytes` of `time_slicing_fec_descriptor`.

4 Carriage of MPE-IFEC sections (normative)

4.1 Generalities

When `time_slice_fec_identifier_descriptor` specifies that MPE-IFEC is used on an elementary stream, the MPE-IFEC parity of each Time-Slice burst SHALL be delivered in MPE-IFEC sections described in table 2. MPE-IFEC sections are carried in the same elementary stream as the corresponding MPE data sections.

4.2 Syntax and semantics

MPE-IFEC sections are compliant to the DSMCC_section Type "User private" in ISO/IEC 13818-6 [23]. The mapping of the section into MPEG-2 Transport Stream packets is defined in ISO/IEC 13818-1 [8]. The syntax and semantics of the MPE-IFEC section are defined in table 2.

Table 2: MPE-IFEC section

Syntax	Number of bits	Identifier
`MPE-IFEC_section () {`		
`    table_id`	8	uimsbf
`    section_syntax_indicator`	1	bslbf
`    Private_indicator`	1	bslbf
`    Reserved`	2	bslbf
`    section_length`	12	uimsbf
`    Burst_number`	8	uimsbf
`    Reserved_for_future_use`	7	Bslbf
`    Current_next_indicator`	1	Bslbf
`    section_index`	8	uimsbf
`    max_iFDT_column`	8	
`    IFEC_burst_size`	8	
`    real_time_parameters()`	32	uimsbf
`    for( i=0; i<Nmax; i++ ) {`		
`        IFEC_data_byte`	8	uimsbf
`    }`		
`    CRC_32`	32	uimsbf
`}`		

The semantics for the MPE-IFEC section:

`table_id`: Shall be set to value of 0x79 (TBC with GBS).

`section_syntax_indicator`: This field SHALL be set to 1 and be interpreted as defined by ISO/IEC 13818-6 [23].

`private_indicator`: This field SHALL be set to 0 and be interpreted as defined by ISO/IEC 13818-6 [23].

`reserved`: Shall be set to "11".

`section_length`: Specifies the number of remaining bytes in the section immediately following this field up to the end of the section, including the CRC. Nmax is equal to section_length–9.

`burst_number`: This carries a burst continuity counter that SHALL vary between 0 and k_{max}-1, where k_{max} is the largest multiple of *EP* below 256. See section 3.5.

`reserved_for_future_use`: These seven bits SHALL be set, when not used, to "1111111".

`current_next_indicator`: Shall be set to a value of "1".

`section_index`: this 8-bit field gives the index of the section as defined in section 3.5.

`max_iFDT_column`: the largest iFDT column number of the iFDT addressed in this section to be sent as defined in section 3.5.

IFEC_burst_size: this 8-bit field gives total number of IFEC sections transmitted in the current time_slice_burst as defined in section 3.5.

Real time parameters: see section 4.3.

IFEC_data_byte: Contains the IFEC data delivered as specified in section 3.5.

CRC_32: This field SHALL be set as defined by ISO/IEC 13818-6 [23]. It is calculated over the entire MPE-IFEC_section.

4.3 Real time parameters

Each MPE-IFEC section SHALL carry real time parameters described in table 3.

Table 3: Time Slicing and MPE-IFEC real time parameters

Syntax	Number of bits	Identifier
real_time_parameters () {		
delta_t	12	uimsbf
MPE_boundary	1	bslbf
frame_boundary	1	bslbf
prev_burst_size	18	uimsbf
}		

delta_t: Usage of this 12-bit field depends on whether Time Slicing is used on the elementary stream.

The following applies when Time Slicing is used:

- The field indicates the time (delta-t) to the next Time-slice burst within the elementary stream. The time information is in each MPE-IFEC sections within a burst and the value MAY differ section by section. The resolution of the delta-t is 10 ms. Value 0x00 is reserved to indicate that no more bursts will be transmitted within the elementary stream (e.g. end of service). In such a case, all MPE-IFEC sections within the burst shall have the same value in this field as the MPE section.

- Delta-t information is the time from the start of the transport packet carrying the first byte of the current IFEC section to the start of the transport packet carrying the first byte of next burst. Therefore the delta-t information MAY differ between IFEC sections within a burst.

- The time indicated by delta-t shall be beyond the end of the maximum burst duration of the actual elementary stream. This ensures a decoder can always reliably distinguish two sequential bursts within an elementary stream.

The following applies when Time Slicing is not used:

- The field supports a cyclic ADST index within the elementary stream. The value of the field increases by one for each subsequent ADST. After value "111 111 111 111", the field restarts from "000 000 000 000".

- In case of large portions of lost data this parameter makes it possible to identify to which ADST the actual received section belongs.

MPE_boundary: This 1-bit flag, when set to "1", indicates that there are not any more MPE sections after this IFEC section in the time slice burst.

frame_boundary: This 1-bit flag, when set to "1", indicates that the current section is the last IFEC section of the time slice burst.

prev_burst_size: The field carries a description of the previous datagram burst size as specified in section 3.5.

5 Time Slice and FEC identifier descriptor (normative)

5.1 Introduction (informative)

This descriptor identifies whether MPE-IFEC is used on an elementary stream, in addition to the signalling of MPE-FEC. So the current specification comes in addition to the one found in EN 301 192 [9], clause 9.5 without contradicting it. In particular absolutely no change has been done on fields used by MPE-FEC.

The differences are:

- the use of the `id_selector_bytes` (text in `id_selector_byte` table);
- the `time_slice_fec_id` that can be set to 0x0 and 0x1 values (a new value MUST be inserted wherever a reference to `time_slice_fec_id` is made).

These differences need to be validated with GBS.

For clarity purpose, the text before the `id_selector_bytes` has been reproduced.

5.2 Descriptor (normative)

Table 4: Time Slice and FEC identifier descriptor

Syntax	Number of bits	Identifier
`time_slice_fec_identifier_descriptor () {`		
`descriptor_tag`	8	uimsbf
`descriptor_length`	8	uimsbf
`time_slicing`	1	bslbf
`mpe_fec`	2	uimsbf
`reserved_for_future_use`	2	bslbf
`frame_size`	3	uimsbf
`max_burst_duration`	8	uimsbf
`max_average_rate`	4	uimsbf
`time_slice_fec_id`	4	uimsbf
`for( i=0; I<id_selector_length; i++ ) {`		
`id_selector_byte`	8	bslbf
`}`		
`}`		

Semantics for Time Slice and FEC identifier descriptor:

`descriptor_tag`: Shall be set to value of 0x77.

`descriptor_length`: This 8-bit field specifies the number of bytes of the descriptor immediately following this field.

`time_slicing`: This 1-bit field indicates, whether the referenced elementary stream is Time Sliced. The value "1" indicates Time Slicing being used, and the value "0" indicates that Time Slicing is not used.

`mpe_fec`: This 2-bit field indicates whether the referenced elementary stream uses MPE-FEC, and which algorithm is used. Coding is according to table 5.

Table 5: Syntax and semantics for `mpe_fec`

value	MPE-FEC	algorithm
00	MPE-FEC not used	n/a
01	MPE-FEC used	Reed-Solomon(255, 191, 64)
10	Reserved for future use	
11	Reserved for future use	

`reserved_for_future_use`: This 2-bit field shall be set, when not used, to "11".

frame_size: This 3-bit field is used to give information that a decoder MAY use to adapt its buffering usage:

- in case Time Slicing is used (i.e. time_slicing is set to "1"), this field indicates the maximum number of bits on section payloads allowed within a Time-slice burst on the elementary stream, excluding MPE-IFEC sections. For MPE sections, bits are counted over ip_datagram_data_bytes or LLC_SNAP field (whichever is supported), excluding any possible stuffing_bytes. For MPE-FEC sections, bits are counted over rs_data_bytes;
- when MPE-FEC is used (i.e. mpe_fec is set to 0x1), this field indicates the exact number of rows on each MPE-FEC Frame on the elementary stream;
- if both Time Slicing and MPE-FEC are used on an elementary stream, both constraints (i.e. the maximum burst size and the number of rows) apply;
- when MPE-IFEC is used (time_slice_fec_id is set to 0x1) this field is computed as in the case both time slicing and MPE-FEC are used;
- coding of the frame_size is according to table 6.

Table 6: Syntax and semantics for frame_size

Size	Max Burst Size	MPE-FEC Frame rows
0x00	512 kbits = 524 288 bits	256
0x01	1 024 kbits	512
0x02	1 536 kbits	768
0x03	2 048 kbits	1 024
0x04 to 0x07	Reserved for future use	reserved for future use

max_burst_duration: This 8-bit field is used to indicate the maximum burst duration in the elementary stream when time slicing is used. A burst SHALL not start before T1 and SHALL end not later than at T2, where T1 is the time indicated by delta-t in the previous burst, and T2 is T1 + maximum burst duration. If the time_slice_fec_id is set to 0x0 or 0x1, the indicated value for maximum burst duration SHALL be from 20 ms to 5,12 s, the resolution is 20 ms, and the field is decoded according to the following formula:

Maximum burst duration = (max_burst_duration + 1) × 20 ms

If the time_slice_fec_id is set to any other value than 0x0 or 0x1, the coding of the max_burst_duration is currently not defined. When time_slicing is set to "0" (i.e. Time Slicing not used), this field is reserved for future use and SHALL be set to 0xFF when not used.

max_average_rate: This 4-bit field is used to define the maximum average bit rate in MPE section payload level over one time slicing cycle or MPE-FEC frame or ADST cycle and it is given by:

$$C_b = \frac{B_s}{T_c},$$

where B_s is the size of the current Time Slicing burst or MPE-FEC Frame or ADST cycle in MPE section payload bits and T_c is the time from the transport packet carrying the first byte of the first MPE section in the current burst/frame to the transport packet carrying the first byte of the first MPE section in the next burst/frame within the same elementary stream. Note that, when MPE-FEC is used, the RS data is not included in B_s nor is the IFEC parity data when MPE-IFEC is used. If time_slice_fec_id is set to 0x0 or 0x1, the coding of the max_average_rate is according to table 7. If time_slice_fec_id is set to any other value, coding of the max_average_rate is currently not defined.

Table 7: Syntax and semantics for `max_average_rate`

Code	Bit rate
0000	16 kbps
0001	32 kbps
0010	64 kbps
0011	128 kbps
0100	256 kbps
0101	512 kbps
0110	1 024 kbps
0111	2 048 kbps
1000 to 1111	reserved for future use

`time_slice_fec_id`: This 4-bit field identifies the usage of following `id_selector_byte`(s). Note that this field affects on coding of `frame_size`, `max_burst_duration` and `max_average_rate` fields on the actual descriptor, and the address field of real-time parameters on the referred elementary stream. If this field is set to value 0x0 `id_selector_byte`(s) SHALL not be present. If this field is set to value 0x1 then the id bytes are the following. Other values are not defined.

`id_selector_length`:

This 8-bits field gives the lengths of the `id_selector_byte` field counted in bytes:

if `time_slice_fec_id` is set to 0, this value is set to 0 (no `id_selector_bytes`).

if `time_slice_fec_id` is set to 1, this value is set to 8 (`id_selector_bytes` is set to 8 bytes).

`id_selector_byte`:

if `time_slice_fec_id` is set to 0x1 the bytes have the following definition:

Table 8: semantics for time_slice_fec_id

Syntax	Number of bits	Identifier
`Time_slice_fec_id_0x1() {`		
`T_code`	2	Uimsbf
`G_code`	3	uimsbf
`Reserved for future use`	3	bslbf
`R`	8	uimsbf
`C`	13	uimsbf
`Reserved for future use`	3	bslbf
`B`	8	uimsbf
`S`	8	uimsbf
`D`	8	uimsbf
`EP`	8	uimsbf
`}`		

`T_code`: This 2-bit field indicates the type of IFEC code used. Currently only value 0 is used and corresponds to MPE-FEC Reed Solomon (255,191).

Section 5.4 adds an informative section introducing the use of Raptor codes as specified in TS 102 472 [22], clause C.4 in the framework.

Table 9: Syntax and semantics for `T_code`

Bit rate	Description
00	Reed Solomon code (EN 301 192 [9], clause 9.5.1)
01	Raptor Codes ([22], clause C.4)
01 to 11	reserved for future use

`G_code`: This 3-bit field indicates value of the *G* parameter as *G*=2^`G_code` (number of symbols).

Table 10: Syntax and semantics for G_code parameter

Bit rate	*G*
000	1
001	2
010	4
011	8
100	16
101	32
110	64
111	128

reserved_for_future_use: This 3-bit field SHALL be set, when not used, to "111".

EP: This 8-bit field indicates value of the encoding period parameter.

R: This 8-bit field indicates the maximum number of MPE-IFEC sections transmitted in an IFEC burst.

C: This 13-bit field indicates the number of columns for the ADST.

reserved_for_future_use: This 3-bit field SHALL be set, when not used, to "111".

B: This 8-bit field indicates value of the encoding process depth parameter normalized by *EP*. *B* is the number of ADTs over which the datagram bursts are interleaved.

S: This 8-bit field indicates value of the spreading process parameter normalized by *EP*: the IFEC sections in one IFEC burst are interleaved from S FDT.

D: This 8-bit field indicates value of the delaying process parameter. *D* is the number of datagram bursts by which the current transmission lags the current datagram burst.

5.3 Sliding Encoding with RS code (normative)

5.3.1 General

This encoding scheme is an extension of the MPE-FEC. It uses the same encoding parity code as the MPE-FEC, benefiting from its existing hardware support.

5.3.2 Parameter Definitions

The following parameter restrictions apply:

- the T_code value SHALL be set to 00;
- the encoding period MUST be set to 1, i.e. *EP* is set to 1;
- the IFEC data spread, *B*, is signalled in the id_selector_byte->B, [0;255];
- the IFEC spread, *S*, is signalled in the id_selector_byte->S, [0;255];
- the data burst delay, *D*, is signalled in the id_selector_byte->D, [0;255];
- the number of data columns in the ADST, *C*, is signalled in the id_selector_byte->C, [0;191];
- the maximum number of sections in one IFEC burst, *R*, is signalled in id_selector_byte->R, [0;64];
- the number of rows in the ADST, ADT, and iFDT *T* is signalled in frame_size->mpe_fec_frame_rows $\in$ {256; 512; 768; 1024};
- the maximum number of iFDT columns in one MPE-IFEC section, *G*, MUST be set to 1, i.e. G_code MUST be set to 000;
- the number of encoding matrices, *M*, is given as $M=B+\max(0, S-D)+\max(0, D-B)$;

- furthermore, with *EP*=1, it is obvious that the number of columns in each ADT is given as *K*=*C* and, in addition with *G*=1 the number of columns in each iFDT is given as *N*=*R*;
- the modulo burst counter is set to k_{max}=256-256[M];
- the modulo counter for the previous burst signalling is set to j_{max}=*M*.

For the special case *D*=0, this encoding scheme results in the case that each Datagram Burst is interleaved in a sliding window of B ADTs out of the total of *M*=*B*+*S* ADTs. For each datagram burst only one iFDT is computed, that will be sent in the IFEC Bursts of the S following time-slice bursts.

5.3.3 Mapping Functions

The following parameter mapping functions shall be used:

the ADT index for a given modulo datagram burst number *k*' and a given ADST column number *j* is given as:

$$\texttt{adt\_index}(k', j) = (k' + j[B])[M];$$

the iFDT index for a given modulo datagram burst number *k*' and a given section index number *j* is given as:

$$\texttt{ifdt\_index}(k', j) = (k' - j[S] - 1 + \mathrm{M})[M];$$

the iFDT column for a given modulo datagram burst number *k*' and a given section index number *j* is given as:

$$\texttt{ifdt\_column}(k', j) = j.$$

5.3.4 Generation of iFDT

For a given ADT, the generation of the iFDT follows the definition of EN 301 192 [9], clause 9.5.1.

Any *K*<191 is achieved by shortening as specified in clause 9.3.3.1.

Any *N*<64 is achieved by puncturing as specified in clause 9.3.3.2.

5.4 Generalized Encoding with Raptor code (informative)

5.4.1 General

The inclusion of a Raptor encoding scheme extends the flexibility compared to the sliding RS scheme as presented in section 5.3. This is achieved as the ADT size can be increased significantly and one MAY benefit from the possibility of a full software implementation of the IFEC decoding.

In case that this encoding scheme is used, this is signalled by the FEC identifier descriptor by T_code value 01.

5.4.2 Parameter Definitions

The following parameter restrictions apply:

- the `T_code` value shall be set to 01;
- the encoding period *EP* is signalled in the `id_selector_byte->EP`, [0;255];
- the IFEC data spread, *B*, is signalled in the `id_selector_byte->B`, [0;255];
- the IFEC spread, *S*, is signalled in the `id_selector_byte->S`, [0;255];
- the data burst delay, *D*, is signalled in the `id_selector_byte->D`, [0;255];
- the number of data columns in the ADST, *C*, is signalled in the `id_selector_byte->C`, [4;8192];
- the maximum number of sections in one IFEC burst, *R*, is signalled in `id_selector_byte->R`, [0;256];

- the maximum number of iFDT columns in one MPE-IFEC section, G, is signalled in the `id_selector_byte->G_code`, whereby G=2^(G_code);
- the number of rows in the ADST, ADT, and iFDT T is derived from the `frame_size->mpe_fec_frame_rows` $\in$ {256; 512; 768; 1024} and the parameter G as T=(`frame_size->mpe_fec_frame_rows`)/G;
- the number of columns in an each ADT is given as $K=C*EP$, whereby it MUST be ensured that $K \leq 8192$;
- the number of columns in each iFDT is given as $N=R*G*EP$;
- the number of encoding matrices, M, is defined as:

$$M = \max(B, S, \mathrm{ceil}(D/EP), B+S-\mathrm{floor}(D/EP));$$

- the modulo burst counter is set to $k_{max} = \mathrm{floor}(256/(M*EP))*M*EP$;
- the modulo counter for the previous burst signalling is set to $j_{max}=M*EP$.

5.4.3 Mapping Functions

The following parameter mapping functions shall be used:

the ADT index for a given modulo datagram burst number k' and a given ADST column number j is given as:

$$\texttt{adt\_index}(k', j) = (\ \mathrm{floor}(k'/EP) + j[B])[M];$$

the ADT column for a given modulo datagram burst number k' and a given ADST column number j is given as:

$$\texttt{adt\_column}(k', j) = \mathrm{floor}((\mathrm{floor}(s/EP)*C)/B)+C*(s[EP]) + \mathrm{floor}(j/B),$$

$$\text{with } s = (B-j+(k'[EP])[B]-1)*EP + k'[EP];$$

the iFDT index for a given modulo datagram burst number k' and a given section index number j is given as:

$$\texttt{ifdt\_index}(k', j) = (\ \mathrm{floor}(k'/EP) - j[S] - 1 + M)[M];$$

the iFDT index for a given modulo datagram burst number k' and a given section index number j is given as:

$$\texttt{ifdt\_column}(k', j) = (k'[EP])*R + j*G.$$

5.4.4 Generation of iFDT

The iFDT is generated using the Systematic Raptor Encoder as specified in TS 102 472 [22], clause C.4, whereby:

- the ADT corresponds to a source block;
- columns of the ADT correspond to source symbols;
- columns of the iFDT correspond to repair symbols.

The source block size K is defined as $C*EP$, the symbol size T corresponds to the column size of the ADT and iFDT. Sub-blocking shall not be used.

5.4.5 Example Parameters

This section provides recommendations for the setting of the transport parameters `id_selector_byte->G_code` and `id_selector_byte->C` at the sender. Note that other parameters MAY be applied and signalled based on specific criteria.

This recommendation is based on the following input parameters:

- N_{Burst} the maximum burst size as signalled by the `frame_size` value;
- nof_fec_rows, the MPE-FEC frame rows as signalled by the `frame_size`;

- r the target code rate between 0 and 1;
- EP the encoding frequency;
- K_{MAX} the maximum number of columns per ADT with $K_{MAX} = 8192$ according to the maximum systematic index in TS 102 472 [22], clause C.5;
- K_{MIN} the minimum number of columns per ADT with $K_{MIN} = 4$ according to the minimum systematic index in TS 102 472 [22], clause C.5;
- K_{TARGET} the target number of columns per ADT;
- T_{MIN} the minimum symbol size for the ADT.

Then, $N_{ADST} = \text{ceil}(r * N_{Burst})$ is the expected maximum ADST size.

Then, $T = \max\{T_{MIN}, \min_{i=0,1,2\ldots,7}\{ G=R/2^i \mid (EP * \text{ceil}(N_{ADST}/(R/2^i)) <= K_{TARGET}) \} \}$ is an appropriate symbol size, i.e. T is selected such that the Raptor source block has at least K_{TARGET} symbols, but the symbol size is always at least T_{MIN} bytes and divides the number of MPE-FEC frame rows by a power of 2.

Then the proposed parameters are as follows:

G = nof_fec_rows / T and `id_selector_byte->G_code` = $\log_2(G)$

$C = \text{ceil}(N_{ADST}/T)$ and `id_selector_byte->C` = C

Recommended settings for the input parameters are $K_{TARGET} = 2048$ and $T_{MIN}=32$.

Assume a frame_size=0x03 with N_{Burst}=2048 kbits and nof_fec_rows=1024, then the above algorithm results in the following parameter settings.

Table 11: Recommended parameter settings for frame_size=0x03

	***EP*=1**		***EP*=4**		***EP*=8**		***EP*=16**		***EP*=32**	
	C	G	C	G	C	G	C	G	C	G
R=7/8	3584	16	896	4	224	1	448	2	224	1
R=3/4	3072	16	768	4	192	1	384	2	192	1
R=2/3	2730	16	682	4	170	1	341	2	170	1
R=1/2	2048	16	512	4	128	1	256	2	128	1

6 Receiver Operation (informative)

6.1 Introduction

In this clause we give informative background on how an MPE-IFEC decoder compliant to the coding profiles defined in sections 5.3 and 5.4 could be implemented:

- we give a general process overview;
- then we give more details on particular steps of this process:
    - parameters derivation from the time_slice_fec_descriptor;
    - burst number detection;
    - section reception;
    - padding ADST;
    - main options in the decoding process.

6.2 General Process

The sender operation takes as its input the sections received in one time-slice burst and generates at its output datagram bursts that contains a sequence of datagram. The basic steps of an example receiver operation are shown in figure 6.

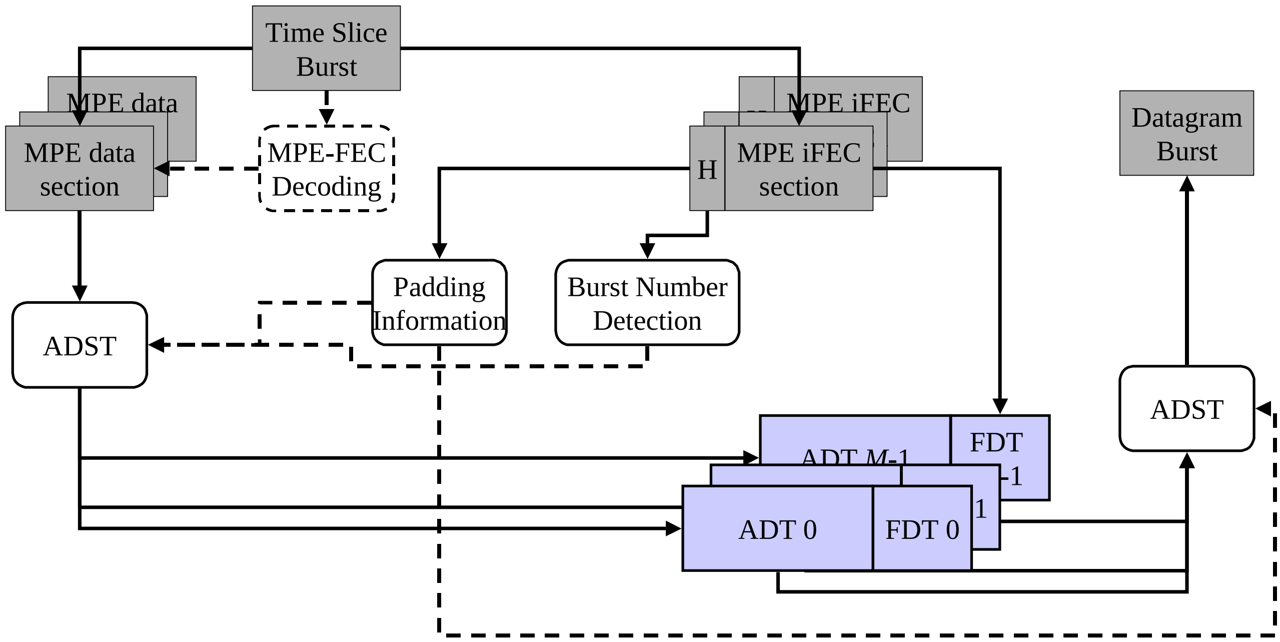


Figure 6: Example Receiver Operation

Assume that we have detected the reception of a time-slice burst that MAY contain:

- none, one or several MPE data sections;
- none, one or several MPE-FEC sections;
- none, one, or several MPE-IFEC sections.

In case that MPE data sections are detected to be missing, and if MPE-FEC sections for this time-slice burst have been received, an MPE-FEC decoding MAY be initiated to recover the MPE data sections for this time-slice burst. In any case, even if everything is correct, the following IFEC decoding process should be started because the data can be used to correct other burst.

The following steps are given for information; other alternatives MAY be given to optimize display time (e.g. fast zapping):

0. detect current burst number; different techniques are presented in section 6.4;
1. use received IFEC sections to perform IFEC decoding; this is presented in section 6.7. If a combination of ADT and iFDT has sufficient data to recover, an ADT MAY be recovered and corresponding datagram bursts MAY be recovered;
2. use received MPE section and, using ADST function, map them onto the ADTs; this is detailed in sections 6.5 and 6.6;
3. go to 0.

When one or several successive complete burst losses happen, it MAY be impossible to detect burst numbering straight away. When this is possible (for instance when another burst is received and the loss is detected) then step 2 is applied for each previously lost burst.

Usually one IFEC section is enough to reconstruct the burst number, reconstruct some padding information, and to insert the IFEC section parity bytes in the relevant positions in the iFDTs (see section reception).

6.3 Parameters

With the availability of an INT and `time_slicing_fec_descriptor`, the following parameters can be determined following the same procedure as at the sender.

Table 12: parameters for receiver operations

Static Parameters	Description
EP	IFEC Encoding period
B	IFEC Data Interleaving
S	IFEC Spread
D	Data delay at sender
C	Maximum number of data columns per ADST
T	Symbol size of code (Number of rows in ADST/ADT/iFDT)
G	Maximum number of symbols per MPE-IFEC section
M	Number of Concurrent ADT and iFDT
K	Number of columns in ADT
N	Number of columns in iFDT
R	maximum number of IFEC sections in an IFEC burst
k_{max}	Modulo counter for burst
T_{bmax}	maximum burst duration

6.4 Burst number detection

Since MPE sections do not carry the time-slice burst number, the first thing that a DVB-IFEC capable receiver must do is to determine to which time-slice burst the section belongs to. This can be done through different ways:

- an MPE-IFEC section must have already been received beforehand so that we have a reference number of burst in the past;
- we can derive the current burst number using real time parameters of the MPE section:

  let $I_p/T_p/\Delta T_p$ be the burst number and time of the preceding MPE-IFEC section;

  let $I_n/T_n/\Delta T_n$ be the burst number and time of the following MPE-IFEC section;

  let T and ΔT the time and delta-t information of the current MPE section.

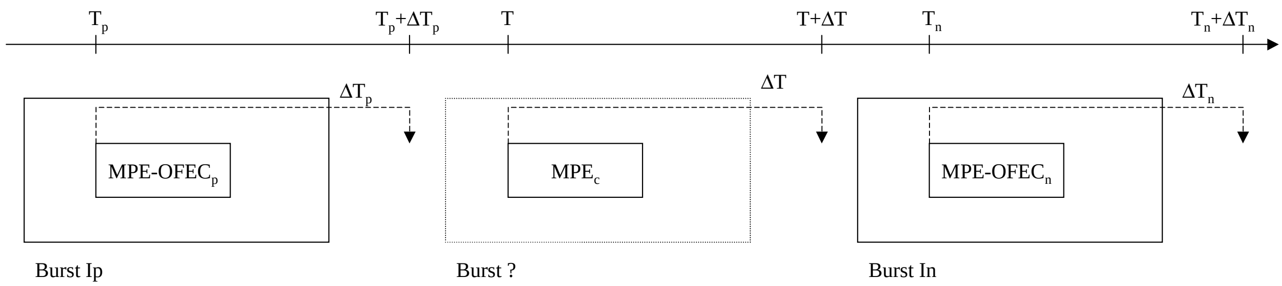


Figure 7: burst numbering, general case

We are looking for the current burst number:

- if (I_p==I_n) burst_number=$I=I_p=I_n$;
- otherwise, the result MAY be based on some timing heuristic:
  - by definition, $T_p<T$ and $T<T_n$;
  - if $T_p+\Delta T_p<T$ and $T_n<T+\Delta T$ THEN $I=I_n$.

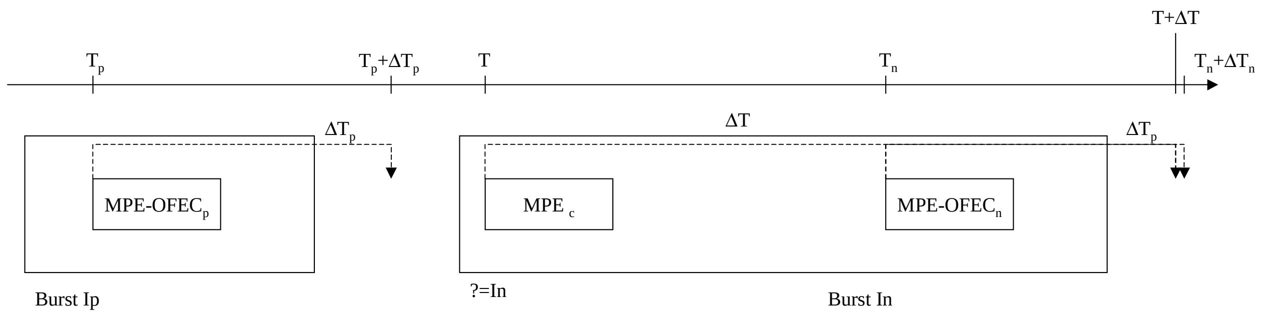


Figure 8: burst numbering, case of next burst

- ELSE $T<T_p+\Delta T_p$ THEN $I=I_p$

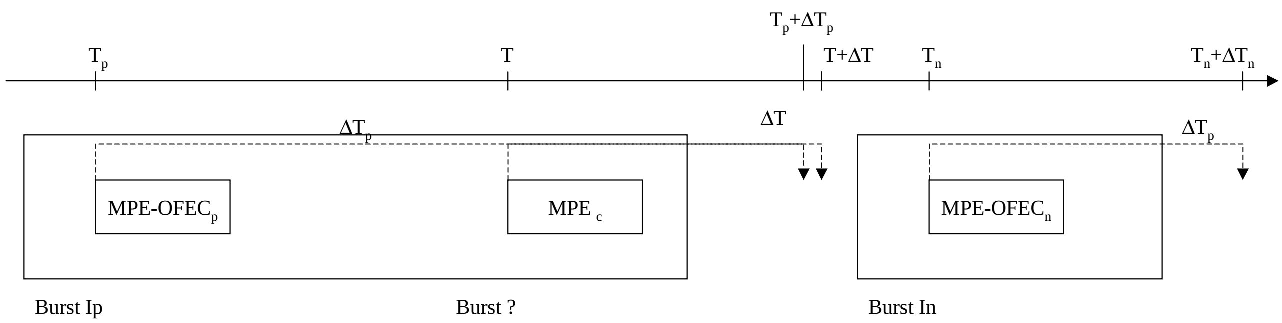


Figure 9: burst numbering, case of previous burst

- ELSE IGNORE that MPE section for MPE-IFEC decoding

- Obviously, the discovery of the $I=I_p$ can be achieved only if there are MPE-IFEC sections preceding the MPE data sections, the latter being completely implementation dependent.

6.5 Section Reception

Once the burst numbering has been done, it is possible to place each received section inside their ADT or iFDT:

- MPE-IFEC sections:
  - each MPE-IFEC section carries a `section_index` that enables to position this IFEC section inside one iFDT according to the algorithm described in section 3.5;
  - note that in any case (and contrary to MPE-FEC), even if all sections in current datagram burst have been received correctly, the receiver shall continue to receive remaining MPE-IFEC sections since they can be used in different FDTs than the ones used to decode the current ADST, or they could be used in the same FDTs as those used to decode the current ADST but for decoding other more erroneously received ADSTs;
  - the number of received sections in the current burst cannot exceed the signalled `fec_burst_size` in every IFEC section received. If there are fewer sections delivered, this means that some IFEC sections have been lost;
  - when reception is over, in case one section is missing, there MAY be a decision from the receiver with regard to decoding because it has learnt the total number of IFEC columns transmitted for an iFDT using `max_iFDT_column` signalled in the sections carrying the columns of this iFDT. If losses are too severe on the ADT, it MAY be impossible to decode;
  - the IFEC sections carry also `prev_burst_size` information enabling to delineate the padding part within the previously received burst.

- MPE sections:
  - each correctly received (id est with a correct CRC32) MPE-section can be placed at the right position inside the ADST since each MPE section carries in the section header a start address for the payload. The receiver will then be able to put the received datagram in the right byte positions in the ADST and mark these positions as "reliable";
  - when all possible MPE sections correctly received have been processed, the padding or erased bytes are added according to chapter 6.6. Erased bytes are positioned as unreliable whereas padding bytes are positioned as reliable. Then the *C* columns are interleaved inside their relevant ADT using same algorithm as in the sender described in section 3.6.

6.6 Padding in ADST mapping

The receiver introduces the number of padding columns during the ADST mapping by marking these padding bytes as reliable. Some of the following guidelines MAY be used to recover the padding information whereby the description is supported by some examples in figure 10:

- if the receiver has received the last MPE section of the time-slice burst correctly, it can mark any remaining padding bytes as reliable (case k'-2 in the figure below);
- if the receiver did not receive the last section correctly, it has to check if the `prev_burst_size` or this burst is available; if yes, this implies that all byte positions after the last correctly received section until the burst size signal by `prev_burst_size` are lost data and will mark them as unreliable (case k'-1);
- if the receiver has received neither the signalling of the padding size using `prev_burst_size,` nor the last MPE section, then it will have to mark all bytes positions after the last correctly received section as lost data and mark the corresponding byte positions as unreliable (case k'-3);
- if there are missing sections before the last section, the receiver can set as unreliable the corresponding bytes (case k'-4).

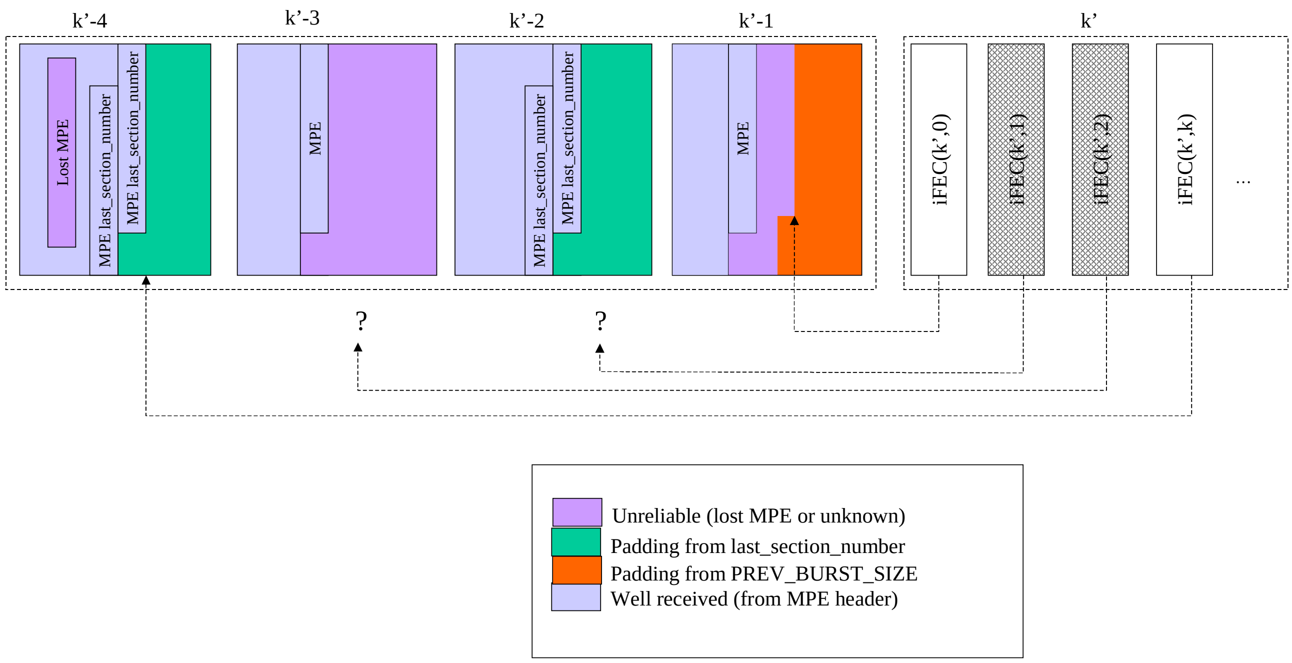


Figure 10: padding strategy

6.7 Decoding

The scope of this section is to exemplify typical receiver procedures for the reconstruction of a flow of IP datagram bursts from a sequence of received time-slice bursts. The reference receiver presented in this section shall be intended as to meet minimum performance requirement and therefore its structure is intentionally kept simple. Specific implementations MAY differ in some implementation aspects, but they shall still be considered fully compliant as long as they fulfil at least the minimum performance of the reference receiver.

6.7.1 Input

The IFEC decoding attempts to recover an ADT with use of the information in the ADT, the padding information in the ADT, as well as the corresponding iFDT. The ADT consists of K columns and T rows, the iFDT consists of $R*EP$ columns and T rows. Only CRC32 valid sections are positioned: all the bytes that constitute the section are signalled as unreliable; in order to save memory, each column in the ADT is MAY be marked as unreliable, if any of the T bytes in the column is erased. Then, if the total number of non-erased columns in ADT and iFDT is at least K, IFEC decoding shall be applied.

6.7.2 RS decoding

For the case of RS codes, at every burst, one (ADT, iFDT) couple can be decoded using the same principles as for MPE-FEC presented in EN 301 192 [9], clause 9.3.3.

6.7.3 Raptor decoding

For the case of Raptor codes, an appropriate decoding algorithm is provided in TS 102 472 [22], clause C.7, whereby

- the source block is determined by the number of symbols K and the symbol size T;
- the ensemble of encoding symbols consists of;
  - the non-erased columns in the ADT, whereby the encoding symbol ID (ESI) corresponds column number in the ADT;
  - the available columns in the iFDT, whereby the ESI corresponds to the column number in iFDT plus K.

If IFEC decoding is successful, the ADT columns are updated by the reconstructed encoding symbols.

6.7.4 Output

Depending on the number of erroneous bytes after frame decoding, several options are possible:

- all missing bytes have been corrected, all datagrams are forwarded to the application;
- some bytes have not been corrected; only correct datagrams are forwarded to the application;
- some bytes have not been corrected; the erroneous datagrams are also forwarded to the application.

Annex B (informative): Interoperability with cellular telephony networks

B.1 Introduction

Mobile terminals can contain several radios and therefore co-existence of the DVB-SH receiver working in L-band (1 452 MHz to 1492 MHz) and S-band (2 170 MHz to 2 200 MHz) with other radios, either cellular or connectivity, is required. The cellular radios can be GSM/EDGE (including GSM850, GSM900, DCS1800 and PCS1900) or UMTS while the connectivity radios can be wireless LAN or Bluetooth. Simultaneous operation of the DVB-SH receiver in combination with one of the cellular or connectivity radios in a small sized terminal is very challenging.

B.1.1 General coexistence issues

The co-existence issues described in this clause consider worst case conditions:

two main co-existence issues for the DVB-SH receiver can be distinguished:

1) interference of the uplink signal of a simultaneous (cellular or connectivity) transmission with the DVB-SH receiver:
   - wanted transmitted signal: desensitization;
   - unwanted transmitted signal: power amplifier noise and spurious responses;
2) interference of the downlink signal of a simultaneous (cellular or connectivity) reception with the DVB-SH receiver:
   - wanted transmitted signal from a base station: adjacent channel selectivity (ACS);
   - unwanted transmitted signal from a base station: downlink Adjacent Channel Leakage Ratio (ACLR).

Next to this, undisturbed operation of the cellular or connectivity radio in presence of the DVB-SH receiver must also be maintained. Because DVB-SH does not have a uplink signal, the possible impairments caused by DVB-SH receiver are limited to:

1) out of band unwanted signals in cellular or connectivity downlink (RX) band;
2) effects to the cellular antenna pattern.

These problems are pure implementation issues and can be solved by proper terminal design.

B.1.2 Terminal Architectures

The architecture (relevant parts) of a typical terminal category 3 containing a DVB-SH receiver and a GSM, UMTS or WLAN radio is presented in figure B.1.

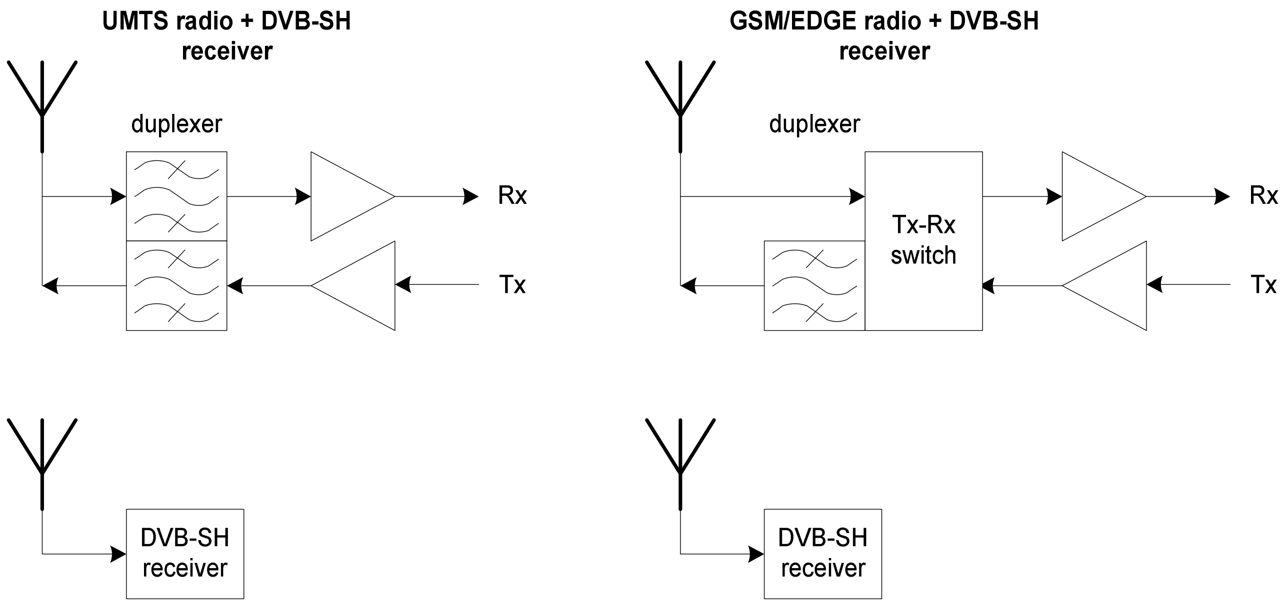


Figure B.1: Terminal architectures with cellular radios

Most probably the DVB-SH receiver and the cellular radio will have physically separate antennas. The antenna isolation between the antennas is frequency dependent and will be detailed in the next clause.

An important difference between UMTS and GSM radios is the duplex filter. UMTS will use duplex filters, but the majority of modern GSM radios uses a Tx/Rx switch. This has a major implication on the co-existence. The cellular radio uplink unwanted signal interference to the DVB-SH receiver will not be a problem in UMTS terminal if a duplexer is used. However the problem will be severe in a GSM terminal with Tx/Rx switch.

It is worth noting that in case of terminal category 2 the cellular radio transmitters are not considered. However, the WLAN radio transmitter might be present causing also signal interferences.

B.2 Frequency bands and power levels

Table B.1 gives an overview of the most relevant interferers for DVB-SH including frequency bands and power. For uplink signals, the transmit power and the assumed antenna coupling values are given for S-band receiver. Together they give the received power at the DVB-SH antenna.

Table B.1: Overview of interferers for DVB-SH receiver

Name	Frequency band	Transmit power	Antenna coupling	Power at DVB-SH antenna
GSM850 Tx	824 MHz to 849 MHz	33 dBm	-18 dB	15 dBm
GSM900 Tx	880 MHz to 915 MHz	33 dBm	-18 dB	15 dBm
DCS1800 TX	1 710 MHz to 1 785 MHz	30 dBm	-15 dB	15 dBm
PCS1900 TX	1 850 MHz to 1 910 MHz	30 dBm	-15 dB	15 dBm
FDD 3G band 1 Tx	1 920 MHz to 1 980 MHz	24 dBm	-9 dB	15 dBm
FDD 3G band 1 Rx	2 110 MHz to 2 170 MHz	n.a.	n.a.	
ISM2400 (BT-WLAN)	2 400 MHz to 2 484 MHz	20 dBm	-9 dB	11 dBm

The most problematic cases are the wanted uplink and downlink signals from the FDD 3G band 1 (UMTS) due to its close proximity to the DVB-SH signal band in S-band.

B.3 Interference due to uplink signal

B.3.1 Desensitization

The transmitted cellular or connectivity signal has a very high power compared to the received DVB-SH signals. FDD 3G band 1 is the closest one with high power and will be considered as the worst-case situation.

FDD 3G band 1 transmitted power is +24 dBm. Part of this power is coupled to the DVB-SH antenna. The coupling loss is between the FDD 3G band 1 antenna and the DVB-SH antenna is assumed to be 9 dB. The received power at the DVB-SH antenna is therefore +15 dBm. Without any filtering the cellular TX signal present in the DVB-SH receiver input would be also +15 dBm. This very high interference signal level would cause severe blocking effects and reciprocal mixing.

The practical solution for co-existence is to insert a rejection filter in front of the DVB-SH receiver. The filter has to attenuate the FDD 3G band 1 Tx-signal to the allowed out of band unwanted signal level. The maximum level for hand portable terminal is -25 dBm. Stop band attenuation of the filter thus becomes:

$$\boxed{A_{filter} = P_{TX} - A_A - P_{max} = +24\,dBm - 9dB - (-25\,dBm) = 40\,dB} \tag{B.1}$$

where:
- A_{filter} = stop band attenuation of the reject filter;
- P_{TX} = Tx output power;
- A_A = coupling between the antennas (-9 dB);
- P_{max} = maximum allowed power at the DVB-SH receiver input.

Typically the insertion loss of the filter is in the order of 1,5 dB as described in clause 10. The sensitivity degradation is affected by the attenuation of the RF filter, by the receiver desensitization and phase noise performances and by antenna coupling between the two antennas at the blocker frequency.

B.3.2 Spurious and transmitted PA noise

Besides the wanted part of the FDD 3G band 1 uplink signal, the signal also contains unwanted parts like spurious responses and the out of band PA noise for terminal architectures without duplexer.

The GSM specification (TS 100 910 [30]) defines that within 100 kHz measurement bandwidth the power shall not be greater than -79 dBm within GSM900 Rx frequency band (-71 dBm in case of DCS1800). This noise floor corresponds to the noise contribution of the transmitter power amplifier. That value can be extended for a noise floor value within the DVB-SH bands as a maximum value.

Assuming a 18 dB coupling loss between GSM900 and DVB-SH S-band antennas, the interference power entering DVB-SH receiver would be -80 dBm for 5 MHz bandwidth. With a 4,5 dB receiver noise figure, the total receiver input noise floor is –102,7 dBm. From this it is obvious that the transmitter output noise reduces the DVB-SH receiver sensitivity considerably. In order to degrade DVB-SH receiver sensitivity "only" by 3 dB the transmitter output noise would need to be lower or equal to the receiver thermal noise floor (-102,7 dBm).

Fortunately, noise in S band from GSM quadriband PA is lower than ETSI specification (20 dB typically). Nevertheless, if DVB-SH high sensitivity performance is targeted, an additional filtering will be required on TX power amplifier outputs.

Interoperability requirements

To guarantee interoperability between the radio systems the noise power at the DVB-H receiver input (@ 5 MHz band) must fulfil the mask shown in figure B.2.

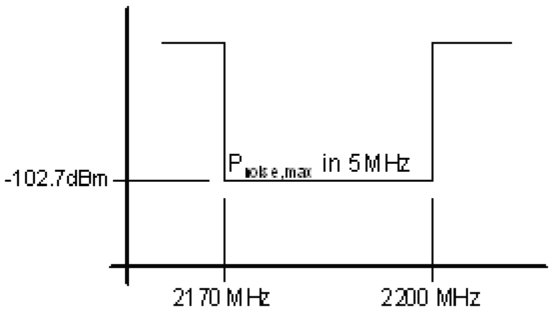


Figure B.2: Tx PA-noise Mask in S-band DVB-SH Receiver Input

The noise level is affected by the noise power of the power amplifier, by the attenuation of the possible low pass filter at the output of the transmitter PA as shown in figure B.3 and by antenna coupling between the two antennas at the DVB-SH reception frequency band.

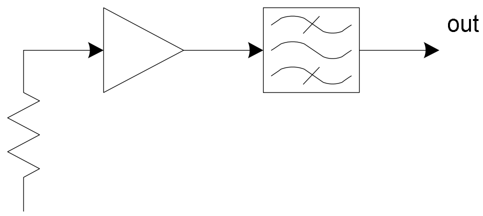


Figure B.3: GSM Tx Block Diagram

B.4 Cellular Radio Downlink Signal Interference to DVB-SH Receiver

The location of the DVB-SH received signal in S-band is next to the UMTS FDD downlink frequency range. This represents an additional interferer. Consequently, this is the most critical case for downlink signal interference.

B.4.1 Adjacent channel

Adjacent channel denomination will characterize hereafter 5 MHz adjacent channel (UMTS downlink channel next to DVB-SH channel) and per extension 10 MHz adjacent channel (one unoccupied UMTS downlink channel between UMTS downlink interferer and DVB-SH channel).

Adjacent interference is summarized through Adjacent Channel Interference Ratio (ACIR). This is defined by:

$$\frac{1}{ACIR} = \frac{1}{ACS} + \frac{1}{ACLR}$$

ACLR: Adjacent Channel Leakage Ratio which determines the power ratio transmitted by the UMTS downlink Base Station in its adjacent channel.

ACS: Adjacent Channel Selectivity which defines the ability of the DVB-SH receiver to demodulate the wanted signal in presence of interferer in its adjacent channel.

The UMTS specification TS 125 104 [15] gives a minimum of 45 dB ACLR for 5 MHz adjacent channel and a minimum of 50 dB ACLR for 10 MHz adjacent channel.

B.4.2 ACS requirements

Minimum receiver ACS higher than 50 dB for 5 MHz adjacent channel and higher than 60 dB for 10 MHz adjacent channel should be achieved in order to avoid DVB-SH receiver ACS contribution in ACIR value considering implementation margins in UMTS downlink ACLR (ACIR degradation less than 1 dB for 45 dB ACLR).

Annex C (informative): Spectrum efficiency and system throuhput

C.1 Spectrum efficiency analysis

The following spectral efficiency calculations shall be considered as illustrative and by no means representative of real systems. In particular, the frequency plan is assumed to be homogenous over the whole coverage region. Finally when the common content retransmission takes place in another frequency sub-band (MFN operations) we assumed that the terrestrial physical and upper layer parameters can be different to the satellite ones. This allows inclusion of local content on top of the common content in the terrestrial retransmission.

When in MFN (and only MFN), a possible issue arises for the hybrid frequency (i.e. the terrestrial sub-bands carrying both common and local contents). Indeed, these frequencies are common to an adjacent satellite beam and the terrestrial cells that make use of them are not guaranteed to be far away from the foot print of this adjacent beam (it can be inside the "adjacent" foot print due to overlap). The "Split band" technique was introduced to overcome this interference issue. It consists in partitioning into two the satellite frequency slot, one part for the satellite downlink and the other for the hybrid terrestrial transmission. This approach reduces spectral efficiency for the common content.

The following configurations can be identified.

Table C.1

	Waveform	SFN/MFN	Split of satellite beam slot for minimization of exclusion zones
Config 1	SH-A	MFN	Yes
Config 2	SH-A	MFN	No
Config 3	SH-A	SFN	No
Config 4	SH-B	MFN	Yes
Config 5	SH-B	MFN	No

Examples for a 3-color reuse pattern of a 3x5 MHz satellite system are depicted below. The drawings show the frequency plan in a particular beam, with "Slot A" designating the frequency assigned to the satellite beam in question (this is the "satellite-protected" frequency, in which no adjacent beam interference occurs). Not the drawings are not to scale and that the frequency slots can be of different width in the different cases.

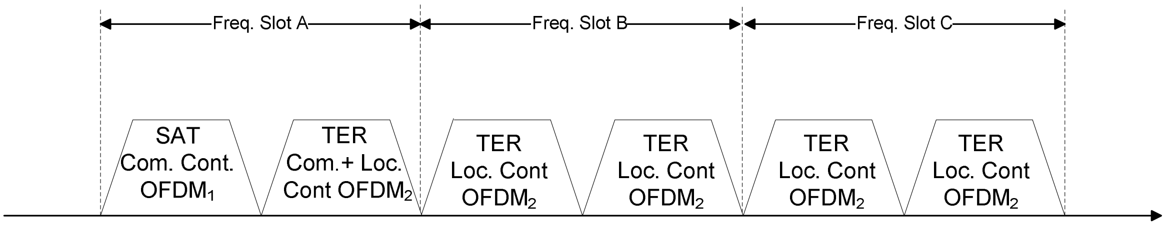


Figure C.1: Configuration 1 (SH-A, MFN, Split band)

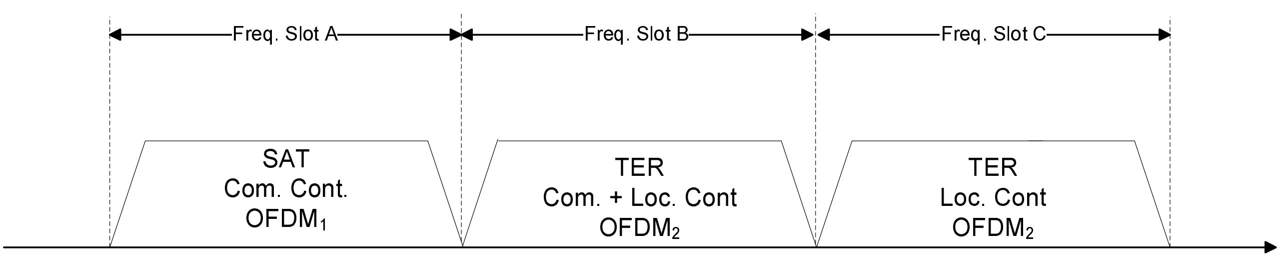


Figure C.2: Configuration 2 (SH-A, MFN, No split band)

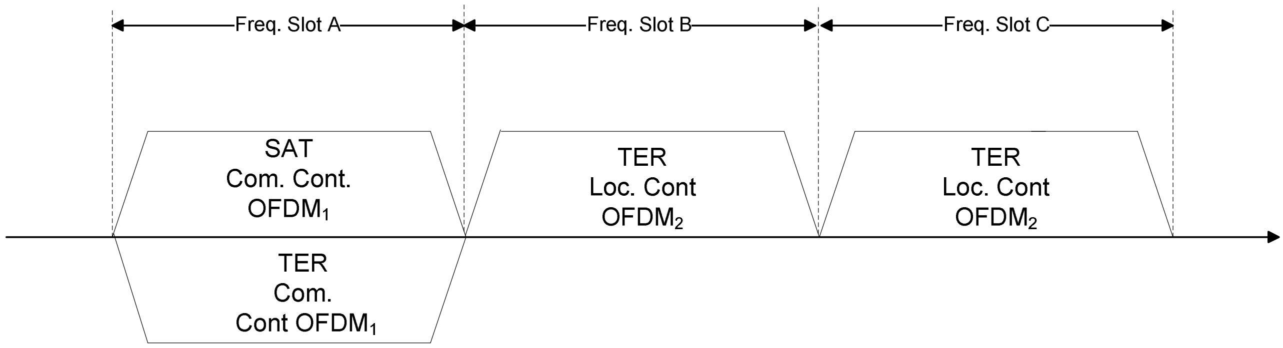


Figure C.3: Configuration 3 (SHA, SFN, No Split band)

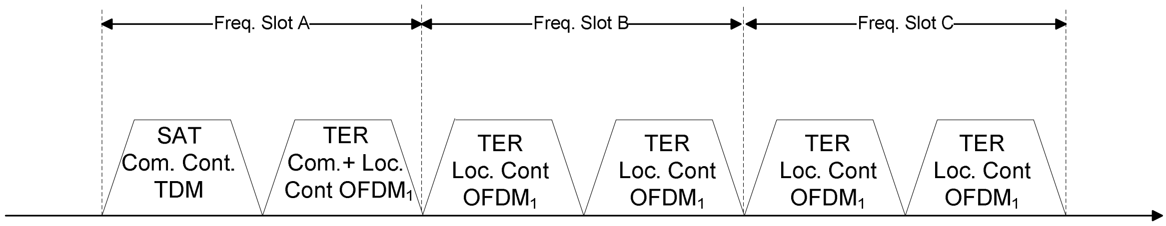


Figure C.4: Configuration 4 (SH-B, MFN, split band)

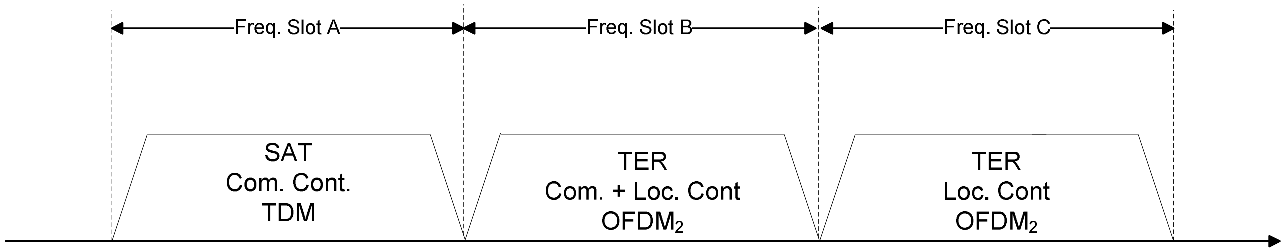


Figure C.5: Configuration 5 (SH-B, MFN, No split band)

C.1.1 Spectral Efficiency Calculations

We define the *waveform* spectral efficiency as the net bit rate of each waveform divided by the channelization bandwidth. The following equations give the spectral efficiencies and the net bit rate as a function of modulation and coding parameters:

for TDM:

the net bit rate supported by the TDM carrier $Cap\_TP[TDM]$ is computed in clause 7. The TDM spectral efficiency is then:

$$\eta(\mathrm{TDM}) = Cap\_TP[TDM] / B_W^{TDM},$$

being B_W^{TDM} the TDM channelization bandwidth;

for OFDM:

the net bit rate supported by the OFDM carrier $Cap\_TP[OFDM]$ is computed in clause 7. The OFDM spectral efficiency is then:

$$\eta(\mathrm{OFDM}) = Cap\_TP[OFDM] / B_W^{OFDM},$$

being B_W^{OFDM} the OFDM channelization bandwidth.

Table C.2 contains numerical examples for the bit rate and spectral efficiencies calculations for TDM and OFDM waveforms having assumed:

channelization bandwidth = B_W^{OFDM} = B_W^{TDM} = 5 MHz;

TDM roll-off factor α_{TDM} = 0,15;

OFDM guard interval GI = 1/8.

Table C.2

TDM, QPSK, FEC_rate 1/3	Net bit rate in 5 MHz= 2,633 Mbps	η(TDM) = 0,53 b/s/Hz
TDM, 16APSK, FEC_rate ¼	Net bit rate in 5 MHz= 3,950 Mbps	η(TDM) = 0,79 b/s/Hz
OFDM, QPSK, FEC_rate 1/3	Net bit rate in 5 MHz= 2,468 Mbps	η(OFDM) = 0,49 b/s/Hz
OFDM, 16QAM, FEC_rate ¼	Net bit rate in 5 MHz= 3,730 Mbps	η(OFDM) = 0,75 b/s/Hz

We are interested in two spectral efficiencies at beam level:

- the *Total spectral efficiency per beam*, η(TOTAL), defined as the net bit rate used for all content (Common+Local) available in the CGC coverage of a beam divided by the *satellite-protected* bandwidth allocated to that beam;
- the *Common content spectral efficiency per beam*, η(COM), defined as the net bit rate of the Common content available in the beam divided by the same bandwidth as above.

For the clarity of the formulas below, we use the following notations to further distinguish between different use cases of the OFDM waveform:

- η(OFDM): Efficiency of the OFDM waveform when used in CGC only;
- η(OFDM,Sat): Efficiency of the OFDM waveform when used in satellite, without SFN;
- η(OFDM,Hyb): Efficiency of the OFDM waveform when used jointly in satellite and CGC (SFN).

Let f_R be the colour reuse factor (equivalently, the total number of sub-bands in a beam). Table C.3 gives the formulas for computing these efficiencies, once the waveform spectral efficiencies and the satellite frequency reuse have been selected.

Table C.3

Configuration	Total spectral efficiency per beam, η(TOTAL)	Common content spectral efficiency per beam, η(COM)
1) SH-A, MFN, Split	(f_R – ½) x η(OFDM)	½ x η(OFDM,Sat)
2) SH-A, MFN, No split	(f_R -1) x η(OFDM)	η(OFDM,Sat)
3) SH-A, SFN, No Split	(f_R -1) x η(OFDM) + η(OFDM,Hyb)	η(OFDM,Hyb)
4) SH-B, MFN, Split	(f_R –1/2) x η(OFDM)	½ x η(TDM)
5) SH-B, MFN, No Split	(f_R -1) x η(OFDM)	η(TDM)

Comments:

In deriving the above formulas for MFN, it has been implicitly assumed that the terrestrial efficiency is always higher than or equal to the satellite efficiency so that local content can be included on top of the common content in the terrestrial retransmission.

In configuration 3), if the Hybrid frequency sub-band uses the same modulation and coding as the other terrestrial sub-bands then η(OFDM) = η(OFDM, Hyb) and the Total spectral efficiency per beam is f_R x η(OFDM). The terrestrial frequency reuse in a beam is exactly the satellite frequency reuse between beams.

Note that when the satellite-protected frequency is split, the waveform efficiency is usually higher since the spectral density of the signal is 3 dB higher than in the non-split case. This has not been made explicit in the formulas above (the η(OFDM) of Config 1 and Config 2 should not be the same).

If Config 3 is to be compared with Config 5, their Total spectral efficiency per beam differs by the η(OFDM,Hyb), in favour of Config 3. As concerns the Common content spectral efficiency per beam, η(TDM) is usually higher than η(OFDM,Hyb), in favour of Config 5.

If the satellite system is composed of N_{beams}, the *system spectral efficiency* can be defined as the total bitrates summed over all beams divided by the total frequency allocated to that satellite system. The following relationship can be easily shown:

$$\text{System spectral efficiency} = (N_{beams}/f_R) \text{ x (Total spectral efficiency per beam)}$$

C.1.2 Numerical examples

In the following example of efficiency comparison between the different 5 possible configurations in a 6-beam, 3-color reuse system. For sake of simplicity the waveform used in all cases has a 5 MHz bandwidth. For this reason the split configurations 1) and 4) require a 10 MHz beam sub-slot bandwidth instead of 5 MHz as for cases 2), 3) and 5).

Table C.4

TDM, QPSK, FEC 1/3	η(TDM) = 0,53 b/s/Hz	Satellite in Config 4 and 5
OFDM, QPSK, GI=1/8, FEC 1/3	η(OFDM) = 0,49 b/s/Hz	Sat/Hybrid in Config 1, 2, 3
OFDM, QPSK, GI=1/8, FEC ½	η(OFDM) = 0,75 b/s/Hz	CGC only

Table C.5

Configuration	η(TDM)	η(OFDM)	η(OFDM) Hybrid or Sat	η(COM)	η(TOTAL) with f_R =3	System (6 beams)
1) SH-A, MFN, Split	NA	0,75	0,49	0,25	(3 - ½) x 0,49 = 1,23	6/3 x 1,23 = 2,46
2) SH-A, MFN, No split	NA	0,75	0,49	0,49	2 x 0,75 = 1,5	6/3 x 1,5 = 3
3) SH-A, SFN, No Split	NA	0,75	0,49	0,49	2 x 0,75 + 0,49 = 1,99	6/3 x1,99 = 3,98
4) SH-B, MFN, Split	0,53	0,75	NA	0,27	(3 - ½) x 0,53 =1,33	6/3 x 1,33 = 2,66
5) SH-B, MFN, No Split	0,53	0,75	NA	0,53	2 x 0,75 = 1,5	6/3 x1,5 = 3

The efficiencies above can be converted into net bitrates as follows, assuming a 5 MHz waveform channelization.

Table C.6

Configuration	Total system bandwidth/Beam sub-slot bandwidth	Common bit rate in a beam	Total bit rate in a beam	System (6 beams)
1) SH-A, MFN, Split	30/10 MHz	2,47 Mbps	12,3 Mbps	73,8 Mbps
2) SH-A, MFN, No split	15/5 MHz	2,47 Mbps	7,5 Mbps	45 Mbps
3) SH-A, SFN, No Split	15/5 MHz	2,47 Mbps	10 Mbps	60 Mbps
4) SH-B, MFN, Split	30/10 MHz	2,63 Mbps	13,3 Mbps	79,8 Mbps
5) SH-B, MFN, No Split	15/5 MHz	2,63 Mbps	7,5 Mbps	45 Mbps

It should be remarked that the total bit rate for cases 1) and 4) (split) is higher than cases 2), 3), 5) (no split) because the allocated overall bandwidth has been doubled. The spectral efficiencies for the split cases are however smaller than the non split ones.

C.2.3 Conclusions on System Spectrum Efficiency

It is difficult to provide general recommendations on the best system frequency planning configuration as decision depends not only on spectral efficiency. In fact when deciding the optimal frequency planning one should also consider other system parameters such satellite antenna beam patters and gap fillers location which may impact the location and extent of exclusion zones (see clause 11). Based on the examples illustrated before from a pure spectral efficiency perspective the following conclusions can be derived:

- when the content to be broadcasted is different for different regions (e.g. linguistic regions) multibeam satellite configuration increase the common content spectral efficiency proportionally to the ratio between the number of beams and the frequency reuse factor $\left(N_{beams} / f_R\right)$;
- SH-A/SFN configuration provides the highest overall spectral efficiency thanks to SFN operation between satellite and terrestrial gap fillers;
- SH-B/No-split achieves a slightly higher spectral efficiency than SH-A for common content delivery but has a lower spectral efficiency for local content distribution than SH-A. Power efficiency issues (e.g. HPA nonlinearity effects) are not part of this comparison.

Any split configuration provides half of the spectral efficiency for common content distribution compared to the corresponding non split ones. This strong efficiency reduction is balanced by potential advantages in terms of exclusion zones for common content delivery (see clause 4). Furthermore the split configuration makes it possible to increase the satellite C/N in power limited cases.

As stated above the above discussion is only looking at the spectral efficiency aspects. More extensive discussion about SH-A and SH-B waveform pro and contra can be found in clause 7.

C.2 Throughput calculations

DVB-SH capacity at MPEG TS interface level may be calculated from the parameters defined in the waveform definition [1]. Specifically, it is found by first calculating the duration of the SH frame, and the number of MPEG TS packets in an SH Frame, and then using these results to determine the capacity.

To support seamless hand-over between OFDM and TDM, the SH frame length of the TDM part has been aligned to the SH frame length in OFDM. Therefore, the TDM parameters can not be calculated independently but imply a selection of OFDM parameters which is represented in the following equations.

C.2.1 Reference DVB-SH parameters

The throughput calculation is based on some reference DVB-SH parameters which are recalled in this section.

C.2.1.1 OFDM Frame

The time related parameters of the OFDM frame are defined in multiple of the elementary period *T*. From EN 302 583 [1], clause 5.7.4.1,

$$T = \frac{7}{8 \cdot BW\_OFDM} \qquad \text{(C.1)}$$

with BW_OFDM being the OFDM signal bandwidth, in MHz; BW_OFDM=1,6 MHz (for a "1,7 MHz" channel), 5 MHz, 6 MHz, 7 MHz or 8 MHz.

NOTE 1: *T* is in µS in the above equation; alternatively, *T* is in seconds if *BW_OFDM* is in Hz.

NOTE 2: the elementary period of the 1,7 MHz Channel, defined in EN 302 583 [1], clause 5.7.4, is equal to 7/12,8 µs; this corresponds to a *BW_OFDM* of 1,6 MHz.

The duration *Ts* of an OFDM symbol, defined in EN 302 583 [1], clause 5.7.4.1, is:

$$T_S = 1024 \cdot Mod\_FFT \cdot (1+GI) \cdot T \qquad \text{(C.2)}$$

with *Mod_FFT* being the FFT length in k; Mod_FFT = 1, 2, 4 or 8 and GI being the Guard Interval ratio; GI=1/4,1/8,1/16 or 1/32.

The number *Nb_Symb_OFDM_Frame* of OFDM symbol per OFDM frame, defined in EN 302 583 [1], clause 5.7.4.1, is:

Nb_Symb_OFDM_Frame = 68 (C.3)

The number of data subcarriers Nb_Data_Carrier per OFDM symbol, defined in EN 302 583 [1], clause 5.7.2, is:

Nb_Data_Carrier = 756 Mod_FFT (C.4)

with Mod_FFT being the FFT length in k; Mod_FFT = 1, 2, 4 or 8.

The number of bits per modulated data subcarrier in OFDM BpS_OFDM is:

BpS_OFDM=2 or 4 (C.5)

with BpS_OFDM= 2 for QPSK modulation, and 4 for 16QAM modulation.

The SH Frame is, as defined in EN 302 583 [1], clause 5.5.2.3, composed of 816 Capacity Units (CU) of 2 016 bit each, thus:

Nb_CU_SHF_OFDM= 816 (C.6)

Nb_Bit_CU=2 016 (C.7)

C.2.1.2 TDM Frame

The TDM signal symbol rate defined in EN 302 583 [1], clause 5.6.3 is calculated with the following formula:

$$Sr\_TDM = \mathrm{int}\left\{32\left(\frac{1+GI}{1+\alpha}\right)\right\}\frac{BW\_TDM}{32(1+GI)} \qquad \text{(C.8)}$$

with α being the TDM Square-Root Raised-Cosine (SRRC) filter roll off factor; α =0,15; 0,25; or 0,35, GI being the OFDM guard interval, *BW_TDM* being the TDM signal bandwidth with *BW_TDM=BW_OFDM.*

NOTE: The rationale of this equation is that the TDM symbol rate may be simply derived from the OFDM elementary period *T*, *in order to* optimize the receiver design. Using this definition as well as (1) and (2):

$$Sr\_TDM \cdot T = \mathrm{int}\left\{32\left(\frac{1+GI}{1+\alpha}\right)\right\}\left[\frac{7}{8*32(1+GI)}\right]$$

result is a rational number, whatever the GI and α values.

Thus *Sr_TDM* may be easily generated in any case from the 1/*T* clock reference.

As defined in EN 302 583 [1], clause 5.6.4, the TDM SH Frame is composed of Physical Layer Slots (PL SLOTS) of *Nb_Symb_TDM_SLOT* symbols:

Nb_Symb_TDM_SLOT=2 176 (C.9)

Each PL SLOT carries 2, 3, or 4 CU, equal to the order of modulation *BpS_TDM* (see EN 302 583 [1], clause 5.6.4.1):

Nb_CU_PLSLOT=BpS_TDM =2, 3 or 4 (C.10)

C.2.2 Calculation of the SH Frame duration

The SH frame length is defined as a function of the length of OFDM frames. Therefore, the frame length durations of both an SH-frame in OFDM and TDM modes are identical. The number of OFDM Frames per SH frame is defined in the table 5.10 of the waveform definition [1]. It may be calculated from the equations (C.3) to (C.7) from above. The related equation is the following:

$$Nb\_OFF\_SHF = \frac{Nb\_CU\_SHF\_OFDM \cdot Nb\_Bit\_CU}{BpS\_OFDM \cdot Nb\_Data\_Carrier \cdot Nb\_Symb\_OFDM\_Frame}$$

where the numerator is the number of data bits per SH Frame, and the denominator is the number of bits per OFDM Frame.

Using (3)-(7), the above equation reduces to:

$$Nb\_OFF\_SHF = \frac{32}{Mod\_FFT \cdot BpS\_OFDM} \qquad \text{(C.11)}$$

So, the equation of the SH Frame duration is the following:

$$T_{SHF} = Nb\_OFF\_SHF \cdot Nb\_Symb\_OFDM\_Frame \cdot T_S \qquad \text{(C.12)}$$

Using (2), (3) and (8), it reduces to:

$$T_{SHF} = 2176 * 896 \cdot \frac{1+GI}{BpS\_OFDM \cdot BW\_OFDM}, \qquad \text{(C.13)}$$

with T_{SHF} in µS.

C.2.3 Calculation of the number of MPEG TS packets per SH Frame

Each turbo code word contains exactly 8 MPEG TS Packets. Thus, the number of MPEG TS packets per SH Frame is:

$$Nb\_TP\_SHF = Nb\_TP\_CW \cdot Nb\_CW\_SHF, \qquad \text{(C.14)}$$

with *Nb_CW_SHF* being the number of FEC code words per SH frame, and *Nb_TP_CW* being equal to 8.

Following FEC coding and rate adaptation, the FEC codeword length in bits is:

$$Nb\_Bit\_CW = \frac{Nb\_Bit\_CW\_0}{cr},$$

with Nb_Bit_CW_0 = 12 096, that is the codeword length assuming a code rate of 1, and cr being the nominal FEC code rate, as defined in EN 302 583 [1], clause 5.3.1.

The number of code word per SH Frame is related to the number of CU per SH Frame:

$$Nb\_CW\_SHF = \mathrm{int}\left(\frac{Nb\_CU\_SHF}{Nb\_Bit\_CW / Nb\_Bit\_CU}\right)$$

with *Nb_CU_SHF* being the number of CU per SH Frame, int*(x)* being a function which returns the largest integer less than the value *x*.

$$Nb\_CW\_SHF = \mathrm{int}\left(\frac{Nb\_CU\_SHF \cdot Nb\_Bit\_CU \cdot cr}{Nb\_Bit\_CW\_0}\right), \quad \text{(C.15)}$$

NOTE: $\frac{Nb\_Bit\_CW\_0}{Nb\_Bit\_CU}$ simplifies to 6 (a codeword occupies 6/cr CUs).

For OFDM, as defined in (6), the number of CU per SH frame is:

$$Nb\_CU\_SHF\_OFDM = 816 \quad \text{(C.16)}$$

For TDM, the number of CU per SH frame is given in EN 302 583 [1], clause 5.6.1. It is obtained from the following calculation:

$$Nb\_CU\_SHF\_TDM = Sr\_TDM \cdot T_{SHF} \cdot \frac{Nb\_CU\_PLSLOT}{Nb\_Symb\_TDM\_SLOT} \quad \text{(C.17)}$$

Using (8), (9), (10) and (13), it reduces to the following expression:

$$Nb\_CU\_SHF\_TDM = 28 \cdot \mathrm{int}\left(32\frac{1+GI}{1+\alpha}\right) \cdot \frac{BpS\_TDM}{BpS\_OFDM} \quad \text{(C.18)}$$

So, the number of MPEG TS packets per SH Frame is:

for OFDM, using (14), (15) and (16):

$$Nb\_TP\_SHF = 8 \cdot \mathrm{int}(136 \cdot cr) \quad \text{(C.19)}$$

for TDM, using (14), (15) and (18):

$$Nb\_TP\_SHF = 8 \cdot \mathrm{int}\left(\mathrm{int}\left(32\frac{1+GI}{1+\alpha}\right) \cdot \frac{28 \cdot cr \cdot BpS\_TDM}{6 \cdot BpS\_OFDM}\right) \quad \text{(C.20)}$$

C.2.4 Calculation of the capacity

An MPEG TS packet is 188 bytes long, that is *Nb_Bit_TP* bits.

$$\text{Nb\_Bit\_TP}=1\,504 \quad \text{(C.21)}$$

So the bit rate capacity at the MPEG-TS interface is:

$$Cap\_TP = Nb\_Bit\_TP \cdot \frac{Nb\_TP\_SHF}{T_{SHF}} \text{ [bps]}, \quad \text{(C.22)}$$

for OFDM, using (13), (19) and (22):

$$Cap\_TP\_OFDM = \frac{47 \cdot 8}{2176 \cdot 28} \cdot \frac{\mathrm{int}(136.cr)}{1+GI} \cdot BpS\_OFDM \cdot BW\_OFDM\,;$$

for TDM, using (13), (20) and (22):

$$Cap\_TP\_TDM = \frac{47 \cdot 8}{2176 \cdot 28} \cdot \frac{\mathrm{int}(\mathrm{int}(32 \cdot \frac{1+GI}{1+\alpha}) \cdot \frac{28.cr.BpS\_TDM}{6.BpS\_OFDM})}{1+GI} \cdot BpS\_OFDM \cdot BW\_OFDM\,.$$

History

Document history		
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