ETSI TR 101 290 V1.4.1 (2020-06)

TECHNICAL REPORT

Digital Video Broadcasting (DVB); Measurement guidelines for DVB systems

Reference
RTR/JTC-DVB-389

Keywords
broadcasting, digital, DVB, TV, video

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Foreword

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

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

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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.

Modal verbs terminology

In the present document "**should**", "**should not**", "**may**", "**need not**", "**will**", "**will not**", "**can**" and "**cannot**" are to be interpreted as described in clause 3.2 of the ETSI Drafting Rules (Verbal forms for the expression of provisions).

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1 Scope

The present document provides guidelines for measurement in Digital Video Broadcasting (DVB) satellite, cable and terrestrial and related digital television systems. The present document defines a number of measurement techniques, such that the results obtained are comparable when the measurement is carried out in compliance with the appropriate definition.

The present document uses terminology used in ETSI EN 300 421 [i.5], ETSI EN 300 429 [i.6], ETSI EN 300 468 [i.7] and ETSI EN 300 744 [i.9] and it should be read in conjunctions with them.

2 References

2.1 Normative references

Normative references are not applicable in the present document.

2.2 Informative references

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

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

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

[i.1] ISO/IEC 13818-1 (ITU-T Recommendation H.222.0): "Information technology - Generic coding of moving pictures and associated audio information: Systems".

[i.2] ISO/IEC 13818-4: "Information technology - Generic coding of moving pictures and associated audio information - Part 4: Conformance testing".

[i.3] ISO/IEC 13818-9: "Information technology - Generic coding of moving pictures and associated audio information - Part 9: Extension for real time interface for systems decoders".

[i.4] Void.

[i.5] ETSI EN 300 421: "Digital Video Broadcasting (DVB); Framing structure, channel coding and modulation for 11/12 GHz satellite services".

[i.6] ETSI EN 300 429: "Digital Video Broadcasting (DVB); Framing structure, channel coding and modulation for cable systems".

[i.7] ETSI EN 300 468: "Digital Video Broadcasting (DVB); Specification for Service Information (SI) in DVB systems".

[i.8] ETSI TR 101 211: "Digital Video Broadcasting (DVB); Guidelines on implementation and usage of Service Information (SI)".

[i.9] ETSI EN 300 744: "Digital Video Broadcasting (DVB); Framing structure, channel coding and modulation for digital terrestrial television".

[i.10] EN 50083-9: "Cable networks for television signals, sound signals and interactive services - Part 9: Interfaces for CATV/SMATV headends and similar professional equipment for DVB/MPEG-2 transport streams", produced by CENELEC.

[i.11] Void.

[i.12] Recommendation ITU-T O.151: "Error performance measuring equipment operating at the primary rate and above".

[i.13] ETSI EN 300 473: "Digital Video Broadcasting (DVB); Satellite Master Antenna Television (SMATV) distribution systems".

[i.14] ETSI TS 101 191: "Digital Video Broadcasting (DVB); DVB mega-frame for Single Frequency Network (SFN) synchronization".

[i.15] ETSI EN 300 748: "Digital Video Broadcasting (DVB); Multipoint Video Distribution Systems (MVDS) at 10 GHz and above".

[i.16] ETSI EN 300 749: "Digital Video Broadcasting (DVB); Microwave Multipoint Distribution Systems (MMDS) below 10 GHz".

[i.17] ISO 639: "Code for the representation of names of languages".

[i.18] ETSI EN 301 210: "Digital Video Broadcasting (DVB); Framing structure, channel coding and modulation for Digital Satellite News Gathering (DSNG) and other contribution applications by satellite".

[i.19] ETSI ETS 300 813: "Digital Video Broadcasting (DVB); DVB interfaces to Plesiochronous Digital Hierarchy (PDH) networks".

[i.20] ETSI ETS 300 814: "Digital Video Broadcasting (DVB); DVB interfaces to Synchronous Digital Hierarchy (SDH) networks".

[i.21] Void.

[i.22] Void.

[i.23] EN 50221: "Common interface specification for conditional access and other digital video broadcasting decoder applications", produced by CENELEC.

[i.24] ETSI TS 102 773 (V1.3.1), January 2012: "Digital Video Broadcasting (DVB); Modulator Interface (T2-MI) for a second generation digital terrestrial television broadcasting system (DVB-T2)".

[i.25] ETSI TS 102 034 (August 2009): "Digital Video Broadcasting (DVB); Transport of MPEG-2 TS Based DVB Services over IP Based Networks".

[i.26] SMPTE 2022-1 (May 2007): "Forward Error Correction for Real-Time Video/Audio Transport Over IP Networks".

[i.27] ETSI EN 302 755 (V1.3.1) (April 2012): "Digital Video Broadcasting (DVB); Frame structure channel coding and modulation for a second generation digital terrestrial television broadcasting system (DVB-T2)".

[i.28] ETSI EN 302 769 (V1.2.1) (April 2011): "Digital Video Broadcasting (DVB); Frame structure channel coding and modulation for a second generation digital transmission system for cable systems (DVB-C2)".

[i.29] ETSI TS 102 991 (V1.2.1) (June 2011): "Digital Video Broadcasting (DVB); Implementation Guidelines for a second generation digital cable transmission system (DVB-C2)".

[i.30] ETSI TS 101 154: "Digital Video Broadcasting (DVB); Specification for the use of Video and Audio Coding in Broadcast and Broadband Applications".

[i.31] ISO/IEC 13818-2: "Information technology - Generic coding of moving pictures and associated audio information - Part 2: Video".

[i.32] ISO/IEC 13818-3: "Information technology - Generic coding of moving pictures and associated audio information - Part 3: Audio".

[i.33] ETSI EN 301 192: "Digital Video Broadcasting (DVB); DVB specification for data broadcasting".

[i.34] IETF RFC 791: "Internet Protocol".

[i.35] IETF RFC 768: "User Datagram Protocol".

[i.36] IETF RFC 3171: "IANA Guidelines for IPv4 Multicast Address Assignments".

[i.37] IETF RFC 4445: "A Proposed Media Delivery Index (MDI)".

[i.38] Proakis John G.: "Digital Communication", McGraw Hill, 1989.

[i.39] Begin G., Haccoun D. and Chantal P.: "High-Rate Punctured Convolutional Codes for Viterbi and Sequential Decoding", IEEE Trans. Commun, vol 37, pp. 1113-1125, November 1989.

[i.40] Begin G., Haccoun D. and Chantal P.: "Further Results on High-Rate Punctured Convolutional Codes for Viterbi and Sequential Decoding", IEEE Trans. Commun., vol 38, pp. 1922-1928, November 1990.

[i.41] Odenwalder J.P.: "Error Control Coding Handbook", Final report prepared for United States Airforce under Contract No. F44620-76-C-0056, 1976.

[i.42] Pratt, Timothy and Bostian Charles W.: "Satellite Communications", John Wiley & Sons, 1986.

[i.43] SMPTE 2022-2:2007: "Unidirectional Transport of Constant Bit Rate MPEG-2 Transport Streams on IP Networks".

[i.44] COST 207: "Digital land mobile radio communications".

NOTE: Available at https://op.europa.eu/en/publication-detail/-/publication/61fc77e7-bca2-4229-8eb4-77741f0d2ab2.

3 Definition of terms, symbols and abbreviations

3.1 Terms

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

MPEG-2: Refers to the ISO/IEC 13818, Systems coding is defined in ISO/IEC 13818-1 [i.1], video coding is defined in ISO/IEC 13818-2 [i.31] and audio coding is defined in ISO/IEC 13818-3 [i.32].

multiplex: stream of all the digital data carrying one or more services within a single physical channel

Service Information (SI): digital data describing the delivery system, content and scheduling/timing of broadcast data streams, etc.

NOTE: Includes MPEG-2 Program Specific Information (PSI) as defined in ISO/IEC 13818-1 [i.1] together with independently defined extensions.

Transport Stream (TS): data structure used in many of the Digital Video Broadcasting (DVB) related standards

NOTE: Defined in ISO/IEC 13818-1 [i.1].

3.2 Symbols

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

BW_{SYS}	System noise power bandwidth
F_H	Frequency (high)
F_L	Frequency (low)
K_{MAX}	Maximum Carrier Number
MIP_N	N-th Mega-frame Initialization Packet

P_B	(Variable name of) Bit Error Probability (BEP)
P_{BLOCK}	(Variable name of) the probability of an undetected error for a block of N symbols
P_I	Number of interleaving T2 frames
P_N	Variable for MIP pointer value
P_S	Symbol Error Probability
printf	symbol in the C programming language
P_{SIN}	Error probability of the incoming symbols
Q_S	Symbol error probability
R_I	Information rate
R_T	Transmission rate
STS_M	Value of M-th Synchronization Time Stamp
T_U	Symbol duration

3.3 Abbreviations

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

ACE	Active Constellation Extension
ACLR	Adjacent Channel Leakage Ratio
AFC	Automatic Frequency Control
AI	Amplitude Imbalance
ASCII	American Standard Code for Information Interchange
ASI	Asynchronous Serial Interface
ATM	Asynchronous Transfer Mode
AWGN	Additive White Gaussian Noise
BAT	Bouquet Association Table
BB	Baseband
BBFER	Baseband Frame Error Rate
BCH	Bose - Chaudhuri - Hocquenghem code
BEP	Bit Error Probability
BER	Bit Error Rate
bslbf	bit string, left bit first
BW	BandWidth
C/N	ratio of RF or IF signal power to noise power
CA	Conditional Access
CAT	Conditional Access Table
CATV	Community Antenna TeleVision
CBR	Constant Bit Rate
CCDF	Cumulative Complementary Distribution Function
CCI	Co-channel Interference
CF	Correction Factor
CFC	Number of Frame Closing symbols
CI	Common Interface
COFDM	Coded Orthogonal Frequency Division Multiplex
CPE	Common Phase Error
CRC	Cyclic Redundancy Check
CS	Carrier Suppression
CSO	Composite Second Order
CTB	Composite Triple Beat
CW	Continuous Wave
DC	Direct Current
DF	Delay Factor
DSNG	Digital Satellite News Gathering
DTG	Digital TV Group
DVB	Digital Video Broadcasting
DVB-C	Digital Video Broadcasting baseline system for digital cable television

NOTE: See ETSI EN 300 429 [i.6].

DVB-CS Digital Video Broadcasting baseline system for SMATV distribution systems

NOTE: See ETSI EN 300 473 [i.13].

DVB-MC Digital Video Broadcasting baseline system for Multi-point Video Distribution Systems below 10 GHz

NOTE: See ETSI EN 300 749 [i.16].

DVB-MG DVB Measurement Group
DVB-MS Digital Video Broadcasting baseline system for Multi-point Video Distribution Systems at 10 GHz and above

NOTE: See ETSI EN 300 748 [i.15].

DVB-S Digital Video Broadcasting baseline system for digital satellite television

NOTE: See ETSI EN 300 421 [i.5].

DVB-T Digital Video Broadcasting baseline system for digital terrestrial television

NOTE: See ETSI EN 300 744 [i.9].

DVB-X2 2^{nd} generation DVB systems
EB Errored Block
EIT Event Information Table
EIT-F Event Information Table - Future
EIT-P Event Information Table - Present
EMM Entitlement Management Message
ENB Equivalent Noise Bandwidth
END Equivalent Noise Degradation
ENF Equivalent Noise Floor
ES Errored Second
ESR Errored Second Ratio
ETI Errored Time Interval
ETIR Errored Time Interval Ratio
ETR ETSI Technical Report
ETS European Telecommunication Standard
EVM Error Vector Magnitude
EVM_V Error Vector Magnitude - Voltage
FEC Forward Error Correction
FEF Future Extension Frames
FFT Fast Fourier Transform
GI Guard Interval
GOP Group of Pictures
GPS Global Positioning System
GSE Generic Stream Encapsulation
GSM Global System for Mobile communications
HEX Hexadecimal
HP High Priority
IBS In-band signalling
ICI Inter-Carrier Interference
IEC International Electrotechnical Commission
IERS International Earth Rotation Service
IF Intermediate Frequency
IFFT Inverse FFT (Fast Fourier Transform)
IP Internet Protocol
IQ In-phase/Quadrature components
IRD Integrated Receiver Decoder
ISO International Organization for Standardization
ISSY Input Stream SYnchronizer
ITU International Telecommunication Union
LAT Link Available Time
LDPC Low Density Parity Check (codes)

LO	Local Oscillator
LP	Low Priority
LUAT	Link Unavailable Time
MDI	Media Delivery Index
MED	Maximum Excess Delay
MER	Modulation Error Ratio
MER_V	Modulation Error Ratio - Voltage
MFN	Multi-Frequency Network
MG	Measurement Guidelines
MGPR	MISO Group Power Ratio
MI	Modulator Interface
MIP	Mega-frame Initialization Packet
MISO	Multiple Input Single Output
MLR	Media Loss Rate
MMDS	Microwave Multi-point Distribution Systems (or Multi-channel Multi-point Distribution Systems)
MPEG	Moving Picture Experts Group
MPTS	Multi-Program Transport Stream
MR	Nominal Media Rate
MTU	Maximum Transmission Unit
MUX	Multiplex
MVDS	Multi-point Video Distribution Systems
NIT	Network Information Table
NM	Noise Margin
OB	Occupied Bandwidth
OFDM	Orthogonal Frequency Division Multiplex
PAPR	Peak to Average Power Ratio
PAT	Program Association Table
PCR	Program Clock Reference
PDH	Plesiochronous Digital Hierarchy
PER	Packet Error Rate
PID	Packet Identifier
PJ	Phase Jitter
PLL	Phase Locked Loop
PLP	Physical Layer Pipe
PMT	Program Map Table
PRBS	Pseudo Random Binary Sequence
PSI	MPEG-2 Program Specific Information

NOTE: As defined in ISO/IEC 13818-1 [i.1].

PTS	Presentation Time Stamps
QAM	Quadrature Amplitude Modulation
QAM-M	QAM systems with e.g. M = 16, 32 and 64
Q_B	Bit error probability
QE	Quadrature Error
QEF	Quasi Error Free
QPSK	Quaternary Phase Shift Keying
RBA	Receiver Buffer Assumptions
RBM	Receiver Buffer Model
RBW	Resolution Bandwidth
RC	FEC rate
REC	Receiver
RF	Radio Frequency
RFC	IETF Request For Comments
RMS	Root Mean Square
RS	Reed-Solomon
RST	Running Status Table

NOTE: See ETSI EN 300 468 [i.7].

RTE	Residual Target Error
RTP	Real Time Protocol

SDH	Synchronous Digital Hierarchy
SDP	Severely Disturbed Period
SDT	Service Description Table
SEP	Symbol Error Probability
SER	Symbol Error Rate
SES	Seriously Errored Second
SETI	Severely Errored Time Interval
SFN	Single Frequency Network
SI	Service Information
SINR	Signal to Interference Noise Ratio
SMATV	Satellite Master Antenna TeleVision
SMPTE	Society of Motion Picture and Television Engineers
SNR	Signal-to-Noise Ratio
ST	Stuffing Table
STD	System Target Decoder
STE	System Target Error
STS	Synchronization Time Stamp
SUS	Severely Uncorrectable Second
SUTI	Severely Uncorrectable Time Interval
SYNC	Synchronization
TDT	Time and Date Table
TEV	Target Error Vector
TFS	Time-Frequency-Slicing
TH	Transport Header
TOT	Time Offset Table
TPS	Transmission Parameter Signalling
TS	Transport Stream
TSTD	Transport Stream Description Table
TU	Typical Urban
TV	TeleVision
TX-SIG	Transmitter Signalling
UAT	Unavailable Time
UDP	User Datagram Protocol
UHF	Ultra-High Frequency
UI	Unit Interval
uimsbf	unsigned integer, most significant bit first
UP	Uncorrectable Packet
US	Uncorrectable Second
UTC	Universal Time Co-ordinated
UTI	Uncorrectable Time Interval
VB	Virtual Buffer
VBR	Variable Bit Rate
VBW	Video Bandwidth

4 General

The Digital Video Broadcasting (DVB) set of digital TV standards specify baseline systems for various transmission media: satellite, cable, terrestrial, etc. Each baseline system standard defined the channel coding and modulation schemes for that transmission medium. The source coding was adapted from the MPEG-2 standard.

The design of these new systems has created a demand for a common understanding of measurement techniques and the interpretation of measurement results.

The present document is an attempt to give recommendations in this field by defining a number of measurement techniques in such detail that the results are actually comparable as long as the measurement is carried out in compliance with the given definition.

Engineers seeking to apply the methods described in the present document should be familiar with the standards for the respective baseline systems. Although most of the parameters specified in the present document are well known in communications, most of them should be interpreted with respect to the new environment, especially the transmission of digital TV signals or other related services.

The inclusion of each parameter in the present document is based on requirements from those who envisage having to work alongside the defined procedures. This includes network operators and providers of equipment for network installation, as well as manufacturers of Integrated Receiver Decoders (IRDs) or test and measurement equipment.

The recommendations of the present document can be used:

- to set-up test beds or laboratory equipment for testing hardware for digital TV and other related services;
- to set these instruments to the appropriate parameters;
- to obtain unambiguous results that can be directly compared with results from other test set-ups;
- to form a potential basis for communicating results in an efficient way by using the definitions in the present document as references.

They are not intended to describe a set of compulsory tests.

The recommendations are grouped in several clauses. Since the MPEG-2 TS is the signal format used for the inputs and outputs of all baseline systems, clause 5 is devoted to the description of checking procedures for those parameters which are accessible in the TS packet header, i.e. without decoding scrambled or encrypted data. The aim of these tests is the provision of a simple and fast health check. It is meant neither as a MPEG-2 conformance test nor as a compliance test for all DVB related issues.

Clause 6 contains the parameters which are commonly addressed by various transmission media. For example, the measurement of the availability of transmission systems or links falls into this category, and it may be desirable to have the same definition for availability independent of the actual system in use.

Clauses 7 and 8 address the parameters which are specific for cable and satellite, DVB-C and DVB-S, they are also applicable to SMATV systems, DVB-CS, and possibly MMDS systems such as DVB-MC and DVB-MS.

Clause 9 addressed parameters specific to the terrestrial DVB environment (DVB-T).

Clauses 6, 7, 8, and 9 of the present document follow the same structure. For each parameter there is a description of the purpose of the recommended measurement procedure, the interface to which the measurement instrument should be applied, and a description of the actual method of the measurement itself.

Apart from these clauses a number of annexes are included, containing recommendations for general aspects, examples of test set-ups and certain requirements for the test and measurement equipment.

If the interfaces for a described measurement procedure are to be found within the transmitter, the notation is provided in accordance with figures 4.1 and 9.1 for DVB-T (the 1st generation terrestrial DVB system). If the interfaces for the described measurement procedures are to be found within the receiver (test receiver or IRD), the notation is provided in accordance with figures 4.2 and 9.2 for DVB-T. These figures illustrate the general cases of a DVB transmitter and receiver of the 1st generation systems, although certain functional blocks only appear in certain systems.

Most of the parameters can be measured with standard equipment such as spectrum analysers or constellation analysers. Other parameters are defined in a new way as a request to test and measurement equipment manufacturers to integrate this functionality in their products.

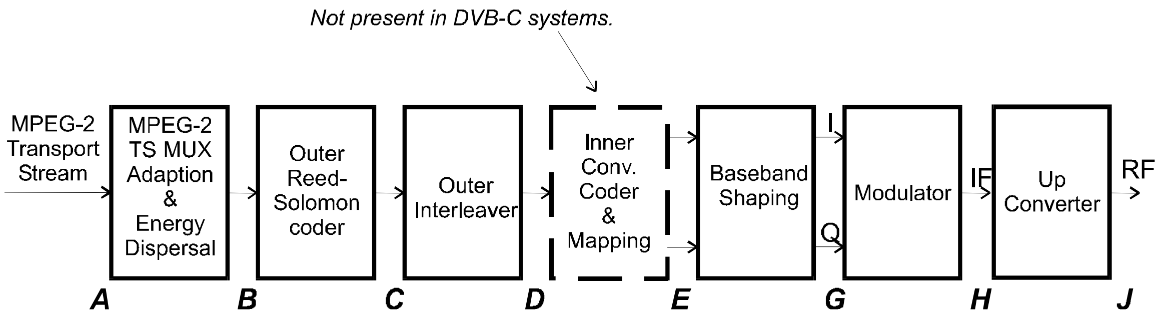


Figure 4.1: Transmitter block diagram

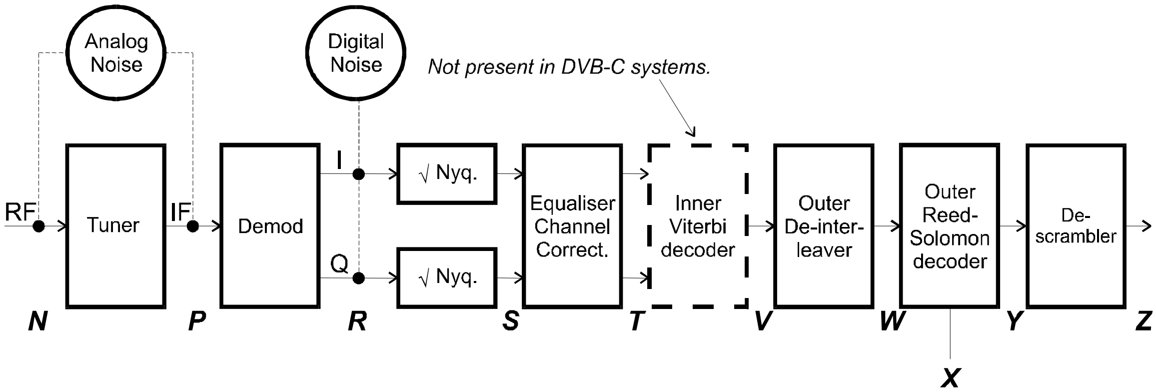


Figure 4.2: Receiver block diagram

The clauses 11 and 12 address two of the DVB transmission systems of the 2nd generation, DVB-T2 and DVB-C2. In clause 11.2 includes the definition of measurement parameters for the DVB-T2 Modulator Interface DVB-T2 MI.

The block diagrams for these 2nd generation transmission systems can be found in the respective clauses.

5 Measurement and analysis of the MPEG-2 Transport Stream

5.1 General

The MPEG-2 Transport Stream (TS) is the specified input and output signal for all the baseline systems, i.e. for satellite, cable, SMATV, MMDS/MVDS and terrestrial distribution, which are defined in the DVB world so far. Therefore these interfaces are accessible in the transmission chain. Direct access is given on the transmitter side at the input of the respective baseline system. At other interfaces where the signal occurs in modulated form, access is possible by an appropriate demodulator that provides the TS interface as an output for further measurements.

5.2 List of parameters recommended for evaluation

5.2.0 Introduction

The present document recommends in this clause a set of syntax and information consistency tests that can be applied to an MPEG-2 TS at the parallel interface, or either of the serial interfaces defined in EN 50083-9 [i.10].

The following assumptions and guiding principles were used in developing these tests:

- the tests are mainly intended for continuous or periodic monitoring of MPEG-2 TSs in an operational environment;

- these tests are primarily designed to check the integrity of a TS at source; clause 5.3 covers other aspects of TSs in networks including impairments created by transport systems;
- the general aim of the tests is to provide a "health check" of the most important elements of the TS. The list of the tests is not exhaustive;
- the tests are consistent with the MPEG-2 Conformance tests defined in ISO/IEC 13818-4 [i.2], they do not replace them;
- the tests are consistent with the DVB-SI documents (ETSI EN 300 468 [i.7], ETSI TR 101 211 [i.8]), they do not replace them.

MPEG-2 and DVB-SI reserved values in the TS do not cause a test error indication.

In general the tests are performed on TS header information so that they are still valid when conditional access algorithms are applied, however a few of the tests may only be valid for an unscrambled or descrambled TS.

The tests are not dependant on any decoder implementation for consistency of results. The MPEG-2 T-STD model constraints, as defined in ISO/IEC 13818-1 [i.1] (MPEG-2 Systems), should be satisfied as specified in ISO/IEC 13818-4 [i.2] (MPEG-2 Compliance).

Off-line tests are performed under stable conditions, no discontinuity or dynamic change can occur during an off-line test process.

Other digital performance parameters such as BER are not considered in this clause.

This clause tabulates the parameters which are recommended for continuous or periodic monitoring of the MPEG-2 TS.

The tests are grouped into three tables according to their importance for monitoring purposes.

The first table lists a basic set of parameters which are considered necessary to ensure that the TS can be decoded. The second table lists additional parameters which are recommended for continuous monitoring. The third table lists optional additional parameters which could be of interest for certain applications.

Any test equipment intended for the evaluation of these parameters should report test results by means of the indicators itemized in the second column of the tables under exactly the preconditions described in the third column of the tables.

If an indicator is set, then the TS is in error. However, since the indicators do not cover the entire range of possible errors, it cannot be concluded that there is no error if the indicator is not set.

If indicator 1.1 is activated then all other indicators are invalid. Each indicator is activated **only as long as** at least one of the described preconditions is fulfilled.

NOTE: In the case of indicators requiring a minimum repetition rate of sections, it is intended that each and every section that is present for this table should have the stated repetition rate.

5.2.1 First priority: necessary for de-codability (basic monitoring)

Table 5.0a: MPEG-2 TS parameters of 1st priority

No.	Indicator	Precondition	Reference
1.1	TS_sync_loss	Loss of synchronization with consideration of hysteresis parameters	ISO/IEC 13818-1 [i.1], clause 2.4.3.3 and annex G.1
1.2	Sync_byte_error	Sync_byte not equal 0x47	ISO/IEC 13818-1 [i.1], clause 2.4.3.3
1.3	PAT_error	PID 0x0000 does not occur at least every 0,5 s a PID 0x0000 does not contain a table_id 0x00 (i.e. a PAT) Scrambling_control_field is not 00 for PID 0x0000	ISO/IEC 13818-1 [i.1], clauses 2.4.4.3, 2.4.4.4
1.3.a (note 1)	PAT_error_2	Sections with table_id 0x00 do not occur at least every 0,5 s on PID 0x0000. Section with table_id other than 0x00 found on PID 0x0000. Scrambling_control_field is not 00 for PID 0x0000	ETSI TS 101 154 [i.30], clause 4.1.7 ISO/IEC 13818-1 [i.1], clauses 2.4.4.3, 2.4.4.4
1.4	Continuity_ count_error	Incorrect packet order a packet occurs more than twice lost packet	ISO/IEC 13818-1 [i.1], clauses 2.4.3.2, 2.4.3.3
1.5	PMT_error	Sections with table_id 0x02, (i.e. a PMT), do not occur at least every 0,5 s on the PID which is referred to in the PAT Scrambling_control_field is not 00 for all PIDs containing sections with table_id 0x02 (i.e. a PMT)	ISO/IEC 13818-1 [i.1], clauses 2.4.4.3, 2.4.4.4, 2.4.4.8
1.5.a (note 2)	PMT_error_2	Sections with table_id 0x02, (i.e. a PMT), do not occur at least every 0,5 s on each program_map_PID which is referred to in the PAT Scrambling_control_field is not 00 for all packets containing information of sections with table_id 0x02 (i.e. a PMT) on each program_map_PID which is referred to in the PAT	ETSI TS 101 154 [i.30], clause 4.1.7 (note 3) ISO/IEC 13818-1 [i.1], clauses 2.4.4.3, 2.4.4.4, 2.4.4.8
1.6	PID_error	Referred PID does not occur for a user specified period.	ISO/IEC 13818-1 [i.1], clause 2.4.4.8
NOTE 1: Recommended for future implementations as a replacement of 1.3. NOTE 2: Recommended for future implementations as a replacement of 1.5; this excludes specifically network_PIDs. NOTE 3: In ETSI TS 101 154 [i.30], it is recommended that the interval between two sections should not exceed 100 ms. For many applications it may be sufficient to check that the interval is no longer than 0,5 s.			

TS_sync_loss
The most important function for the evaluation of data from the MPEG-2 TS is the sync acquisition. The actual synchronization of the TS depends on the number of correct sync bytes necessary for the device to synchronize and on the number of distorted sync bytes which the device cannot cope with.

It is proposed that five consecutive correct sync bytes (ISO/IEC 13818-1 [i.1], clause G.1) should be sufficient for sync acquisition, and two or more consecutive corrupted sync bytes should indicate sync loss.

After synchronization has been achieved the evaluation of the other parameters can be carried out.

Sync_byte_error
The indicator "Sync_byte_error" is set as soon as the correct sync byte (0x47) does not appear after 188 or 204 bytes. This is fundamental because this structure is used throughout the channel encoder and decoder chains for synchronization. It is also important that every sync byte is checked for correctness since the encoders may not necessarily check the sync byte. Apparently some encoders use the sync byte flag signal on the parallel interface to control randomizer re-seeding and byte inversion without checking that the corresponding byte is a valid sync byte.

PAT_error
The Program Association Table (PAT), which only appears in PID 0x0000 packets, tells the decoder what programs are in the TS and points to the Program Map Tables (PMT) which in turn point to the component video, audio and data streams that make up the program (figure 5.2).

If the PAT is missing then the decoder can do nothing, no program is decodable.

Nothing other than a PAT should be contained in a PID 0x0000.

PAT_error_2
The reworded description of the error in PAT_error_2 refers to the possibility that the Program Association Table may consist of several (consecutive) sections with the same table_id 0x00.

Continuity_count_error
For this indicator three checks are combined. The preconditions "Incorrect packet order" and "Lost packet" could cause problems for IRD which are not equipped with additional buffer storage and intelligence. It is not necessary for the test equipment to distinguish between these two preconditions as they are logically OR-ed, together with the third precondition, into one indicator.

The latter is also covering the packet loss that may occur on ATM links, where one lost ATM packet would cause the loss of a complete MPEG-2 packet.

The precondition "a packet occurs more than twice" may be symptomatic of a deeper problem that the service provider would like to keep under observation.

PMT_error
The Program Association Table (PAT) tells the decoder how many programs there are in the stream and points to the PMTs which contain the information where the parts for any given event can be found. Parts in this context are the video stream (normally one) and the audio streams and the data stream (e.g. Teletext). Without a PMT the corresponding program is not decodable.

PID_error
It is checked whether there exists a data stream for each PID that occurs. This error might occur where TS are multiplexed, or demultiplexed and again remultiplexed.

The user specified period should not exceed 5 s for video or audio PIDs (see note). Data services and audio services with ISO 639 [i.17] language descriptor with type greater than '0' should be excluded from this 5 s limit.

NOTE: For PIDs carrying other information such as sub-titles, data services or audio services with ISO 639 [i.17] language descriptor with type greater than '0', the time between two consecutive packets of the same PID may be significantly longer.

In principle, a different user specified period could be defined for each PID.

5.2.2 Second priority: recommended for continuous or periodic monitoring

Table 5.0b: MPEG-2 TS parameters of 2nd priority

No.	Indicator	Precondition	Reference
2.1	Transport_error	Transport_error_indicator in the TS-Header is set to "1"	ISO/IEC 13818-1 [i.1]: clauses 2.4.3.2, 2.4.3.3
2.2	CRC_error	CRC error occurred in CAT, PAT, PMT, NIT, EIT, BAT, SDT or TOT table	ISO/IEC 13818-1 [i.1]: clauses 2.4.4, annex A ETSI EN 300 468 [i.7]: clause 5.2
2.3	PCR_error (see notes 1 and 2)	PCR discontinuity of more than 100 ms occurring without specific indication. Time interval between two consecutive PCR values more than 100 ms	ISO/IEC 13818-1 [i.1]: clauses 2.4.3.4, 2.4.3.5 ISO/IEC 13818-4 [i.2]: clause 9.11.3 ETSI TS 101 154 [i.30]: clause 4.1.5.3
2.3a	PCR_repetition_ error (see notes 1 and 2)	Time interval between two consecutive PCR values more than 100 ms	ETSI TS 101 154 [i.30]: clause 4.1.5.3
2.3b	PCR_discontinuity_ indicator_error	The difference between two consecutive PCR values ($PCR_{i+1} – PCR_i$) is outside the range of 0...100 ms without the discontinuity_indicator set	ISO/IEC 13818-1 [i.1]: clauses 2.4.3.4, 2.4.3.5 ISO/IEC 13818-4 [i.2]: clause 9.1.1.3
2.4	PCR_accuracy_ error	PCR accuracy of selected programme is not within ±500 ns	ISO/IEC 13818-1 [i.1]: clause 2.4.2.2
2.5	PTS_error (see note 3)	PTS repetition period more than 700 ms	ISO/IEC 13818-1 [i.1]: clauses 2.4.3.6, 2.4.3.7, 2.7.4

No.	Indicator	Precondition	Reference
2.6	CAT_error	Packets with transport_scrambling_control not 00 present, but no section with table_id = 0x01 (i.e. a CAT) present Section with table_id other than 0x01 (i.e. not a CAT) found on PID 0x0001	ISO/IEC 13818-1 [i.1]: clause 2.4.4
NOTE 1: The old version of PCR_error (2.3) is a combination of the more specific errors PCR_repetition_error (2.3.a) and PCR_discontinuity_indicator_error (2.3.b) by a logical 'or' function. It is kept in the present document for reasons of consistency of existing implementations. For new implementations it is recommended that the indicators 2.3.a and 2.3.b are used only. NOTE 2: The limitation to 40 ms in the 'Preconditions' of 2.3 PCR_error and 2.3a PCR_repetition_error was removed from ETSI TS 101 154 [i.30] in 2005. The respective clause there now refers only to the 100 ms limitation in [i.1] which is recommended to be applied generally. NOTE 3: The limitation to 700 ms should not be applied to still pictures.			

Transport_error
The primary Transport_error indicator is Boolean, but there should also be a resettable binary counter which counts the erroneous TS packets. This counter is intended for statistical evaluation of the errors. If an error occurs, no further error indication should be derived from the erroneous packet.

There may be value in providing a more detailed breakdown of the erroneous packets, for example, by providing a separate Transport_error counter for each program stream or by including the PID of each erroneous packet in a log of Transport_error events. Such extra analysis is regarded as optional and not part of this recommendation.

CRC_error
The CRC check for the CAT, PAT, PMT, NIT, EIT, BAT, SDT and TOT indicates whether the content of the corresponding table is corrupted. In this case no further error indication should be derived from the content of the corresponding table.

PCR_error
The PCRs are used to re-generate the local 27 MHz system clock. If the PCR do not arrive with sufficient regularity then this clock may jitter or drift. The receiver/decoder may even go out of lock. In DVB a repetition period of not more than 100 ms is permitted, previously a maximum of 40ms was recommended (see note 2 in table 5.0b).

PCR_repetition_error
The PCRs are used to re-generate the local 27 MHz system clock. If the PCR do not arrive with sufficient regularity then this clock may jitter or drift. The receiver/decoder may even go out of lock. In DVB a repetition period of not more than 100 ms is permitted, previously a maximum of 40ms was recommended (see note 2 in table 5.0b).

The error indication that may result from the check of this repetition period should be called PCR_repetition_error in future implementations (after the release of the present document).

PCR_discontinuity_indicator_error
The PCR_discontinuity_indicator_error is set in the case that a discontinuity of the PCR values occurs that has not been signalled appropriately by the discontinuity indicator. The usage of this indicator is recommended for future implementations (after the release of the present document).

PCR_accuracy_error
The accuracy of ±500 ns is intended to be sufficient for the colour subcarrier to be synthesized from system clock.

This test should only be performed on a constant bitrate TS as defined in ISO/IEC 13818-1 [i.1] clause 2.4.2.2.

Further information on PCR jitter measurements is given in clause 5.3.2.

PTS_error
The Presentation Time Stamps (PTS) should occur at least every 700 ms (see note 3 in table 5.0b). They are only accessible if the TS is not scrambled.

CAT_error
The CAT is the pointer to enable the IRD to find the EMMs associated with the CA system(s) that it uses. If the CAT is not present, the receiver is not able to receive management messages.

5.2.3 Third priority: application dependant monitoring

Table 5.0c: MPEG-2 TS parameters of 3rd priority

No.	Indicator	Precondition	Reference
3.1	NIT_error (note 2)	Section with table_id other than 0x40 or 0x41 or 0x72 (i. e. not an NIT or ST) found on PID 0x0010 No section with table_id 0x40 or 0x41 (i.e. an NIT) in PID value 0x0010 for more than 10 s.	ETSI EN 300 468 [i.7], clause 5.2.1 ETSI TR 101 211 [i.8], clauses 4.1, 4.4
3.1.a	NIT_actual_error	Section with table_id other than 0x40 or 0x41 or 0x72 (i.e. not an NIT or ST) found on PID 0x0010 No section with table_id 0x40 (i.e. an NIT_actual) in PID value 0x0010 for more than 10 s. Any two sections with table_id = 0x40 (NIT_actual) occur on PID 0x0010 within a specified value (25 ms or lower).	ETSI EN 300 468 [i.7], clause 5.2.1, 5.1.4 ETSI TR 101 211 [i.8], clauses 4.1, 4.4
3.1.b	NIT_other_error	Interval between sections with the same section_number and table_id = 0x41 (NIT_other) on PID 0x0010 longer than a specified value (10 s or higher).	ETSI TR 101 211 [i.8], clause 4.4
3.2	SI_repetition_ error	Repetition rate of SI tables outside of specified limits.	ETSI EN 300 468 [i.7], clause 5.1.4 ETSI TR 101 211 [i.8], clause 4.4

3.3	Buffer_error	**TB_buffering_error**
		overflow of transport buffer (TB_n)
		TBsys_buffering_error
		overflow of transport buffer for system information (Tb_{sys})
		MB_buffering_error
		overflow of multiplexing buffer (MB_n) or
		if the *vbv_delay method* is used:
		underflow of multiplexing buffer (Mb_n)
		EB_buffering_error
		overflow of elementary stream buffer (EB_n) or
		if the *leak method* is used:
		underflow of elementary stream buffer (EB_n)
		though low_delay_flag and DSM_trick_mode_flag are set to 0
		else (*vbv_delay method*)
		underflow of elementary stream buffer (EB_n)
		B_buffering_error
		overflow or underflow of main buffer (B_n)
		Bsys_buffering_error
		overflow of PSI input buffer (B_{sys})
			ISO/IEC 13818-1 [i.1], clause 2.4.2.3
			ISO/IEC 13818-4 [i.2], clauses 9.11.2, 9.1.4

3.4	Unreferenced_PID	PID (other than PAT, CAT, CAT_PIDs, PMT_PIDs, NIT_PID, SDT_PID, TDT_PID, EIT_PID, RST_PID, reserved_for_future_use PIDs, or PIDs user defined as private data streams) not referred to by a PMT within 0,5 s (note 1).	ETSI EN 300 468 [i.7], clause 5.1.3
3.4.a	Unreferenced_PID	PID (other than PMT_PIDs, PIDs with numbers between 0x00 and 0x1F or PIDs user defined as private data streams) not referred to by a PMT or a CAT within 0,5 s.	ETSI EN 300 468 [i.7], clause 5.1.3
3.5	SDT_error (note 3)	Sections with table_id = 0x42 (SDT, actual TS) not present on PID 0x0011 for more than 2 s Sections with table_ids other than 0x42, 0x46, 0x4A or 0x72 found on PID 0x0011.	ETSI EN 300 468 [i.7], clause 5.1.3 ETSI TR 101 211 [i.8], clauses 4.1, 4.4
3.5.a	SDT_actual_error	Sections with table_id = 0x42 (SDT, actual TS) not present on PID 0x0011 for more than 2 s Sections with table_ids other than 0x42, 0x46, 0x4A or 0x72 found on PID 0x0011. Any two sections with table_id = 0x42 (SDT_actual) occur on PID 0x0011 within a specified value (25 ms or lower).	ETSI EN 300 468 [i.7], clauses 5.2.3, 5.1.4 ETSI TR 101 211 [i.8], clauses 4.1, 4.4

No.	Indicator	Precondition	Reference
3.5.b	SDT_other_error	Interval between sections with the same section_number and table_id = 0x46 (SDT, other TS) on PID 0x0011 longer than a specified value (10s or higher).	ETSI TR 101 211 [i.8], clause 4.4
3.6	EIT_error (note 4)	Sections with table_id = 0x4E (EIT-P/F, actual TS) not present on PID 0x0012 for more than 2 s Sections with table_ids other than in the range 0x4E - 0x6F or 0x72 found on PID 0x0012.	ETSI EN 300 468 [i.7], clause 5.1.3 ETSI TR 101 211 [i.8], clauses 4.1, 4.4
3.6.a	EIT_actual_error	Section '0' with table_id = 0x4E (EIT-P, actual TS) not present on PID 0x0012 for more than 2 s Section '1' with table_id = 0x4E (EIT-F, actual TS) not present on PID 0x0012 for more than 2 s Sections with table_ids other than in the range 0x4E - 0x6F or 0x72 found on PID 0x0012. Any two sections with table_id = 0x4E (EIT-P/F, actual TS) occur on PID 0x0012 within a specified value (25 ms or lower).	ETSI EN 300 468 [i.7], clauses 5.2.4, 5.1.4 ETSI TR 101 211 [i.8], clauses 4.1, 4.4
3.6.b	EIT_other_error	Interval between sections '0' with table_id = 0x4F (EIT-P, other TS) on PID 0x0012 longer than a specified value (10 s or higher); Interval between sections '1' with table_id = 0x4F (EIT-F, other TS) on PID 0x0012 longer than a specified value (10 s or higher).	ETSI TR 101 211 [i.8], clause 4.4
3.6.c	EIT_PF_error	If either section ('0' or '1') of each EIT P/F sub table is present both should exist. Otherwise EIT_PF_error should be indicated.	ETSI EN 300 468 [i.7], clause 5.2.4.
3.7	RST_error	Sections with table_id other than 0x71 or 0x72 found on PID 0x0013. Any two sections with table_id = 0x71 (RST) occur on PID 0x0013 within a specified value (25 ms or lower).	ETSI EN 300 468 [i.7], clause 5.1.3
3.8	TDT_error	Sections with table_id = 0x70 (TDT) not present on PID 0x0014 for more than 30 s Sections with table_id other than 0x70, 0x72 (ST) or 0x73 (TOT) found on PID 0x0014. Any two sections with table_id = 0x70 (TDT) occur on PID 0x0014 within a specified value (25 ms or lower).	ETSI EN 300 468 [i.7], clauses 5.1.3, 5.2.6 ETSI TR 101 211 [i.8], clauses 4.1, 4.4
3.9	Empty_buffer_error	Transport buffer (TB_n) not empty at least once per second or transport buffer for system information (TB_{sys}) not empty at least once per second or if the *leak method* is used multiplexing buffer (MB_n) not empty at least once per second.	ISO/IEC 13818-1 [i.1], clauses 2.4.2.3, 2.4.2.6 ISO/IEC 13818-9 [i.3], annex E ISO/IEC 13818-4 [i.2], clauses 9.1.1.2, 9.1.4
3.10	Data_delay_error	Delay of data (except still picture video data) through the TSTD buffers superior to 1 second; or delay of still picture video data through the TSTD buffers superior to 60 s.	ISO/IEC 13818-1 [i.1], clauses 2.4.2.3, 2.4.2.6

No.	Indicator	Precondition	Reference
NOTE 1: It is assumed that transition states are limited to 0,5 s, and these transitions should not cause error indications.			
NOTE 2: The old version of NIT_error (3.1) has been split into the more specific errors NIT_actual_error (3.1.a) and NIT_other_error (3.1.b). The old version is kept in the document for reasons of consistency of existing implementations. For new implementations it is recommended that the indicators 3.1.a and 3.1.b are used only.			
NOTE 3: The old version of SDT_error (3.5) has been split into the more specific errors SDT_actual_error (3.5.a) and SDT_other_error (3.5.b). The old version is kept in the present document for reasons of consistency of existing implementations. For new implementations it is recommended that the indicators 3.5.a and 3.5.b are used only.			
NOTE 4: The old version of EIT_error (3.6) has been split into the more specific errors EIT_actual_error (3.6.a), EIT_other_error (3.6.b) and EIT_PF_error (3.6.c). The old version is kept in the present document for reasons of consistency of existing implementations. For new implementations it is recommended that the indicators 3.6.a, 3.6.b and 3.6.c are used only.			

NIT_error [i.1]
Network Information Tables (NITs) as defined by DVB contain information on frequency, code rates, modulation, polarization, etc. of various programs which the decoder can use. It is checked whether NITs are present in the TS and whether they have the correct PID.

NIT_actual_error
Network Information Tables (NITs) as defined by DVB contain information on frequency, code rates, modulation, polarization, etc. of various programs which the decoder can use. It is checked whether the NIT related to the respective TS is present in this TS and whether it has the correct PID.

NIT_other_error
Further Network Information Tables (NITs) can be present under a separate PID and refer to other TSs to provide more information on programmes available on other channels. Their distribution is not mandatory and the checks should only be performed if they are present.

SI_repetition_error
For SI tables a maximum and minimum periodicity are specified in ETSI EN 300 468 [i.7] and ETSI TR 101 211 [i.8]. This is checked for this indicator. This indicator should be set in addition to other indicators of repetition errors for specific tables.

Buffer_error
For this indicator a number of buffers of the MPEG-2 reference decoder are checked whether they would have an underflow or an overflow.

Unreferenced_PID
Each non-private program data stream should have its PID listed in the PMTs.

SDT_error
The SDT describes the services available to the viewer. It is split into sub-tables containing details of the contents of the current TS (mandatory) and other TS (optional). Without the SDT, the IRD is unable to give the viewer a list of what services are available. It is also possible to transmit a BAT on the same PID, which groups services into "bouquets".

SDT_actual_error
The SDT (Service Description Table) describes the services available to the viewer. It is split into sub-tables containing details of the contents of the current TS (mandatory) and other TS (optional). Without the SDT, the IRD is unable to give the viewer a list of what services are available. It is also possible to transmit a BAT on the same PID, which groups services into "bouquets".

SDT_other_error
This check is only performed if the presence of a SDT for other TSs has been established.

EIT_error
The EIT (Event Information Table) describes what is on now and next on each service, and optionally details the complete programming schedule. The EIT is divided into several sub-tables, with only the "present and following" information for the current TS being mandatory. The EIT schedule information is only accessible if the TS is not scrambled.

EIT_actual_error
The EIT (Event Information Table) describes what is on now and next on each service, and optionally details the complete programming schedule. The EIT is divided into several sub-tables, with only the "present and following" information for the current TS being mandatory. If there are no 'Present' or 'Following' events, empty EIT sections will be transmitted according to ETSI TR 101 211 [i.8]. The EIT schedule information is only accessible if the TS is not scrambled.

EIT_other_error
This check is only performed if the presence of an EIT for other TSs has been established.

RST_error
The RST is a quick updating mechanism for the status information carried in the EIT.

TDT_error
The TDT carries the current UTC time and date information. In addition to the TDT, a TOT can be transmitted which gives information about a local time offset in a given area.

The carriage of the following tables:

- NIT_other;
- SDT_other;
- EIT_P/F_other;
- EIT_schedule_other;
- EIT_schedule_actual;

is optional and therefore these tests should only be performed when the respective table is present.

When these tables are present this will be done automatically by measuring the interval rather than the occurrence of the first section.

As a further extension of the checks and measurements mentioned above an additional test concerning the SI is recommended: all mandatory descriptors in the SI tables should be present and the information in the tables should be consistent.

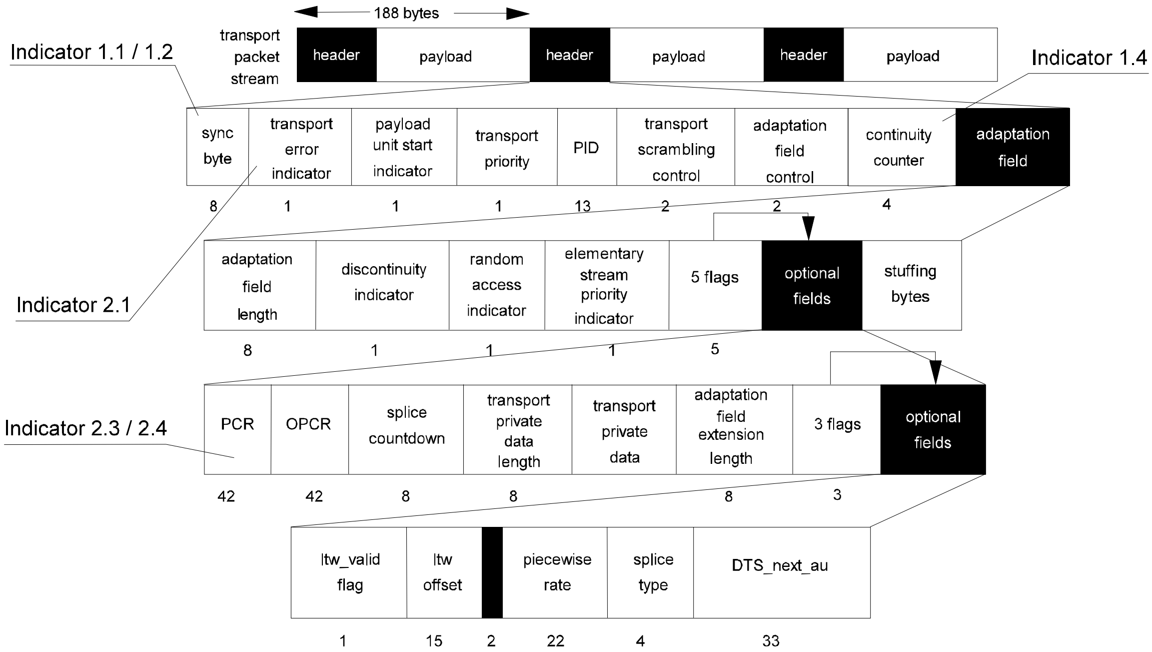


Figure 5.1: Indicators related to TS syntax

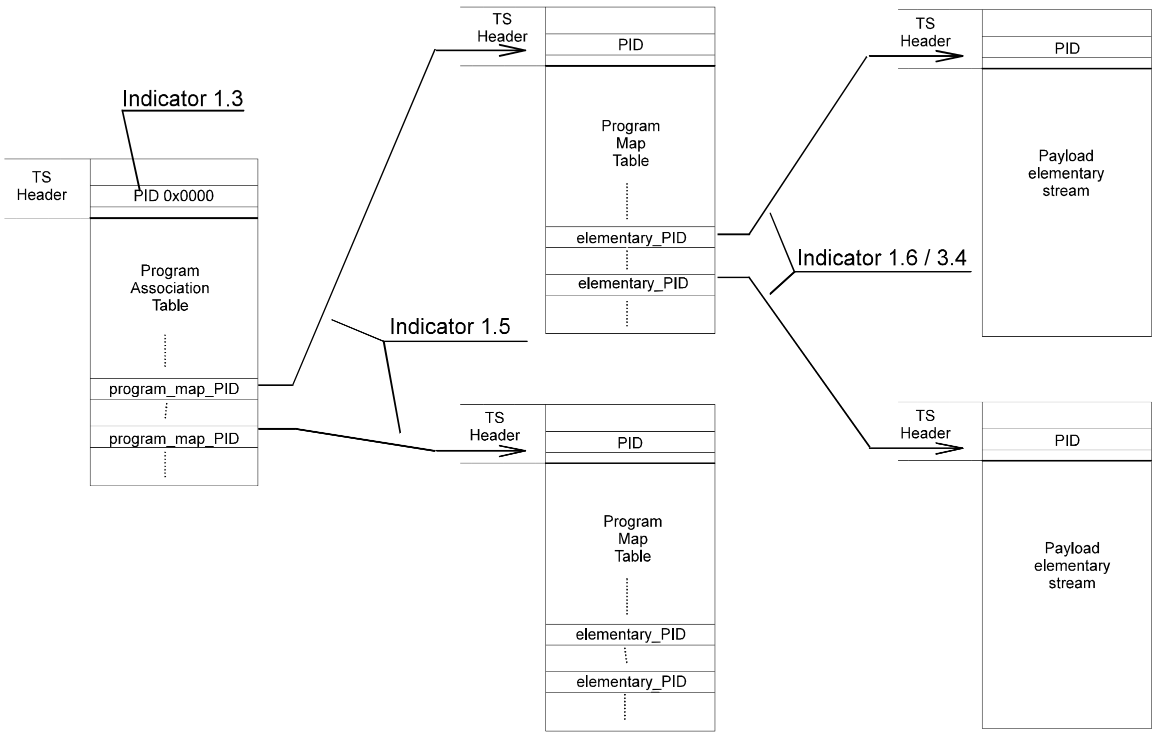


Figure 5.2: Indicators related to TS structure

5.3 Measurement of MPEG-2 Transport Streams in networks

5.3.1 Introduction

A MPEG-2 Transport Stream that is transmitted over any real network, is exposed to certain effects caused by the network components which are not ideally transparent. One of the pre-dominant effects is the acquisition of jitter in relation to the PCR values and their position in the TS. The parameters defined in clause 5.3.2 describe the various jitter components which can be differentiated by demarcation frequencies.

For the measurement of bitrates of Transport Streams, the requirements vary significantly for constant bitrate TS and partial TS/variable bitrate TS. The application of statistical multiplexers led to more dynamic variations in the bitrate, especially of the video components. Other services such as opportunistic data transmission, have typical features which again differ in terms of occurrence or presence of the service and the variation of bitrates. In clause 5.3.3 several profiles are defined to accommodate the majority of such applications, and which can be applied for monitoring and localization of failures.

5.3.2 System clock and PCR measurements

5.3.2.1 Reference model for system clock and PCR measurements

This clause presents a reference model for any source of a transport stream (TS) concerning the generation of PCR values and delivery delays. It models all the timing effects visible at the TS interface point. It is not intended to represent all the mechanisms by which these timing effects could arise in real systems.

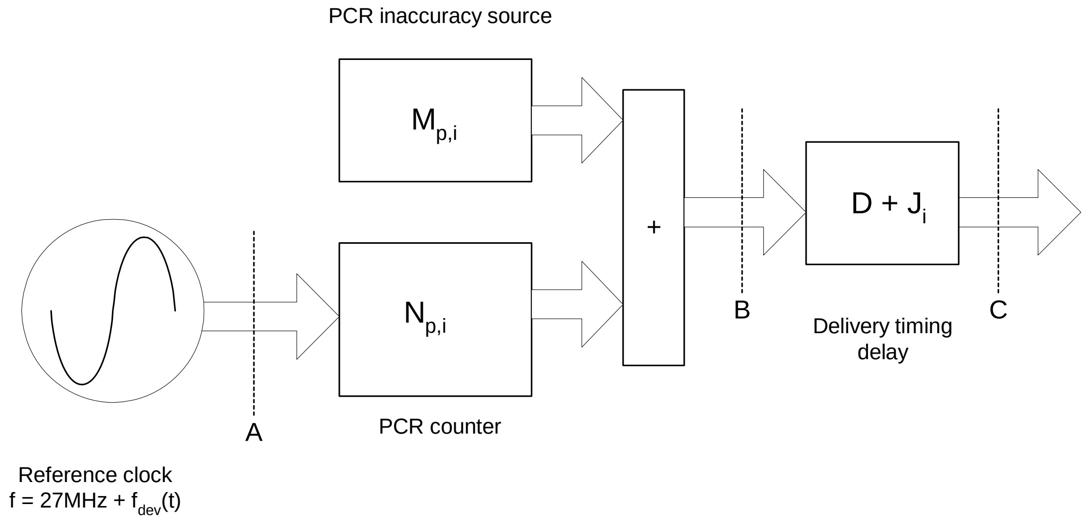


Figure 5.3: Reference model

Reference points are indicated by dashed lines. This is a model of an encoder/multiplexer (up to reference point B) and a physical delivery mechanism or communications network (between reference points B and C). The components of the model to the left of reference point B are specific to a single PCR PID. The components of the model to the right of reference point B relate to the whole Transport Stream. Measuring equipment can usually only access the TS at reference point C.

The model consists of a system clock frequency oscillator with a nominal frequency of 27 MHz, but whose actual frequency deviates from this by a function $f_{dev}(p, t)$. This function depends on the time (t) and is specific to a single PCR PID (p). The "Frequency Offset PCR_FO" measures the value of $f_{dev}(p, t)$. The "Drift Rate PCR_DR" is the rate of change with time of $f_{dev}(p, t)$.

The system clock frequency oscillator drives a PCR counter which generates an idealized PCR count, $N_{p,i}$. p refers to the specific PCR PID p and i refers to the bit position in the transport stream. To this is added a value from a PCR inaccuracy source, $M_{p,i}$ to create the PCR value seen in the stream, $P_{p,i}$. The simple relationship between these values is:

$$P_{p,i} = N_{p,i} + M_{p,i} \tag{1}$$

$M_{p,i}$ represents the "Accuracy PCR_AC".

The physical delivery mechanism or communications network beyond point B introduces a variable delay between the departure time T_i and the arrival time U_i of bits:

$$U_i - T_i = D + J_i \tag{2}$$

In the case of a PCR, U_i is the time of arrival of the last bit of the last byte containing the PCR base (ISO/IEC 13818-1 [i.1], clause 2.4.3.5). D is a constant representing the mean delay through the communications network. J_i represents the jitter in the network delay and its mean value over all time is defined to be zero. $J_i + M_{p,i}$ is measured as the "Overall Jitter PCR_OJ".

In the common case where the Transport Stream is constant bitrate, at reference point B the Transport Stream is being transmitted at a constant bitrate R_{nom}. It is important to note that in this reference model this bitrate is accurate and constant; there is no error contribution from varying bitrate. This gives an additional equation for the departure time of packets:

$$T_i = T_0 + \frac{i}{R_{nom}} \tag{3}$$

T_0 is a constant representing the time of departure of the zero'th bit. Combining equations 2 and 3 the arrival time is:

$$U_i = T_0 + \frac{i}{R_{nom}} + D + J_i \tag{4}$$

5.3.2.2 Measurement descriptions

The following measurements require a demarcation frequency for delimiting the range of drift rate and jitter frequencies of the timing variations of PCRs and/or TSs.

The demarcation frequency used should be chosen from table 5.1 and indicated with the measurement results.

Table 5.1: Profiles for jitter and drift rate measurements

Profile	Demarcation frequency	Comments
MGF1	10 mHz	This profile is provided to give the total coverage of frequency components included in the timing impairments of PCR related measurements. This profile provides the most accurate results in accordance with the limits specified in ISO/IEC 13818-1 [i.1], clause 2.4.2.1. If jitter or drift rate measurements are found out of specification when using other profiles, it is suggested to use this one for better accuracy.
MGF2	100 mHz	This profile is accounting for intermediate benefits between the profiles MGF1 and MGF3, by giving reasonable measurement response as well as reasonable account for low frequency components of the timing impairments.
MGF3	1 Hz	This profile provides faster measurement response by taking in account only the highest frequency components of the timing impairments. This profile is expected to be sufficient in many applications.

Profile	Demarcation frequency	Comments
MGF4	Manufacturer defined	This profile will provide any benefit that the manufacturer may consider as useful when it is designed and implemented in a measurement instrument. The demarcation frequency has to be supplied with the measurement result. Optionally any other data that the manufacturer may consider to be relevant may be supplied. For testing against ISO/IEC 13818-9 [i.3] (±25 µs jitter limit) a demarcation frequency of 2 mHz is required. A filter for such demarcation may be implemented under this MGF4 profile.

5.3.2.3 Program Clock Reference - Frequency Offset PCR_FO

Definition	PCR_FO is defined as the difference between the program clock frequency and the nominal clock frequency (measured against a reference which is not PCR derived, neither TS derived). The units for the parameter PCR_FO should be in Hz according to: • Measured Frequency - Nominal Frequency, • or in ppm expressed as: – [Measured Frequency (in Hz) - Nominal Frequency(in Hz)]/Nominal Frequency (in MHz).
Purpose	The original frequency of the clock used in the digital video format before compression (program clock) is transmitted to the final receiver in form of numerical values in the PCR fields. The tolerance as specified by ISO/IEC 13818-1 [i.1] clause 2.4.2.1 is ±810 Hz or ±30 ppm.
Interface	A, Z

5.3.2.4 Program Clock Reference – Drift Rate PCR_DR

Definition	PCR_DR is defined as the first derivative of the frequency and is measured on the low frequency components of the difference between the program clock frequency and the nominal clock frequency (measured against a reference which is not PCR derived, neither TS derived). The format of the parameter PCR_DR should be in mHz/s (@ 27 MHz) or ppm/hour.
Purpose	The measurement is designed to verify that the frequency drift, if any, of the program clock frequency is below the limits set by ISO/IEC 13818-1 [i.1]. This limit is effective only for the low frequency components of the variations. The tolerance as specified by ISO/IEC 13818-1 [i.1] is ±75 mHz/s@ 27 MHz or ±10 ppm/hour.
Interface	A, Z
NOTE: A break frequency of 10 mHz is recommended for the separation of PCR_jitter (higher frequencies) and PCR_drift (lower frequencies). See also MGF1 in Table 5.1.	

5.3.2.5 Program Clock Reference - Overall Jitter PCR_OJ

Definition	PCR_OJ is defined as the instantaneous measurement of the high frequency components of the difference between when a PCR should have arrived at a measurement point (based upon previous PCR values, its own value and a reference which is not PCR or TS derived) and when it did arrive. The format of the parameter PCR_OJ should be in nanoseconds.
Purpose	The PCR_OJ measurement is designed to account for all cumulative errors affecting the PCR values during program stream generation, multiplexing, transmission, etc. All these effects appear as jitter at the receiver but they are a combination of PCR inaccuracies and jitter in the transmission. This value can be compared against the maximum error specification by ISO/IEC 13818-1 [i.1] for PCR Accuracy of ±500 ns only if the jitter in the transmission is assumed to be zero.
Interface	A, Z
NOTE: A break frequency of 10 mHz is recommended for the separation of PCR_jitter (higher frequencies) and PCR_drift (lower frequencies). See also MGF1 in table 5.1.	

5.3.2.6 Program Clock Reference – Accuracy PCR_AC

Definition	The accuracy of the PCR values PCR_AC is defined as the difference between the actual PCR value and the value it should have in the TS represented by the byte index for its actual position. This can be calculated for constant bitrate TS, the measurement may NOT produce meaningful results in variable bitrate TS. The units for the parameter PCR_AC should be in nanoseconds.
Purpose	This measurement is designed to indicate the total error included in the PCR value with respect to its position in the TS. The tolerance as specified by ISO/IEC 13818-1 [i.1] is ±500 ns. This measurement is considered to be valid for both: real time and off-line measurements. The measurement should trigger the indicator under clause 5.2.2, item 2.4.
Interface	A. Z

NOTE: Note that PCR Accuracy is defined by ISO/IEC 13818-1 [i.1]: "A tolerance is specified for the PCR values. The PCR tolerance is defined as the maximum inaccuracy allowed received PCRs. This inaccuracy may be due to imprecision in the PCR values or to PCR modification during re-multiplexing. It does not include errors in packet arrival time due to network jitter or other causes".

5.3.3 Bitrate measurement

5.3.3.0 General

The bitrate value from a measurement system depends on a number of parameters:

- when the bitrate measurement is started;
- what is counted (packets, bytes, bits);
- the time duration (gate) over which the bitrate is measured;
- the way in which the time-gate function moves between measurements (timeSlice).

5.3.3.1 Bitrate measurement algorithm

This clause defines the parameter "**MG bitrate**" which is an instantaneous bitrate value. The bitrate is averaged over a fixed time gate (or "window"). This gating function is moved by a discrete time slice (or interval) to produce the bitrate value for each time slice. (The window "hops" from one time slice to the next) The items that are counted can be bits, bytes or Transport Stream packets, and the meaning of the measured value should be made clear by accurate labelling (see Nomenclature below). The measurement can be applied to the entire Transport Stream or a partial transport stream obtained by applying a PID filter or even a filter to remove packet headers.

The following equation defines "MG bitrate":

$$MG_bitrate_at_timeSlice_t = \frac{elementSize}{\mathrm{T}} \times \sum_{n=0}^{n=N\tau-1} num_elements_in_timeSlice_{t-n\tau}$$

Where:

N is the integer number of time slices during the time gate.

T = Nτ is the duration of the time gate in seconds.

τ is the width of each time slice in seconds.

element is the fundamental unit which is being counted by the bitrate measurement algorithm.

elementSize is the size (measured in the appropriate units) of the element being measured. For example if bitrate units are packets/s then the elementSize should be expressed in packets. If bitrate units are bits/s then the elementSize is expressed in bits. Hence if an element is a 188 byte packets then elementSize can be expressed as:

elementSize = 188 bytes/packet × 8 bits/byte = 1 504 bits

num_elements_in_timeSlice

is the integer number of element starts which have occurred in the timeSlice. If an element is a 188 byte packet then this corresponds to counting sync bytes. If an element is a byte then this may correspond to counting the first bit in transmission order on a serial link.

The units of $MG_bitrate_at_timeSlice_t$ are not part of the present document, but should be the same as the units used to express elementSize. This is because the bitrate can be expressed in a number of different ways as is described in the Nomenclature clause below.

The measurement is discrete. A new measurement value is available every timeSlice and is held for the duration of a timeSlice. Display of a bitrate value in a piece of measurement equipment may not be a precise display of this value as is indicated in figure 5.4.

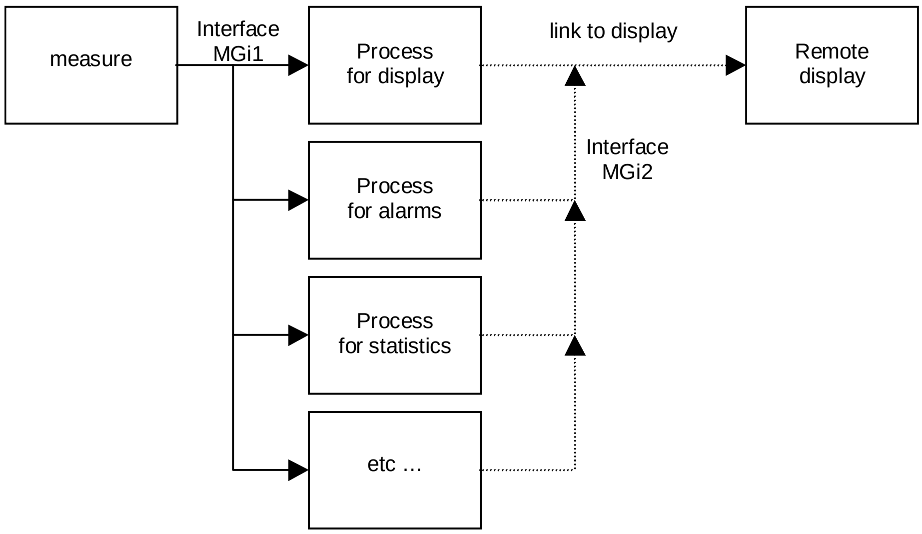


Figure 5.4: Display of a bitrate value

5.3.3.2 Preferred values for Bitrate Measurement

The preferred values for the algorithm are application dependent. One set of values may be appropriate for monitoring and another may be appropriate for precise measurements. In order to have consistent measurements between different equipment vendors, the following profiles are defined. (Note that the timeSlice interval τ can be expressed as a time or as a frequency for precision).

Table 5.1a: Preferred profiles for bitrate measurements

MG Profile	Profile Description	Stream Type/Rate	τ	N	T=Nτ	element
MGB1	This Profile is best geared towards applications where the bitrate is constant or slowly varying. It is compatible with much equipment developed before the present document was created.	All	1 s	1	1 s	188 byte packet
MGB2	This Profile provides overall consistent rate calculations while providing reasonable accuracy for most monitoring and troubleshooting applications. It is intended for CBR measurements whereas rapidly varying bitrates are more appropriately measured with the MGB3 or MGB4 profiles.	All	100 ms	10	1 s	188 byte packet
MGB3	This Profile provides for tracking of small variations in the multiplex rate of each element.	All	1/90 kHz	1 800	20 ms	188 byte packet
MGB4	This Profile provides for a longer term average for rate calculation but with repeatability between two different measurements of the same data.	All	1/90 kHz	9×10^4	1 s	188 byte packet
MGB5	This Profile allows the user to tune bitrate calculations based on the parameters that are most appropriate for a particular transport stream. It is very important that when this is done, the nomenclature used to define the bitrate clearly shows that bitrates for components are not directly comparable with each other: TS@MGB1 video@MGB3 audio@MGB4 the_rest@188,1s,100s etc. This follows the nomenclature guide in the present document and shows that it is unlikely that the sum of the bitrates of the TS components will equal the overall transport stream rate.	Complete or partial transport stream	User Def.	User Def.	User Def.	188 byte packet

Applications of the profiles are given in the informative annex J.

5.3.3.3 Nomenclature

It is important to display bitrate values in a way which allows comparison. Correct nomenclature can indicate for example that correction factors need to be applied to convert from a 204 byte packet bitrate measurement to a 188 byte packet measurement. This recommendation is for the "MG-bitrate" nomenclature. If the "MG bitrate" algorithm has been used, then bitrates are of the form:

<bitrate_value> <units>@ MGprofile

or <bitrate_value> <units> @ MG<element>, <timeslice>, <time_gate> [,<filter>]

For example if the full transport stream bitrate of a 204 byte packet system is to be measured, then it is important to know the size of the packet (i.e. the elementSize) and the size of the time window which was measured to ensure repeatability. Hence a bitrate should be expressed as:

EXAMPLE 1: 10,300 Mbit/s @ MG 204,1/90 kHz, 1,1 s *1*

It is assumed by default that the bitrate was for the full transport stream.

If the bitrate of all the service components for a service called "Test Transmission" (i.e. all PIDs listed in the PMT + the bitrate of the PMT excluding the bitrate of EITp and EITf for that service) is to be measured, then it would be expressed as:

EXAMPLE 2: 4,154 Mbit/s @ MG 188, 1/90 kHz, 1 s, service: Test Transmission

or

4,154 Mbit/s @ MGB4, service: Test Transmission

To express example 2 as a percentage of the total bitrate in example 1, it is obvious now that a 188/204 correction factor needs to be applied before the division takes place:

Test Transmission = 100 × (4,154 × 204/188)/10,300 % of bitrate

43,8 % of bitrate

Note that this nomenclature is independent of the measurement technique, but is vital to allow results to be compared.

Note also that when writing MG-bitrate measurements, the values kbit/s and Mbit/s are taken to mean 10^3 bits per second and 10^6 bits per second respectively. It is also recommended that the values kB/s (10^3 bytes/s) and MB/s (10^6 bytes/s) are not used.

5.3.4 Consistency of information check

5.3.4.0 General

The information provided in the various SI/PSI tables in different Transport Streams needs to be consistent and coherent to provide access to all services for the user. Wherever these tables are created, modified or extracted, there is a need for checking the tables of the outgoing Transport Stream.

In many cases, these applications are user-defined in the sense that providers and operators may wish to minimize the complexity of these checks.

As a first example for such a check, the Transport_Stream_ID check is defined hereafter.

5.3.4.1 Transport_Stream_ID check

Definition	Each MPEG-2 Transport Stream should be identifiable by its Transport_Stream_ID carried in the PAT.
Purpose	As DVB networks become more and more complex, there is an increased risk of transmitting the wrong Transport Stream. Providers and operators may wish to make sure that the TS they actually process is the intended one.
Interface	A, Z
Method	The Transport Stream ID (as referenced in the PAT) should be checked and the actual TS ID should be compared with a user defined value. By this it can be tested whether the actual Transport Stream is the correct one.

5.3.5 TS parameters in transmission systems with reduced SI data

Certain transmission systems, e.g. DSNG Transport Streams conforming to ETSI EN 301 210 [i.18] contain simplified PSI/SI information (see annex D of ETSI EN 301 210 [i.18]). When testing such Transport Streams, table 5.1b indicates which of the tests recommended in clause 5.2 can be used.

Table 5.1b: TS parameters in systems with reduced SI data

No.	Indicator	Comment
1.1	TS_sync_loss	Essential for access to TS data
1.2	Sync_byte_error	May not necessarily prevent decoding of content
1.3	PAT_error	Essential for access to TS data
1.3.a	PAT_error_2	Essential for access to TS data
1.4	Continuity_count_error	May not necessarily prevent decoding of content
1.5	PMT_error	Essential for access to TS data
1.5.a	PMT_error_2	Essential for access to TS data
1.6	PID_error	May not necessarily prevent decoding of content
2.1	Transport_error	
2.2	CRC_error	Applies to PAT and PMT only
2.3	PCR_error	
2.3a	PCR_repetition_error	
2.3b	PCR_discontinuity_indicator_error	
2.4	PCR_accuracy_error	
2.5	PTS_error	
2.6	CAT_error	
3.3	Buffer_error	
3.4	Unreferenced_PID	
3.4.a	Unreferenced_PID	
3.9	Empty_buffer_error	
3.10	Data_delay_error	

5.4 Measurement of availability at MPEG-2 Transport Stream level

Definitions of error events

The following definitions are used to establish criteria for System Availability, Link Availability, and System Error Performance (e.g. for coverage measurement purposes) for distribution networks such as satellite (DVB-S and DVB-DSNG), cable (DVB-C), terrestrial (DVB-T) and microwave systems (DVB-MS, DVB-MC and DVB-MT) as well as for contribution networks (DVB-PDH ETSI ETS 300 813 [i.19] and DVB-SDH ETSI ETS 300 814 [i.20]).

These definitions may also be used to test the performance of TSs in IRDs via Common Interfaces.

Table 5.2: Error Events

5.4.1	**Severely Disturbed Period (SDP):**	A period of sync loss (as defined in clause 5.2.1 of the present document, parameter 1.1) or loss of signal.
5.4.2	**Errored Block (EB):**	An MPEG-2 TS packet with one or more uncorrectable errors, which is indicated by the transport_error_indicator flag set. See clause 5.2.2.
5.4.3	**Errored Time Interval (ETI):**	A given time interval with one or more EBs.
5.4.3.a	**Errored Second (ES):**	A specific case of the ETI where the given time interval is one second.
5.4.4	**Severely Errored Time Interval (SETI):**	A given time interval which contains greater than a specified percentage of errored blocks, or at least one SDP or part thereof. This percentage will not be specified in the present document, but should be the subject of agreements between the network operators and the program providers.
5.4.4a	**Severely Errored Second (SES):**	A specific case of the SETI where the given time interval is one second.

5.4.5	**Unavailable Time UAT**	A start of a period of Unavailable Time can be defined as: • either the onset of N consecutive SES/SETI events; or • the onset of a rolling window of length T in which M SES/SETI events occur. These time intervals/seconds are considered to be part of the Unavailable Time. A end of period of Unavailable Time can be defined accordingly as • the onset of N consecutive non-SES/SETI events; or • the onset of a rolling window of length T in which no SES/SETI events occur. These time intervals/seconds are considered to be part of Available Time. The values N, M and T could differ for different types of service (video, audio, data, etc.).

Note that these tests are only possible if Reed-Solomon encoding was used upstream with respect to the measurement point.

5.5 Evaluation of service performance by combination of TS related parameters

Void.

5.6 Parameters for CI related applications

5.6.0 Introduction

The Common Interface (CI) is - in principle - a Transport Stream interface but it has particular properties which require additional tests.

The parameters defined in this clause are intended to enable reproducible and comparable measurements on the CI. As in the previous clauses on Transport Stream related tests and measurements, it cannot be assumed that these tests provide a complete analysis. They are also designed as a 'health check', not as an overall compliance or conformance test.

The following reference model pictures the interfaces and the functional blocks which are referred to in the definitions of the tests.

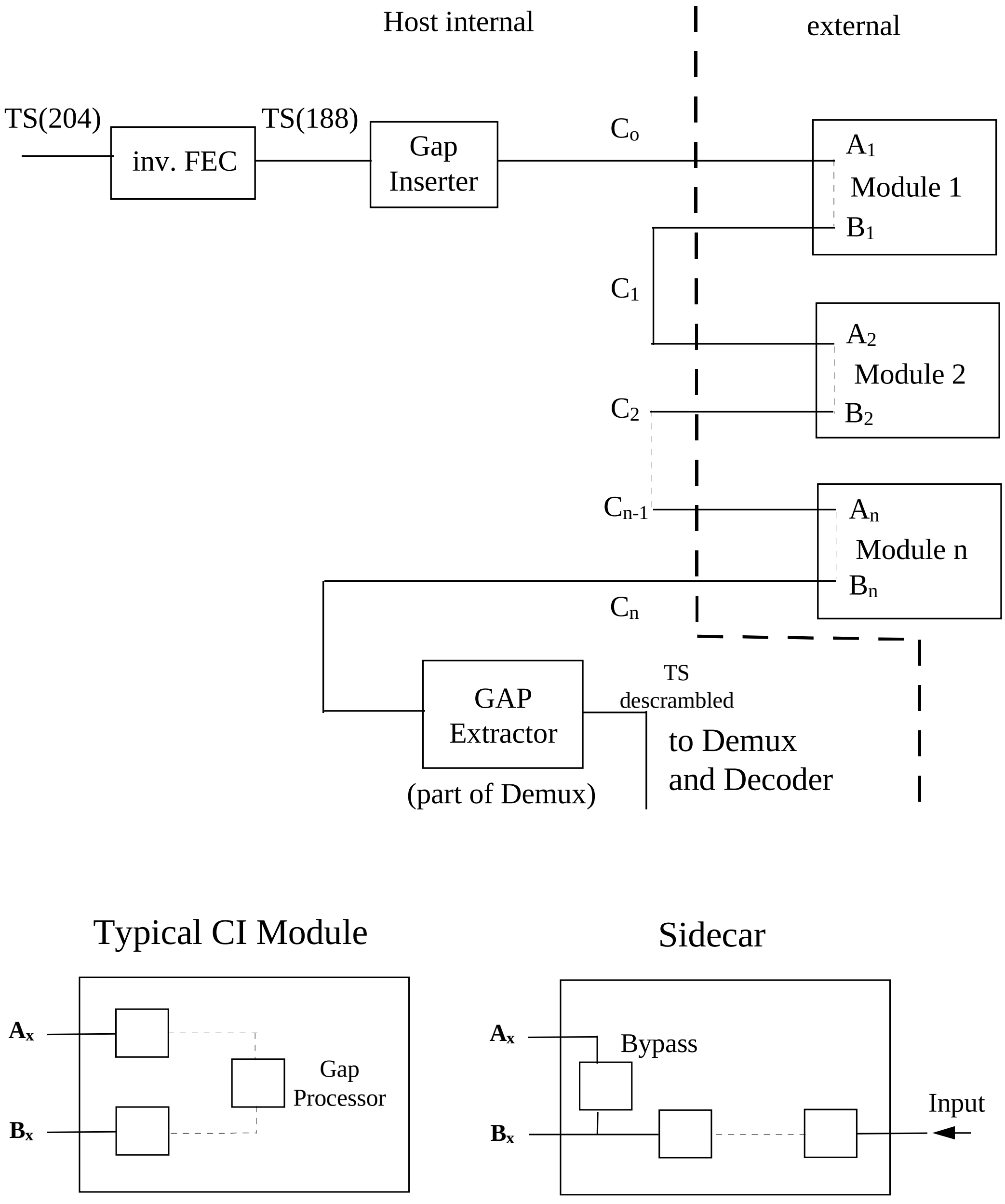


Figure 5.5: CI Reference model

5.6.1 Latency

Table 5.3

Parameter	Purpose	Interface	Method
Latency	To determine the impact of one CI module on latency (or average delay)	A_n - B_n	Measure arrival time of synch bytes of corresponding TS packets at both interfaces

5.6.2 CI_module_delay_variation

Table 5.4

Parameter	Purpose	Interface	Method	Reference
CI_module_delay_ variation	To check compliance with CI spec, to limit additional PCR jitter and support decodability	A_x - B_x	Measure delay for all corresponding bytes of each TS packet between input A_x and output B_x and calculate peak delay variation for each TS packet	EN 50221 [i.23], clause 5.4.2
NOTE: A_x and B_x are the input and output of any one CI Module.				

5.6.3 Input_output_TS comparison

Table 5.5

Parameter	Purpose	Interface	Method
Input-output TS comparison	To ensure that modules under test do not impair other parts of the TS	C_o - C_n	TS with at least 1 PID unaffected by the CI modules + other PIDs which will activate each module under test and carry out a bitwise comparison for the unaffected PIDs; additionally the CI modules should be tested while inactive

5.6.4 CI_module_throughput

Table 5.6

Parameter	Purpose	Interface	Method	Limits
Period between consecutive synch bytes	To ensure compliance with CI spec	A_x, B_x or C_x	Measure time between 2 synch bytes after processing in modules @ A_x: modules able to accept input TS @ B_x: module outputs TS within limits	58 Mbit/s from EN 50221 [i.23]
NOTE: A_x and B_x are the input and output of any one CI Module, C_x is any corresponding interface of the host device.				

5.6.5 Valid TS on CI

Table 5.7

Parameter	Purpose	Interface	Method
Valid TS	To ensure decodability	A_x, B_x or C_x	Checks as in Table 5.0a (1st priority) and 2.6 in Table 5.0b
NOTE: A_x and B_x are the input and output of any one CI Module, C_x is any corresponding interface of the host device.			

6 Common parameters for satellite and cable transmission media

6.1 System availability

Purpose	The system availability describes the long-term quality of the complete digital transmission system from MPEG-2 encoder to the measurement point.
Interface	Z
Method	The definition of System Availability is based on the list of performance parameters of table 5.2: Severely Disturbed Period (SDP) Errored Block (EB) Errored Time Interval ETI/Errored Second (ES) Severely Errored Time Interval SETI/Severely Errored Second (SES) Unavailable Time UAT The System Availability is defined as the ratio of (Total Time - Unavailable Time) to Total Time.

6.2 Link availability

Purpose	The link availability describes the long term quality of a specified link in a digital transmission chain. It could be used as a quality of service parameter in contracts between network operators and program providers.	
Interface	X (Overload indicator of the Reed Solomon decoder).	
Method	The definition of Link availability is based on following performance parameters:	
	Uncorrectable Packet (UP)	An MPEG-2 TS packet with an uncorrectable error, which is indicated by overload at the Reed-Solomon decoder.
	Uncorrectable Time Interval UTI/Uncorrectable Second (US)	A given time interval with one or more UPs. The US is a specific case of the UTI where the given time interval is one second.
	Severely Uncorrectable Time Interval (SUTI)/Severely Uncorrectable Second (SUS):	A given time interval which contains greater than a specified percentage of Uncorrectable Packets, or at least one SDP (see clause 5.4) or part thereof. NOTE: This percentage will not be specified in the present document, but should be the subject of agreements between the network operators and the service providers. The SUS is a specific case of the SUTI where the given time interval is one second.
	Link Unavailable Time LUAT	A start of a period of Link Unavailable Time can be defined as: • either the onset of N consecutive SUS/SUTI events; or • the onset of a rolling window of length T in which M SUS/SUTI events occur. These time intervals/seconds are considered to be part of the Link Unavailable Time. A end of period of Link Unavailable Time can be defined accordingly as: • the onset of N consecutive non-SUS/SUTI events; or • the onset of a rolling window of length T in which no SUS/SUTI events occur. These time intervals/seconds are considered to be part of Link Available Time. The values N, M and T could differ for different types of service (video, audio, data, etc.).

6.3 BER before RS decoder

6.3.0 Introduction

Purpose	The Bit Error Rate (BER) is the primary parameter which describes the quality of the digital transmission link.
Interface	W
Method	The BER is defined as the ratio between erroneous bits and the total number of transmitted bits. Two alternative methods are available; one for "Out of Service" and a second for "In Service" use. In both cases, the measurement should only be done within the "link available time" as defined in clause 6.2.

6.3.1 Out of service

The basic principle of this measurement is to generate within the channel encoder a known, fixed, repeating sequence of bits, essentially of a pseudo random nature. In order to do this the data entering the sync-inversion/randomization function is a continuous repetition of one fixed TS packet. This sequence is defined as the *null TS packet* in ISO/IEC 13818-1 [i.1] with all data bytes set to 0x00. i.e. the fixed packet is defined as the four byte sequence 0x47, 0x1F, 0xFF, 0x10, followed by 184 zero bytes (0 x 00). Ideally this would be available as an encoding system option (see clause A.2).

6.3.2 In service

The basic assumption made in this measurement method is that the RS check bytes are computed for each link in the transmission chain. Under normal operational circumstances, the RS decoder will correct all errors and produce an error-free TS packet. If there are severe error-bursts, the RS decoding algorithm may be overloaded, and be unable to correct the packet. In this case the transport_error_indicator bit should be set, no other bits in the packet should be changed, and the 16 RS check bytes should be recalculated accordingly before re-transmission on to another link. The BER measured at any point in the transmission chain is then the BER for that particular link only.

The number of erroneous bits within a TS packet will be estimated by comparing the bit pattern of this TS packet before and after RS decoding. If the measured value of BER exceeds 10^{-3} then the measurement should be regarded as unreliable due to the limits of the RS decoding algorithm. Any TS packet that the RS decoder is unable to correct should cause the calculation to be restarted.

6.4 Error events logging

Purpose	Error events logging creates a permanent error log which can subsequently be used to locate possible sources of errors. It may be used as a measure of "system availability" (see clause 6.1 above).
Interface	Z
Method	Loss of sync, loss of signal, and reception of errored TS packets are logged. In case of sync or signal loss, the absolute time of loss should be recorded, along with either the duration of loss or the time of recovery from loss. A default time resolution of 1 second is strongly recommended for this measurement, but other time intervals may be appropriate depending on the application. In case of reception of EBs (see clause 6.1), the number of such events in each second should be logged, together with the PID and the total number of received packets of this PID within the resolution time. Logging of any other parameters (e.g. overloading of Reed-Solomon decoder, original_network_id, service_id) are optional. The error log should store the most recent 1 000 error events as a minimum. Provision should be made to access all of the error information in a form suitable for further data processing.

6.5 Transmitter symbol clock jitter and accuracy

Purpose	Inaccuracies of the symbol clock concerning absolute frequency, frequency drift and jitter may introduce intersymbol interference. Additionally, the accuracy of transmitted clock references like the Program Clock Reference (PCR) can be influenced. Therefore the degradation of signal quality due to symbol clock inaccuracies has to be negligible. Symbol clock jitter and accuracy can be degraded if the symbol clock is directly synthesized from an unstable TS data clock. For this reason, the measurement should be performed while the transmitter is driven by a TS to ensure a worst case measurement is obtained.
Interface	E
Method	For measurements the absolute frequency, frequency wander and timing jitter are of interest. A PLL circuit can be used for synchronization to the symbol clock and according to the loop bandwidth, timing jitter is suppressed and low frequency drift (wander) is still present at the output of the loop oscillator. Jitter can be measured with an oscilloscope by triggering with the extracted clock. Jitter is usually expressed as a peak-to-peak value in UI (Unit Interval) where one UI is equal to one clock cycle (T_{symbol}). For measurements of the absolute frequency and frequency wander the output of the clock extractor can be used or the symbol clock can be measured directly using an appropriate frequency counter. NOTE: This measurement refers to the physical layer of TS interconnection. See clause 5.3.2 for PCR measurements.

6.6 RF/IF signal power

Purpose	Level measurement is needed to set up a network.
Interface	Any RF/IF interface, N, P.
Method	The signal power, or wanted power, is defined as the mean power of the selected signal as would be measured with a thermal power sensor. Care should be taken to limit the measurement to the bandwidth of the wanted signal. When using a spectrum analyser or a calibrated receiver, it should integrate the signal power within the nominal bandwidth of the signal (symbol rate x$(1 + \alpha)$).

6.7 Noise power

Purpose	Noise is a significant impairment in a transmission network.
Interface	N (out of service) or T (in service)
Method	The noise power (mean power), or unwanted power, is measured with a spectrum analyser (out of service) or an estimate is obtained from the IQ diagram (in service), see clause 6.9.9. The noise level is specified using either the occupied bandwidth of the signal, which is equal to the symbol rate x $(1 + \alpha)$. See annex G.

6.8 Bit error count after RS

Purpose	To measure whether the MPEG-2 TS is quasi error free.
Interface	Z
Method	The same principle as used for the "Out of service measurement" of the "BER before the Reed-Solomon decoder" described in clause 6.3.2, with the modification that the result is presented as an error count rather than a ratio. The receiver only has to compare the received TS packets with the Null packets as defined in clause A.2.

6.9 IQ signal analysis

6.9.1 Introduction

Assuming:

- a constellation diagram of M symbol points; and
- a measurement sample of N data points, where N is sufficiently larger than M to deliver the wanted measurement accuracy; and
- the co-ordinates of each received data point j being $I_j + \delta I_j$, $Q_j + \delta Q_j$ where I and Q are the co-ordinates of the ideal symbol point and δI and δQ are the offsets forming the error vector of the data point (see clause A.3).

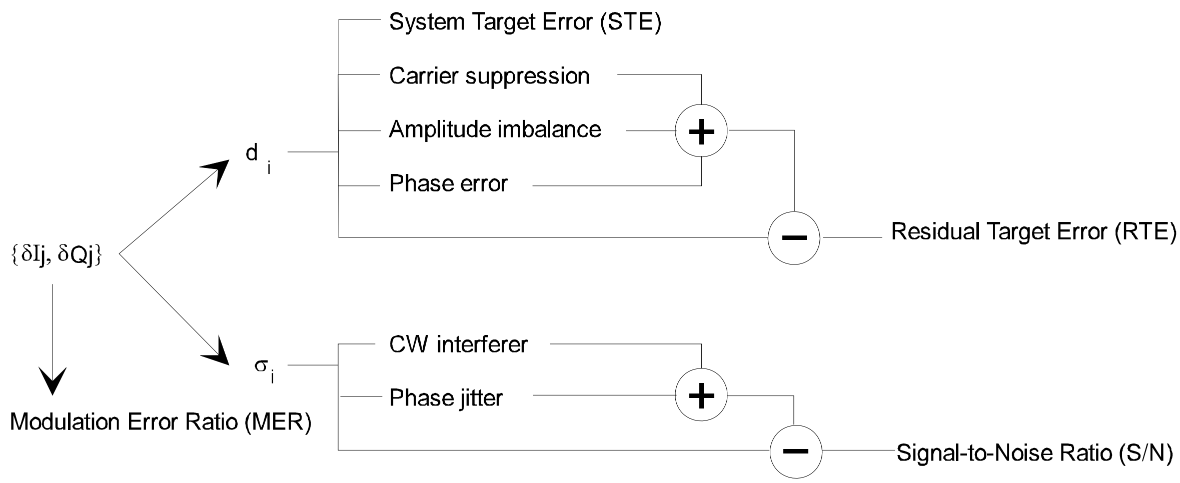


Figure 6.1: Relationship between the parameters describing different IQ distortions

Modulation Error Ratio (MER) and the related Error Vector Magnitude (EVM) are calculated from all N data points without special pre-calculation for the data belonging to the M symbol points.

With the aim of separating individual influences from the received data, for each point i of the M symbol points the mean distance d_i and the distribution σ_i can be calculated from those δI_j, δQ_j belonging to the point i.

From the M values $\{d_1, d_2, \ldots d_M\}$ the influences/parameters:

- origin offset;
- amplitude Imbalance (AI); and
- quadrature Error (QE),

can be extracted and removed from the d_i values, allowing to calculate the Residual Target Error (RTE) with the same algorithm as the System Target Error (STE) from $\{d_1, d_2, \ldots d_M\}$.

From the statistical distribution of the M clouds (denoted by σ_i in figure 6.2) parameters:

- phase jitter; and
- CW interferer,

may be extracted. The remaining clouds (after elimination of the above two influences) are assumed to be due to Gaussian noise only and are the basis for calculation of the signal-to-noise ratio. The parameter may include - besides noise - also some other disturbing effects, like small non-coherent interferers or residual errors from the equalizer. From the SNR value the Carrier/Noise value can be estimated (see clause A.3).

When using the interfaces E or G filtering of the signal before the interface should be considered.

6.9.2 Modulation Error Ratio (MER)

Purpose	To provide a single "figure of merit" analysis of the received signal. This figure is computed to include the total signal degradation likely to be present at the input of a commercial receiver's decision circuits and so give an indication of the ability of that receiver to correctly decode the signal.
Interface	E, G, S, T
Method	The carrier frequency and symbol timing are recovered, which removes frequency error and phase rotation. Origin offset (e.g. cause by residual carrier or DC offset), quadrature error and amplitude imbalance are not corrected. A time record of N received symbol co-ordinate pairs $(\tilde{I}_j, \tilde{Q}_j)$ is captured. For each received symbol, a decision is made as to which symbol was transmitted. The ideal position of the chosen symbol (the centre of the decision box) is represented by the vector (I_j, Q_j). The error vector $(\delta I_j, \delta Q_j)$ is defined as the distance from this ideal position to the actual position of the received symbol. In other words, the received vector $(\tilde{I}_j, \tilde{Q}_j)$ is the sum of the ideal vector (I_j, Q_j) and the error vector $(\delta I_j, \delta Q_j)$. The sum of the squares of the magnitudes of the ideal symbol vectors is divided by the sum of the squares of the magnitudes of the symbol error vectors. The result, expressed as a power ratio in dB, is defined as the Modulation Error Ratio (MER). $$MER = 10 \times \log_{10} \left\{ \frac{\sum_{j=1}^{N} \left(I_j^2 + Q_j^2 \right)}{\sum_{j=1}^{N} \left(\delta I_j^2 + \delta Q_j^2 \right)} \right\} dB$$ The definition of MER does not assume the use of an equalizer, however the measuring receiver may include a commercial quality equalizer to give more representative results when the signal at the measurement point has linear impairments. When an MER figure is quoted it should be stated whether an equalizer has been used. It should be reconsidered that MER is just one way of computing a "figure of merit" for a vector modulated signal. Another "figure of merit" calculation is Error Vector Magnitude (EVM) defined in annex C. It is also shown in annex C that MER and EVM are closely related and that one can generally be computed from the other. MER is the preferred first choice for various reasons itemized in annex C.

6.9.3 System Target Error (STE) (void)

Figure 6.2: Definition of Target Error Vector (TEV) (void)

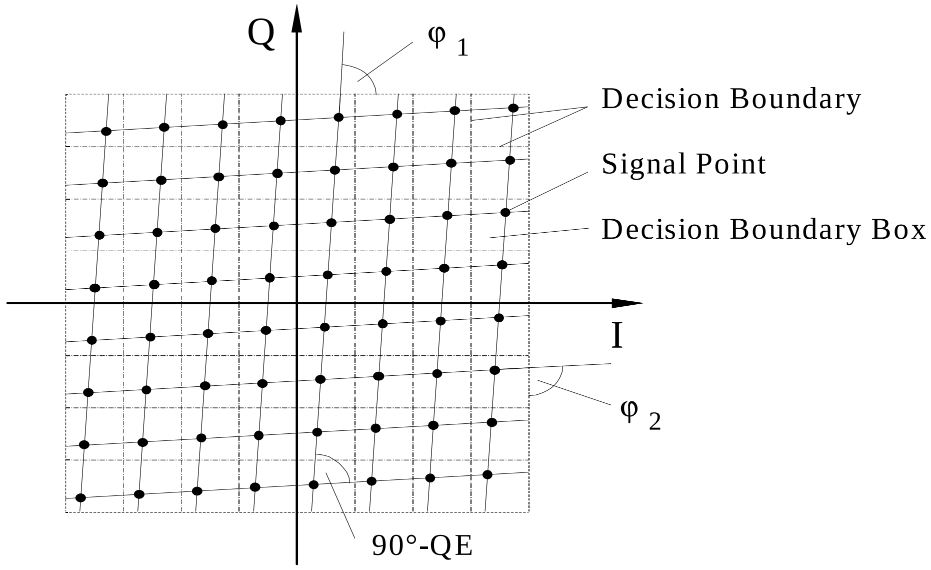
6.9.4 Carrier suppression

Purpose	A residual carrier is an unwanted coherent CW signal added to the QAM signal. It may have been produced by DC offset voltages of the modulating I and/or Q signal or by crosstalk from the modulating carrier within the modulator.
Interface	E, G, S, T
Method	Search for systematic deviations of all constellation points and isolate the residual carrier. Calculate the Carrier Suppression (CS) from the formula: $$CS = 10 \times \log_{10} \left(\frac{P_{sig}}{P_{RC}} \right)$$ where P_{RC} is the power of the residual carrier and P_{sig} is the power of the QAM signal (without residual carrier).

6.9.5 Amplitude Imbalance (AI)

Purpose	To separate the QAM distortions resulting from AI of the I and Q signal from all other kind of distortions.
Interface	E, G, S, T
Method	Calculate the I and Q gain values v_I and v_Q from all points in a constellation diagram eliminating all other influences. Calculate AI from v_I and v_Q: $$AI = \left(\frac{v_2}{v_1} - 1\right) \times 100\,\%$$ $$with \qquad v_1 = \min(v_I, v_Q) \quad \text{and} \quad v_2 = \max(v_I, v_Q)\ .$$ $$v_I = \frac{1}{M}\sum_{i=1}^{M}\frac{I_i + \left(\bar{d}_i\right)_I}{I_i}$$ $$\left(\bar{d}_i\right)_I = \frac{1}{N}\sum_{j=1}^{N}\delta I_j \qquad \text{(I - component of } d_i \text{ as given in clause 6.9.3)}$$ $$v_Q = \frac{1}{M}\sum_{i=1}^{M}\frac{Q_i + \left(\bar{d}_i\right)_Q}{Q_i}$$ $$\left(\bar{d}_i\right)_Q = \frac{1}{N}\sum_{j=1}^{N}\delta Q_j \qquad \text{(Q - component of } d_i \text{ as given in clause 6.9.3)}$$ $$\left(\bar{d}_i\right)_I + \left(\bar{d}_i\right)_Q = \bar{d}_i$$

6.9.6 Quadrature Error (QE)

Purpose	The phases of the two carriers feeding the I and Q modulators have to be orthogonal. If their phase difference is not 90° a typical distortion of the constellation diagram results. The receiver usually aligns its reference phase in such a way that the 90° error ($\Delta\varphi$) is equally spread between φ_1 and φ_2.  **Figure 6.3: Distortion of constellation diagram resulting from I/Q Quadrature Error (QE)**
Interface	E, G, S, T
Method	Search for the constellation diagram error shown in figure 6.3 and calculate the absolute value of the phase difference $\Delta\varphi = \|\varphi_1 - \varphi_2\|$ after having eliminated all other influences and convert this into degrees. $QE = \frac{180°}{\pi} \times \|\varphi_1 - \varphi_2\| \quad [°]$

6.9.7 Residual Target Error (RTE)

Void.

6.9.8 Coherent interferer

Purpose	Coherent interferers (not necessarily related to the main carrier) are usually measured with a spectrum analyser (out of service, and in some cases in service with narrow resolution bandwidth filter and video filter at interfaces N and P) or either of the following methods described below (in service). In a constellation diagram a sine-wave interferer will change the noisy clouds of each system point into a "donut" shape. From the statistical distribution of the clouds, the amplitude of the interferer can be calculated if it is above a certain limit. If the frequency of the interferer is of interest or more than one interferer is present, the Fourier transform method should to be used.
Interface	E, G, S, T
Method	Perform a Fourier transform of a time record of error vectors to produce a frequency spectrum of the interferers. Alternatively, calculate the RMS magnitude a_i of the coherent interferer preferably from the statistical distribution of the 4 inner clouds computed from the measurement sample. Normalize a_i to S_{rms} and express the result in dB. $C/I = 20 \times \log_{10} \frac{S_{rms}}{a_i}$ [dB]

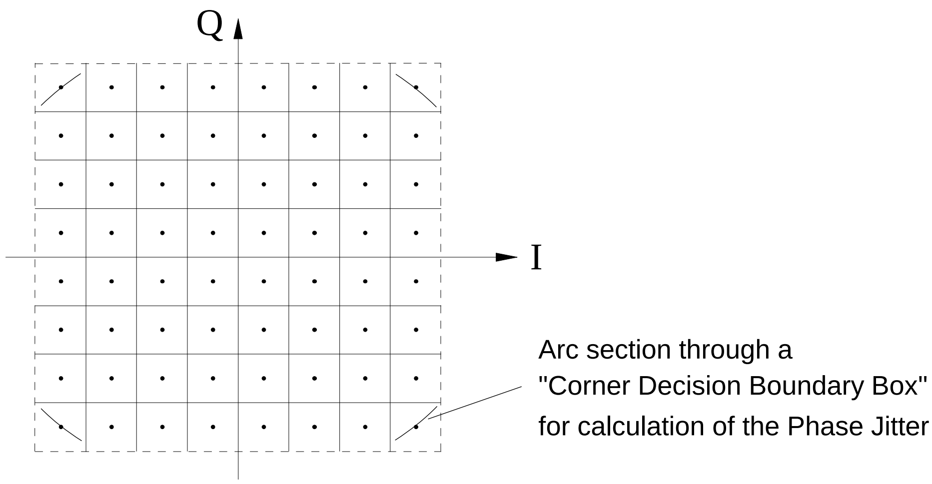
NOTE 1: In the present document, the term "coherent" is applied to signals that have a high degree of correlation with a time shifted version of itself.

EXAMPLE 1: Continuous Waves (CW) or even single channel analogue video modulated carriers, these signals are coherent although they do not need to be related to the carrier of the digital channel under test.

NOTE 2: Non-coherent is applied to signals with very low correlation to a time shifted version of themselves.

EXAMPLE 2: Random noise or digitally modulated carriers, as well as the combined result of inter-modulation by many carriers.

6.9.9 Phase Jitter (PJ)

Purpose	The PJ of an oscillator is due to fluctuations of its phase or frequency. Using such an oscillator to modulate a digital signal results in a sampling uncertainty in the receiver, because the carrier regeneration cannot follow the phase fluctuations. The signal points are arranged along a curved line crossing the centre of each decision boundary box as shown in figure 6.4 the four "corner decision boundary boxes".  **Figure 6.4: Position of arc section in the constellation diagram to define the PJ (example: 64-QAM)**
Interface	E, G, S, T

Method	Phase Jitter (PJ) can be calculated theoretically using the following algorithm: For every received symbol: 1) Calculate the angle between the I-axis of the constellation and the vector to the received symbol $(\tilde{I},\tilde{Q})$: $$\phi_1 = \arctan\frac{\tilde{Q}}{\tilde{I}}$$ 2) Calculate the angle between the I-axis of the constellation and the vector to the corresponding ideal symbol (I,Q): $$\phi_2 = \arctan\frac{Q}{I}$$ 3) Calculate the error angle: $$\phi_E = \phi_1 - \phi_2$$ From these N error angles calculate the RMS phase jitter: $$PJ = \sqrt{\frac{1}{N}\sum_{i=1}^{N}\phi_{E_i}^2 - \frac{1}{N^2}\left(\sum_{i=1}^{N}\phi_{E_i}\right)^2}$$ However, the following method may be more practical. The first approximation of the "arc section" of a "corner decision boundary box" is a straight line parallel to the diagonal of the "decision boundary box". Additionally the curvature of the Phase Jitter (PJ) trace has to be taken into account when calculating the standard deviation of the PJ. The mean value of the PJ is calculated in degrees. $$PJ = \frac{180°}{\pi}\times\arcsin\left(\frac{\sigma_{PJ}}{\sqrt{2}\times\left(\sqrt{M}-1\right)\times d}\right)$$ where M = order of QAM and 2d = distance between two successive boundary lines. Within the argument of the arc sine function, the standard deviation of the PJ is referenced to the distance from the centre of the "corner decision boundary box" to the centre point of the QAM signal.

6.9.10 Signal-to-Noise Ratio (SNR)

Purpose	See 6.9.1
Interface	S, T
Method	See 6.9.1, G.8, A.3

6.10 Interference

Purpose	In a CATV network interference products can be caused by modulators and frequency converters.
Interface	N (out of service) or S, T (in service).
Method	Out of service interference products are measured with a spectrum analyser and in some cases in-service measurements can be done if a narrow resolution bandwidth filter and video filtering is used to lower the response of the instrument to the signal spectrum. If the frequency of the expected interference is known, the measurement can be made easily and quickly. In-service information of coherent interference can be derived from the constellation, clause 6.9.8. In some circumstances the residual carrier level can be measured with a spectrum analyser, by using a narrow resolution bandwidth filter and video filtering, at the interfaces H, J, N, P. The CS can be calculated as ten times the logarithm (base 10) of the ratio of the signal power measured as described in clause 6.6, to the measured remaining carrier power.

7 Cable specific measurements

7.0 Introduction

In SMATV networks that distribute the 1st satellite IF directly to subscribers, some parameters of this clause can be defined accordingly for QPSK modulated signals.

7.1 Noise margin

Purpose	To provide an indication of the reliability of the transmission channel. The noise margin measurement is a more useful measure of system operating margin than a direct BER measurement due to the steepness of the BER curve.
Interface	The reference interface for the noise injection is the RF interface (N). For practical implementation, other interfaces can be used, provided equivalence can be shown, for example P.
Method	The noise margin is computed by adding white Gaussian noise on the received signal. The noise margin will be the difference in dB between the carrier to noise ratio (C/N) of the received signal and the carrier to noise ratio for a BER of 10^{-4} (before RS decoding).

7.2 Estimated noise margin

Purpose	To provide an indication of the reliability of the transmission channel without switching off the service. The noise margin measurement is a more useful measure of system operating margin than a direct BER measurement due to the steepness of the BER curve.
Interface	T
Method	The estimated noise margin is computed by simulating the addition of white Gaussian noise to the demodulated data and predicting the resulting BER by statistical methods. The noise margin will be the difference in dB between the estimated SNR of the received signal and the synthesized SNR which gives a predicted BER of 10^{-4} (before RS decoding).

7.3 Signal quality margin test

Void.

Figure 7.1: Quality thresholds for single constellation in the I/Q plane (void)

7.4 Equivalent Noise Degradation (END)

Purpose	END is a measure of the implementation loss caused by the network or the equipment where the reference is the ideal performance.
Interface	T (BER) and N or P or R (noise injection)
Method	The END is obtained from the difference in dB of the C/N or E_b/N_0 ratio needed to reach a BER of 10^{-4} and the C/N or E_b/N_0 ratio that would theoretically give a BER of 10^{-4}, for a Gaussian channel.

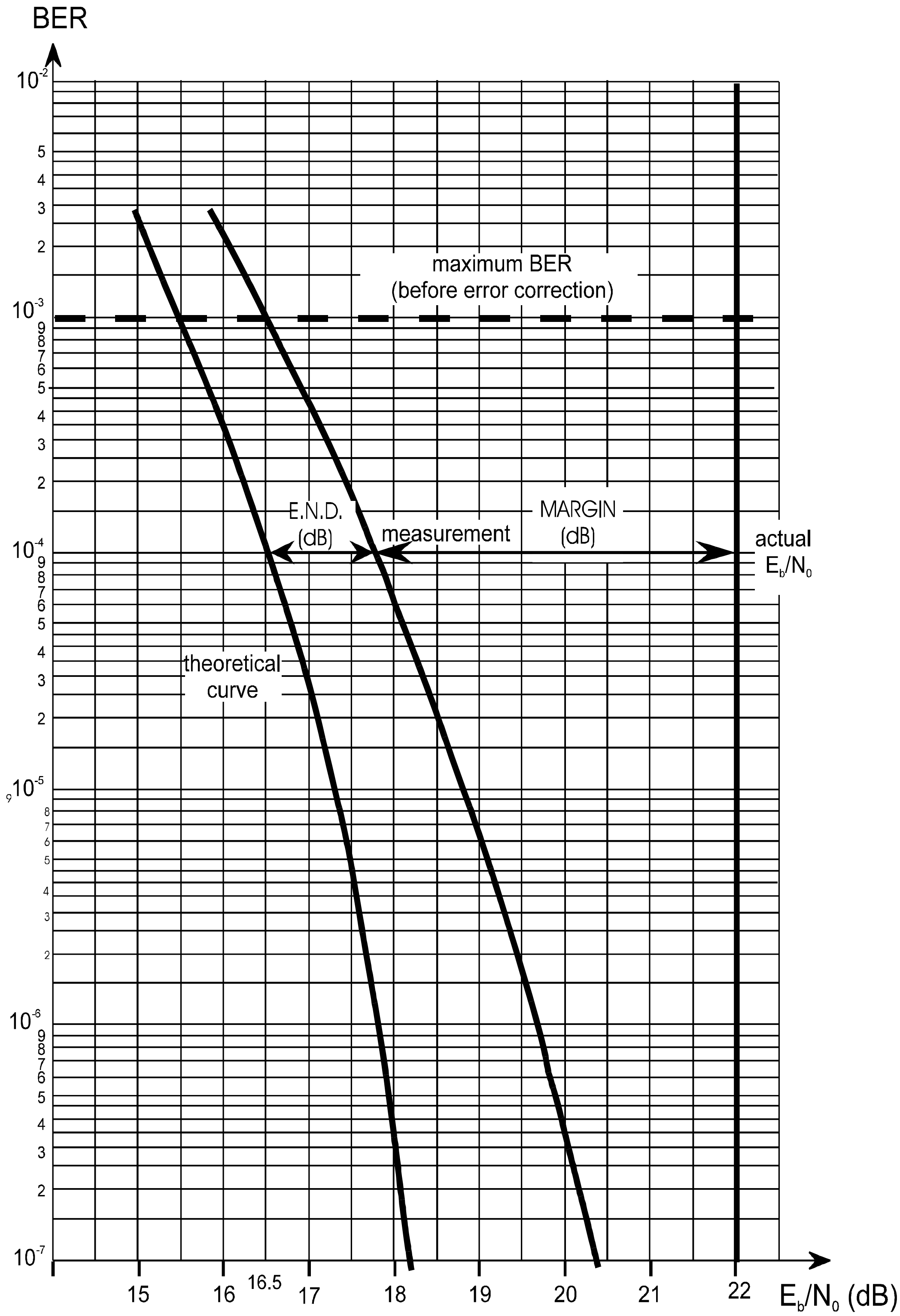


Figure 7.2: Measurement of equivalent noise degradation

Figure 7.2 gives a quantitative curve of the BER in DVB-C systems.

7.5 BER vs. E_b/N_0

Purpose	The BER vs. E_b/N_0 measurement enables a graph to be drawn which shows the implementation loss of the system over a range of Bit Error Rates. The residual BER at high E_b/N_0 values is an indicator of possible network problems. C/N measurements can be converted to E_b/N_0 as shown $$E_b/N_0 = C/N + 10\log_{10}\frac{BW_{noise}}{f_s \times m} \text{ [in dB]}$$ m is the number of bits per symbol (m = 6 for 64-QAM) and N is measured in the Nyquist bandwidth (symbol rate as indicated in clause 6.7).
Interface	T (BER) and N or P or R (noise injection)
Method	The BER vs. E_b/N_0 curve will be measured using the RF and noise power measurements described above. The BER range of interest is 10^{-7} to 10^{-3}. The E_b/N_0 value is based on the gross bitrate (including RS error correction) and the net bitrate value of E_b/N_0 can easily be calculated using the RS rate, using the following conversion factor for a RS (204, 188) code (see annex G). $$10 \times \log_{10}\left(\frac{204}{188}\right) = +0{,}35 \ dB$$

7.6 Phase noise of RF carrier

Purpose	Phase noise can be introduced at the transmitter side or by the receiver due to unstable local oscillators. Phase noise outside the loop bandwidth of the carrier recovery circuit leads to a circular smearing of the constellation points in the I/Q plane. This reduces the operating margin (noise margin) of the system and may directly increase the BER.
Interface	Any RF/IF interface, N, P
Method	Phase noise power density is normally expressed in dBc/Hz at a certain frequency offset from the carrier. Out of service phase noise will be measured with a spectrum- or modulation- analyser.

7.7 Amplitude, phase and impulse response of the channel

Purpose	Linear distortions, like amplitude and phase response errors and echoes, will be caused for instance by long lengths of cable and the cascading of a high number of amplifiers. The impulse response is important to localize the discrete reflections that may occur in cable networks.
Interface	S, T
Method	The impulse response of the transmission channel can be calculated (inverse Fourier transform) from the amplitude and phase response. The amplitude and phase response are defined as the RF-channel response. The amplitude response of the transmission channel can be derived from the equalizer tap coefficients or can be calculated directly from the "I" and "Q" samples, for example by using auto- and cross-correlation functions.

7.8 Out of band emissions

Purpose	To prevent interference in other channels in the network the RF signal should comply with the spectrum mask specified for the network under test.
Interface	Transmitter output, J
Method	Spectrum analyser

8 Satellite specific measurements

8.1 BER before Viterbi decoding

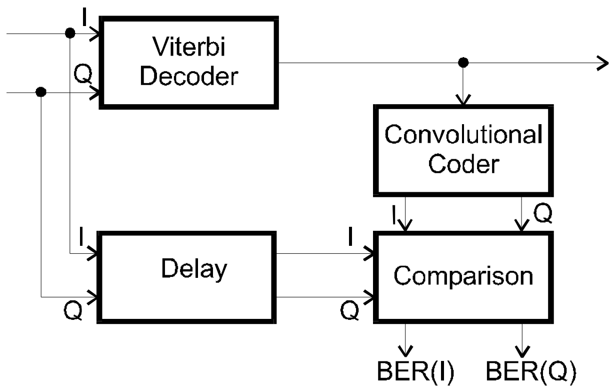
<table>
<tr><td>Purpose</td><td>This measurement gives an indication of the transmission link performance. Due to typical error rates ranging from 7×10^{-2} to 10^{-5} the measurement can be done in a reasonable amount of time. Outside of this range the accuracy of the results may not be guaranteed.</td></tr>
<tr><td>Interface</td><td>The measurement should be done before the Viterbi decoder (Interface T of the receiver).</td></tr>
<tr><td>Method</td><td>The signal after Viterbi decoding in the measurement instrument is coded again using the same coding scheme as in the transmitter, in order to produce an estimate of the originally coded I and Q sequences. These sequences are compared at bit level with the sign-values of the signals that are available before Viterbi decoding.
The BER for the I and Q paths should be made available separately. The measurement should be based on at least several hundred bit errors. For fast evaluation, in the case that the BER is lower than 10^{-4}, it should be possible to stop the measurement after approximately 1 second.
For accurate measurement of E_b/N_0 at the quasi error free threshold, the measurement time and the presentation of the result should be such that an accuracy of three decimal place can be achieved. The quasi error free threshold corresponds to a BER before Viterbi decoding in the range 7×10^{-2} to 7×10^{-3},depending on the selected convolutional code rate; or a BER after Viterbi decoding of 2×10^{-4}.


Figure 8.1: BER measurement before Viterbi decoding</td></tr>
</table>

8.2 Receive BER vs. E_b/N_0

Purpose	To verify overall clear sky link performance and link margin using a reference down link for acceptance tests.
Interface	After Viterbi decoding, V
Method	This is an out-of-service-measurement. The BER measurement should be based on the null packets inserted at the modulator as defined in clause A.1. To obtain the various values necessary for the curve BER over E_b/N_0, white Gaussian noise is injected at the receiver site. In order to get accurate results it should be verified that the inserted noise level is at least 15 dB above the system noise. This can easily be observed on a spectrum analyser by switching the inserted noise on and off. Stable reception conditions are a precondition for accurate measurement results. The RS decoding should be deactivated, or bypassed to avoid excessively long measurement periods. The BER range of interest is 10^{-9} to 10^{-2}. The measurement values are compared with the theoretical values. The value for the Equivalent Noise Degradation (END) at a BER of 10^{-4} can be derived from this information as well. For evaluation of E_b/N_0 only the number of information bits (the net bitrate) should be taken into account.

8.3 IF spectrum

Purpose	To prevent interference into other channels and to be compliant with the DVB specification the modulator output spectrum should be according with the one specified in ETSI EN 300 421 [i.5].
Interface	H, input of the up-converter, typically 70 MHz or 140 MHz (Modulator output plus equipment for the connection to the up-converter input).
Method	Spectrum analyser and template for amplitude response, network analyser and template for group delay response, both as specified in ETSI EN 300 421 [i.5].

9 Measurements specific for a terrestrial (DVB-T) system

9.0 Introduction

The intention of these guidelines is to provide a list of measurements useful in a DVB-T OFDM environment. The different options could be selected by the users of the system. Equipment manufacturers (both transmitters and receivers) as well as the operators, can choose those measurements that best fit their needs. A list of the applicability of the measurement parameters described in the present document to the DVB-T transmitter, receiver and network is given in table 9.1.

The measurements 6.1 "System availability" and 6.2 "Link availability" are also valid for Terrestrial (not only for Cable and Satellite) and for any contribution link like SDH, PDH, etc.

Table 9.1: DVB-T measurement parameters and their applicability

Measurement parameter	Transmitter	Network	Receiver
1) RF frequency measurements			
1.1) RF frequency accuracy (Precision)	X		
1.2) RF channel width (Sampling Frequency Accuracy)	X		
1.3) Symbol Length measurement at RF (Guard Interval verification)	X		
2) Selectivity			X
3) AFC capture range			X
4) Phase noise of local oscillators (LO)	X		X
5) RF/IF signal power	X	X	X
6) Noise power			X
7) RF and IF spectrum	X		
8) Receiver sensitivity/dynamic range for a Gaussian channel			X
9) Equivalent Noise Degradation (END)	X		X
9a) Equivalent Noise Floor (ENF)	X		
10) Linearity characterization (shoulder attenuation)	X		
11) Power efficiency	X		
12) Coherent interferer	X	X	
13) BER vs. C/N ratio by variation of transmitter power	X	X	
14) BER vs. C/N ratio by variation of Gaussian noise power		X	X
15) BER before Viterbi (inner) decoder	X	X	X
16) BER before RS (outer) decoder	X	X	X
17) BER after RS (outer) decoder	X	X	
18) I/Q analysis			
18.1) N/A			
18.2) Modulation Error Ratio	X	X	X
18.3) System Target Error	X		X
18.4) Carrier Suppression	X		X
18.5) Amplitude Imbalance	X		X
18.6) Quadrature Error	X		X
18.7) Phase Jitter	X		X
19) Overall signal delay	X	X	
20) SFN synchronization			
20.1) MIP_timing_error	X		
20.2) MIP_structure_error	X		

Measurement parameter	Transmitter	Network	Receiver
20.3) MIP_presence_error	X		
20.4) MIP_pointer_error	X		
20.5) MIP_periodicity_error	X		
20.6) MIP_ts_rate_error	X		
21) System Error Performance	X	X	X

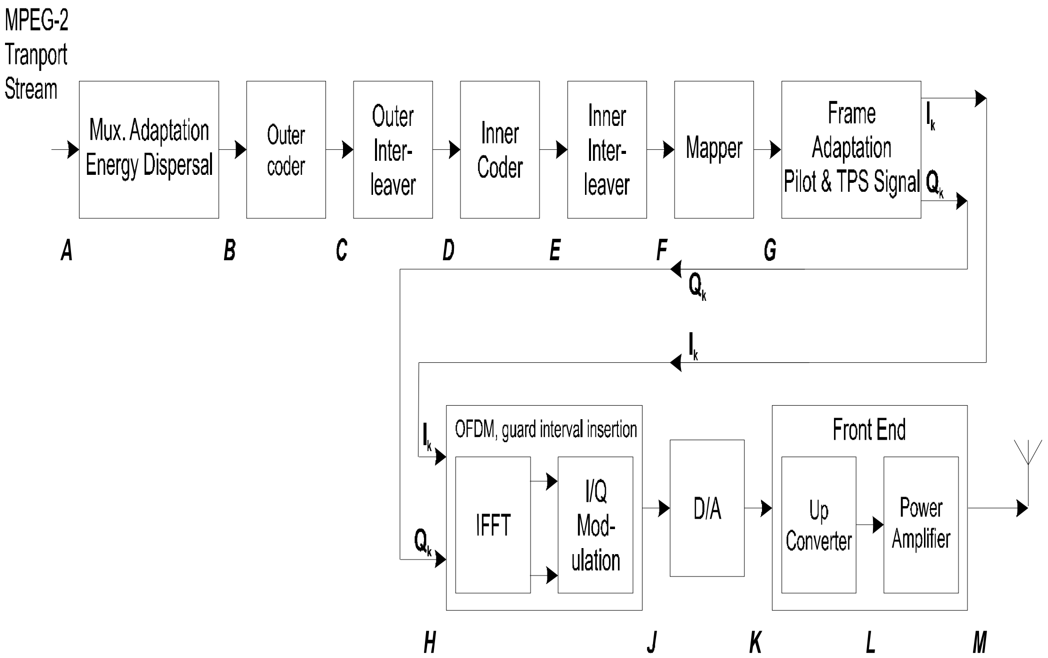


Figure 9.1: Block diagram of a DVB-T transmitter

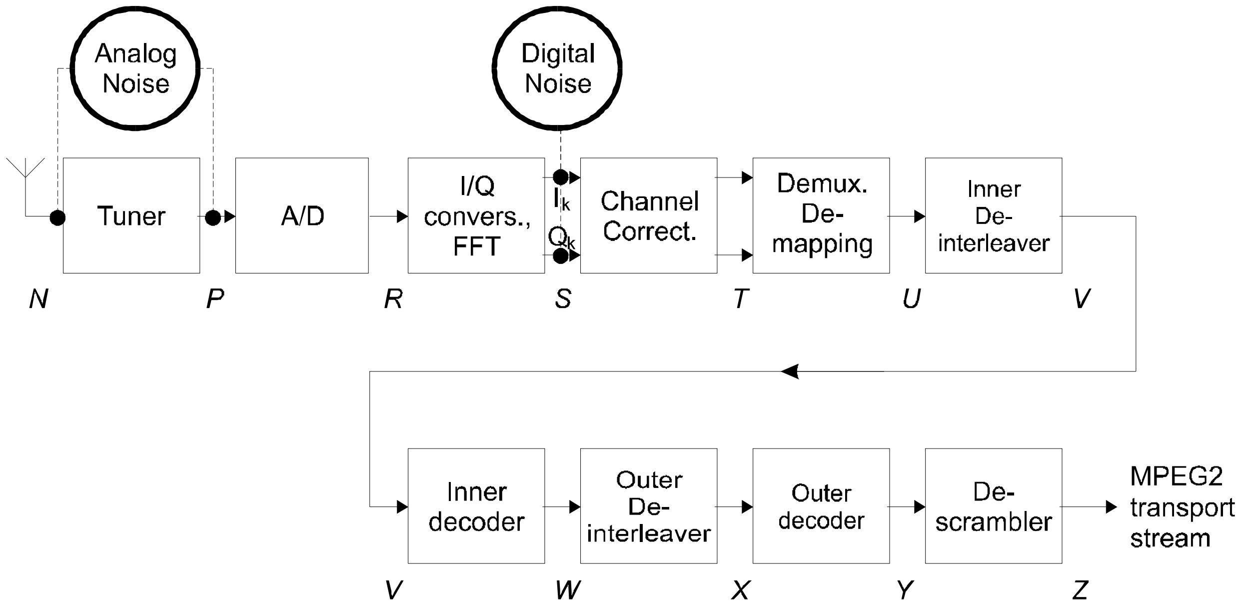


Figure 9.2: Block diagram of a DVB-T receiver

9.1 RF frequency measurements

9.1.0 General

The accuracy of some basic parameters of the OFDM modulation may be carried out at the RF layer of the DVB-T signal.

9.1.1 RF frequency accuracy (Precision)

Purpose	Successful processing of OFDM signals requires that certain carrier frequency accuracy be maintained at the transmitter. Specific network operations modes such as SFN require high accuracy of the carrier frequency.
Interface	L, M
Method	The 8k mode of the DVB-T always has a continual pilot, with continuous phase along successive OFDM symbols, exactly at the channel centre (k = 3 408). Its frequency may be directly measured by any spectrum analyser that has an integrated counter and at least a resolution filter of 300 Hz or less (if necessary by utilizing a reference source of sufficient accuracy). The 2k mode has a continual pilot with continuous phase at k = 1 140. Its frequency may be directly measured by any spectrum analyser that has an integrated counter and at least a resolution filter of 300 Hz or less (if necessary by utilizing a reference source of sufficient accuracy). The centre channel frequency may be inferred by subtracting to the measured frequency: **8 MHz channels:** **285 714 Hz** i.e. [(1 140 - 852) × 4 464,2 857 = 1 285 714 Hz]. **7 MHz channels:** **1 125 000 Hz** i.e. [(1 140 - 852) × 3 906,25 = 1 125 000 Hz]. **6 MHz channels:** **964 286 Hz** i.e. [(1 140 - 852) × 4 464,2857 = 964 286 Hz]. NOTE: For 2k mode this method may have some inaccuracy if the sampling frequency of the modulator is not precise, however such error in the sampling frequency would need to be very high to significantly affect the centre channel measurement. Should more accuracy needed, the two outer continual pilots may be measured as indicated under clause 9.1.2 RF channel width, and the mean of the two values be calculated.

9.1.2 RF channel width (Sampling Frequency Accuracy)

Purpose	Channel width measurements are convenient for verification that sampling frequency accuracy is maintained at the modulator side.
Interface	L, M
Method	The occupied bandwidth of a COFDM modulated channel depends directly from the frequency spacing and this from the sampling frequency. The outermost carriers in a DVB-T signal are continual pilot carriers. Their frequencies are measured (see clauseE.1) and the difference between them should be compared to the nominal channel width of 7 607 142,857 Hz for 8 MHz channels, 6 656 250,000 Hz for 7 MHz channels and 5 705 357,143 Hz for 6 MHz channels. NOTE: Three decimal places are given here for completeness only. Accuracy of 1 Hz at 5 MHz means 0,2 × 10^{-6} per Hz, which may be enough for most cases of sampling frequency measurement. Measurement instruments should have better accuracy and resolution (typically in the order of ten times) than the required measurement accuracy. If the frequency of the outermost carriers is known, see clauses E.1.3 and E.1.4 for how to measure them, then the related values may be calculated as per table below. Denoting the outermost pilot frequencies as F_L and F_H appropriately the occupied bandwidth is OB = F_H _ F_L. The number of carriers is K, and for the 2k mode K-1 = 1 704 while for the 8k mode K-1 = 6 816.

Table 9.2: Calculated values

	8k mode	2k mode
Occupied bandwidth	**F_H - F_L**	
Frequency Spacing	$(F_H - F_L)/6\ 816$	$(F_H - F_L)/1\ 704$
Useful duration	$6\ 816/(F_H - F_L)$	$1\ 704/(F_H - F_L)$
Centre channel 1st IF	$(F_H - F_L) \times 4\ 096/(K-1)$	$(F_H - F_L) \times 1\ 024/(K-1)$
Sampling Frequency	$(F_H - F_L) \times 16\ 384/(K-1)$	$(F_H - F_L) \times 4\ 096/(K-1)$

9.1.3 Symbol Length measurement at RF (Guard Interval verification)

Purpose	Verification of the guard interval used in a received DVB-T signal may be carried out at RF level by careful frequency measurements. This measurement is valid in cases where there is an uncertainty on whether a modulator is correctly working and producing a signal with the expected or assigned Guard Interval.
Interface	L, M
Method	The scattered pilots produce a pulsed-like spectrum every third carrier in a DVB-T spectrum due to their repetition presence at the same phase and location every fourth symbol. The frequency difference between two contiguous spectral lines representing a scattered pilot represents the inverse of the time length of four consecutive DVB-T symbols. Measuring such frequency difference and dividing its inverse by 4 will provide the total symbol length T_S of the measured signal. By subtracting the nominal useful symbol duration T_U the length of the GI is found. See clause E.1 for details on the measurement procedure and symbol lengths.

9.2 Selectivity

Purpose	To identify the capability of the receiver to reject out-of-channel interference.
Interface	The measurement of the signal input level and the interferer should be carried out at the interface N, using interface W or X for the BER monitoring.
Method	The input power is adjusted to 10 dB above the minimum input power as defined in "Receiver sensitivity" (see clause 9.8). The C/I threshold needed for QEF operation after RS decoder (BER < 2×10^{-4} before RS decoder) should be measured as a function of the frequency of a CW interferer.

9.3 AFC capture range

Purpose	To determine the frequency range over which the receiver will acquire overall lock.
Interface	N, for the application of the test signal; Z, for the test of TS synchronization
Method	A signal is applied to the input of the receiver, at a level 10 dB above the minimum input power as defined in "Receiver sensitivity" (see clause 9.8). The signal is frequency shifted in steps (from below and above) towards a nominal value and the Sync_byte_error is verified according to clause 5.2.1 (Measurement and analysis of the MPEG-2 *TS* - First priority: necessary for decodability (basic monitoring)).

9.4 Phase noise of Local Oscillators (LO)

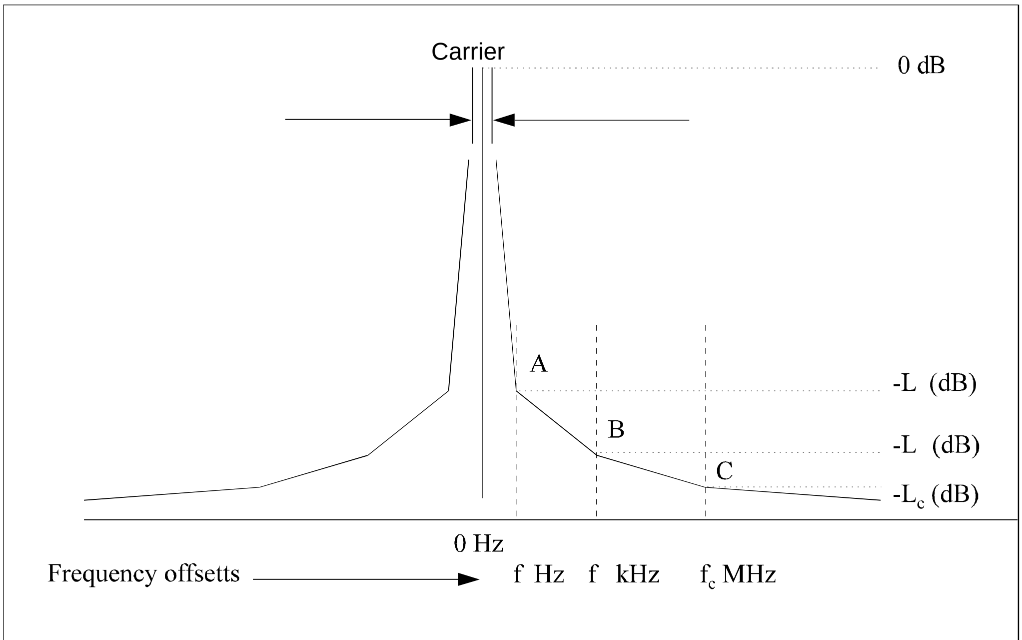
Purpose	Phase noise can be introduced at the transmitter, at any frequency converter or by the receiver due to random perturbation of the phase of the oscillators. In an OFDM system the phase noise can cause Common Phase Error (CPE) which affects all carriers simultaneously, and which can be minimized or corrected by using the continual pilots. However the Inter-Carrier Interference (ICI) is noise-like, cannot be corrected. The effects of CPE are similar to any single carrier system and the phase noise, outside the loop bandwidth of the carrier recovery circuit, leads to a circular smearing of the constellation points in the I/Q plane. This reduces the operating margin (noise margin) of the system and may directly increase the BER. The effects of ICI are peculiar to OFDM and cannot be corrected for. This has to be taken into account as part of the total noise of the system.
Interface	Any access to Local Oscillators (LO), in transmitters, converters and receivers.
Method	Phase noise can be measured with a spectrum analyser, a vector analyser or a phase noise test set.
Method for CPE	Phase noise power density is normally expressed in dBc/Hz at a certain frequency offset from the local oscillator signal. It is recommended to specify a spectrum mask with at least three points (frequency offsets and levels), for example see figure 9.3. NOTE: See clauses A.4 and E.4 for additional information on phase noise measurements. See clause E.4.1 for some practical information.
	 Possible mask for CPE measurements. The points A, B and C to be defined. **Figure 9.3: Possible mask for CPE measurements**
Method for ICI	For the measurement of ICI, the use of multiples of the carrier spacing is recommended for the frequencies, f_a, f_b, f_c.

Table 9.3: Frequency offsets for 2 k and 8 k systems

2 k system	4,5 kHz	8,9 kHz	13,4 kHz
8 k system	1,1 kHz	2,2 kHz	3,4 kHz

Typical use For manufacturing, incoming inspection and maintenance of modulators, transmitters, up/down converters and receivers, either professional or consumer type.

9.5 RF/IF signal power

Purpose	Signal power, or wanted power, measurement is required to set and check signal levels at the transmitter and receiver sites.
Interface	K, L, M, N, P
Method	The signal power of a terrestrial DVB signal, or wanted power, is defined as the mean power of the signal as would be measured with a thermal power sensor. In the case of received signals care should be taken to limit the measurement to the bandwidth at the wanted signal. When using a spectrum analyser or a calibrated receiver, it should integrate the signal power within the nominal bandwidth of the signal ($n \times f_{SPACING}$) where n is the number of carriers.

9.6 Noise power

Purpose	Noise is a significant impairment in a transmission network.
Interface	N,P
Method	The noise power (mean power), or unwanted power, can be measured with a spectrum analyser (out of service). The noise power is specified using the occupied bandwidth of the OFDM signal ($n \times f_{SPACING}$) where n is the number of carriers.

NOTE: The term C/N should be calculated as the ratio of the signal power, measured as described in clause 9.5, to the noise power, measured as described in this clause.

9.7 RF and IF spectrum

Purpose	To avoid interfering with other channels, the transmitted RF spectrum should comply with a spectrum mask, which is defined for the terrestrial network. If the spectrum at the modulator output is defined by a spectrum mask, the same procedure can be applied to the IF signal (with no pre-correction active).
Interface	K, M
Method	This measurement is usually carried out using a spectrum analyser. The spectral density of a terrestrial DVB signal is defined as the long-term average of the time-varying signal power per unity bandwidth (i.e. 1 Hz). Values for other bandwidths can be achieved by proportional increase of the values for unity bandwidth. To avoid regular structures in the modulated signal a non-regular, e.g. a Pseudo-Random Binary Sequence (PRBS) - like or a programme type digital transmitter input signal is necessary. Care has to be taken that the input stage of the selective measurement equipment is not overloaded by the main lobe of the signal while assessing the spectral density of the side lobes, i. e. the out-of-band range. Especially in cases with very strong attenuation of the side lobes non-linear distortion in the measurement equipment can produce side lobe signals that mask the original ones. Selective attenuation of the main lobe has proven to be in principal a way to avoid this masking effects. However, as the frequency response of the band-stop filter has to be included in the evaluation, the whole measurement procedure may become somewhat complex. For the resolution bandwidth, the recommended values should not exceed 30 kHz. Preferred values are approximately 4 kHz. The measurement should be Noise-normalized to 4 kHz.

9.8 Receiver sensitivity/dynamic range for a Gaussian channel

Purpose	For network planning purposes, the minimum and maximum input powers for normal operation of a receiver have to be determined.
Interface	Test signals are applied and measured at interface N; interfaces W or X are used for the monitoring of BER before RS.
Method	The minimum and maximum input power thresholds for QEF (Quasi Error Free) operation after the RS decoder (i.e. BER < 2×10^{-4} before RS decoding) should be measured. The dynamic range is the difference between the measured values.

9.9 Equivalent Noise Degradation (END)

9.9.0 General

Purpose	END is a measure of the implementation loss caused by the network or the equipment where the reference is the ideal performance.
Interface	W or X for BER measurement; N, P or S for noise injection
Method	The END is obtained from the difference in dB of the C/N ratio needed to reach a BER of 2×10^{-4} before RS (outer) decoding, and the C/N ratio that would theoretically give a BER of 2×10^{-4} for a Gaussian channel (see annex A of ETSI EN 300 744 [i.9]).

9.9.1 Equivalent Noise Floor (ENF)

Purpose	ENF is a measure of the implementation loss caused by the transmitting equipment where the reference is the ideal transmitter.
Interface	M for noise power measurement, W or X for BER measurement; N, P or S for noise injection
Method	The ENF is obtained from the measurement of additional noise needed to reach a BER of 2×10^{-4} before RS (outer) decoding, and the noise level that would theoretically give a BER of 2×10^{-4} for a Gaussian channel (see annex A of ETSI EN 300 744 [i.9]) as described in clause B.12.

Note on END and ENF:

The impact of the DVB-T transmitter on the overall system performance, when a certain DVB-T mode is being received by the reference receiver, via a Gaussian channel, is assessed by the measurement of the END.

The reference receiver is in the present document defined as a DVB-T receiver which require a C/N which is 3,0 dB higher than the C/N figures indicated in ETSI EN 300 744 [i.9], on a Gaussian channel.

The END is in the present document defined to be the difference between required C/N, for a BER of 2×10^{-4} after convolutional decoding on the reference receiver, using a real and an ideal DVB-T transmitter.

The END is not only a characteristic of the transmitter itself but is also dependent on the used DVB-T mode and on the receiver implementation loss (this is why a fixed 3,0 dB receiver implementation loss is defined for the reference receiver).

The END should not exceed [0,5] dB and should be independent of the selected guard interval. Depending on the requirements of the network operator typical END values fall in the range [0,1-0,4] dB.

For the determination of the END value another parameter, the Equivalent Noise Floor ENF, can be used. As described in clause B.12, this should result in an improved accuracy for the END.

As opposed to the END the ENF is relatively independent of the DVB-T mode used and on the receiver implementation loss and can therefore be used to characterize the transmitter *by itself*. Depending on whether there is a need for characterizing the DVB-T transmitter by itself, or whether there is a need to characterize *its effect on a receiver*, the ENF can sometimes be used as an alternative to END as a performance parameter.

The influences of intermodulation and amplitude ripple are expected to dominate in practise in the performance parameter END.

(The Group Delay response of a transmitter needs to be defined by network operators depending on the configuration in use (channel combiners, output filters, etc.).)

9.10 Linearity characterization (shoulder attenuation)

Purpose	The "shoulder attenuation" can be used to characterize the linearity of an OFDM signal without reference to a spectrum mask.
Interface	M
Method	Apply the following procedure on the measured RF spectrum of the transmitter output signal: (a) Identify the maximum value of the spectrum by using a resolution bandwidth at approximately 10 times the carrier spacing. (b) Place declined, straight lines connecting the measurement points at 300 kHz and 700 kHz from each of the upper and lower edges of the spectrum. Draw additional lines parallel to these, so that the highest spectrum value within the respective range lies on the line. (c) Subtract the power value of the centre of the line (500 kHz away from the upper and lower edge) from the maximum spectrum value of (a) and note the difference as the "shoulder attenuation" at the upper and lower edge. (d) Take the worst case value of the upper and lower results from (c) as the overall "shoulder attenuation". NOTE: For a quick overview the value at e.g. 500 kHz can be measured directly provided that coherent interferers are not present.

9.11 Power efficiency

Purpose	To compare the overall efficiency of DVB transmitters.
Interface	M
Method	Power efficiency is defined as the ratio of the DVB output power to the total power consumption of the chain from TS input to the RF signal output including all necessary equipment for operation such as blowers, transformers etc. (and is usually quoted in % terms). The operational channel and the environmental conditions need to be specified.

9.12 Coherent interferer

Purpose	To identify any coherent interferer which may influence the reliability of the I/Q analysis or the BER measurements.
Interface	N or P
Method	The measurement is carried out with a spectrum analyser. The resolution bandwidth is reduced stepwise so that the displayed level of the modulated carriers *(and of the unmodulated pilots, due to the influence of the guard interval)* is reduced. The CW interferer is not affected by this process and can be identified after appropriate averaging of the trace.

9.13 BER vs. C/N ratio by variation of transmitter power

Purpose	To evaluate the BER performance of a transmitter as the Carrier to Noise (C/N) ratio is varied, with the measurement repeated for a range of mean transmitted output powers. This measurement can be used to compare the performance of a transmitter with theory or with other transmitters.
Interface	From F to U or from E to V
Method	A Pseudo-Random Binary Sequence (PRBS) is injected at interface F (or E). The various C/N ratios are established at the input of the test receiver by addition of Gaussian noise, and the BER of the received PRBS is measured at point V (or U) using a BER TEST Set. The measurement is repeated for a range of mean transmitted output power. If the ability to generate a PRBS at interface F (or E) is included in the transmitting equipment for test purposes, then it should be a 2^{23} - 1 PRBS as defined by Recommendation ITU-T O.151 [i.12]. For the measurement of carrier and noise power, the system bandwidth is defined as $n \times f_{SPACING}$, where n is the number of active carriers (e.g. 6 817 or 1 705 carriers in an 8 MHz channel) and $f_{SPACING}$ is the frequency spacing of the OFDM carriers.

NOTE: Transmitter back-off is defined as the ratio of the rated pulsed peak power of the transmitter to the mean power of the signal. The rated pulsed peak power is normally equivalent to the peak sync power of a standard B, D, G, H, I or K RF signal.

9.14 BER vs. C/N ratio by variation of Gaussian noise power

Purpose	To evaluate the BER performance of a receiver as the Carrier to Noise (C/N) ratio is varied by changing the added Gaussian noise power. This measurement can be used to compare the performance of a receiver with theory or with other receivers. For example to evaluate the influence of receiver noise floor.
Interface	From F to U or from E to V.
Method	A Pseudo-Random Binary Sequence (PRBS) is injected at interface F (or E). Various C/N ratios are established at the input of the receiver under test by addition of Gaussian noise and the BER of the received PRBS is measured at point V (or U) using a BER test set. A test transmitter should be able to generate the 2^{23} - 1 PRBS as defined by Recommendation ITU-T O.151 [i.12]. For the measurement of carrier and noise power, the system bandwidth is defined as $n \times f_{SPACING}$ where n is the number of active carriers i. e. 6 817 or 1 705 carriers and $f_{SPACING}$ is the frequency spacing of the OFDM carriers.

NOTE: The bandwidth in an 8 MHz channel is approx. 7,61 MHz, in a 7 MHz channel system it is 6,66 MHz and 5,71 MHz in a 6 MHz channel.

9.15 BER before Viterbi (inner) decoder

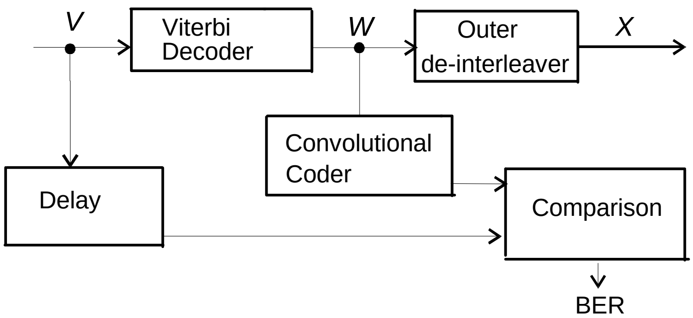
<table>
<tr><td>Purpose</td><td>This measurement gives an in-service indication of the un-coded performance of the transmitter, channel and receiver.</td></tr>
<tr><td>Interface</td><td>V.</td></tr>
<tr><td>Method</td><td>The signal after Viterbi decoding in the test receiver is coded again using the same convolutional coding scheme as in the transmitter in order to produce an estimate of the originally coded data stream. This data stream is compared at bit-level with the signal which is available before Viterbi decoder.
The measurement should be based on at least several hundred bit errors.


Figure 9.4: BER measurement before Viterbi decoding</td></tr>
</table>

9.16 BER before RS (outer) decoder

9.16.0 General

Purpose	The BER is the primary parameter which describes the quality of the digital transmission link.
Interface	W or X
Method	The BER is defined as the ratio between erroneous bits and the total number of transmitted bits. Two alternative methods are available; one for "Out of Service" and a second for "In Service" use. In both cases, the measurement should only be done within the Link Available Time (LAT) as defined in clause 6.2.

9.16.1 Out of Service

The basic principle of this measurement is to generate within the channel encoder a known, fixed, repeating sequence of bits, essentially of a Pseudo-Random nature. In order to do this the data entering the sync-inversion/randomization function is a continuous repetition of one fixed TS packet. This sequence is defined as the *null TS packet* in ISO/IEC 13818-1 [i.1] with all data bytes set to 0x00; i.e. the fixed packet is defined as the four byte sequence 0x47, 0x1F, 0xFF, 0x10, followed by 184 zero bytes (0x00). Ideally this would be available as an encoding system option.

The apparently obvious alternative of injecting a PRBS in the transmitter at the output of the RS encoder is not used because of the requirement to have sync bytes to ensure correct operation of the byte interleaver. Insertion after the byte interleaver is not appropriate because it is not then directly comparable with the in-service measurement.

9.16.2 In Service

The basic assumption made in this measurement method is that the RS check bytes are computed for each link in the transmission chain. Under normal operational circumstances, the RS decoder will correct all errors and produce an error-free TS packet. If there are severe error-bursts, the RS decoding algorithm may be overloaded, and be unable to correct the packet. In this case the transport_error_indicator bit should be set, no other bits in the packet should be changed, and the 16 RS check bytes should be recalculated accordingly before re-transmission on to another link. The BER measured at any point in the transmission chain is then the BER for that particular link only.

The number of erroneous bits within a TS packet will be estimated by comparing the bit pattern of this TS packet before and after RS decoding. If the measured value of BER exceeds 10^{-3} then the measurement should be regarded as unreliable due to the limits of the RS decoding algorithm. Any TS packet that the RS decoder is unable to correct should cause the calculation to be restarted.

9.17 BER after RS (outer) decoder (Bit error count)

Purpose	To gain information about the pattern with which bit errors occur.
Interface	Z
Method	The same principle as used for the "Out of service" measurement of the "BER before the RS decoder" described in clause 9.16.1, with the modification that the result is presented as an error count rather than a ratio. The receiver only has to compare the received TS packets with the Null packets as defined in clause A.1.2. This method is applicable for cases where the BER before RS decoder is lower than approximatively 10^{-3}. This can be used as one parameter for the estimation of the quality of the transmission link as it was defined by the operator, or for localization of specific problems.

9.18 IQ signal analysis

9.18.1 Introduction

The IQ analysis can be applied on single carriers of the OFDM signal as well as on groups of carriers. If groups of carriers are under consideration all received symbols of this group can be superimposed in order to get one common constellation diagram. Since the scattered pilot carriers, the continual pilot carriers and the TPS carriers are transmitted in a different modulation scheme it is recommended to exclude these carriers from the IQ analysis or apply a specific IQ analysis.

Assuming:

- a constellation diagram of M symbol points and K carriers under consideration with $0 < K \le K_{MAX} + 1$ and $K_{MAX} + 1$ is the total number of active OFDM carriers (i.e. 1 705 or 6 817 carriers);
- a measurement sample of N data points, where N is sufficiently larger than $M \times K$ to deliver the wanted measurement accuracy; and

- the co-ordinates of each received data point j being $I_j + \delta I_j$, $Q_j + \delta Q_j$ where I and Q are the co-ordinates of the ideal symbol point and δI and δQ are the offsets forming the error vector of the data point (as long as the respective carrier is a "useful" one).

The following six parameters can be calculated, which give an in-depth analysis of different influences, all deteriorating the signal.

Modulation Error Ratio (MER) and the related Error Vector Magnitude (EVM) are calculated from all N data points without special pre-calculation for the data belonging to the M symbol points.

With the aim of separating individual influences from the received data, for each point i of the M symbol points the mean distance d_i and the distribution σ_i can be calculated from those δI_j, δQ_j belonging to the point i.

From the M values {d1, d2, ... dM} the influences/parameters:

- Origin offset/Carrier suppression (CS);
- Amplitude Imbalance; and
- Quadrature Error (QE)

(only for 2 k modes since the centre carrier needs to carry a complete constellation which is not the case in an 8k system where the centre carrier is a continual pilot) can be extracted and removed from the d_i values, allowing to calculate the Residual Target Error (RTE) with the same algorithm as the System Target Error (STE) from $\{d_1, d_2, ... d_M\}$.

From the statistical distribution of the M clouds the parameters:

- Phase Jitter (PJ); and
- coherent interferer (if it is dominant)

may be extracted. The remaining clouds (after elimination of the above two influences) are assumed to be due to Gaussian noise only and are the basis for calculation of the signal-to-noise ratio. The parameter may include - besides noise - also some other disturbing effects, like small coherent interferers or residual errors from the channel correction.

When using the interfaces S or T filtering of the signal before the interface should be considered.

The parameters Origin offset/Carrier suppression (CS), Amplitude Imbalance (AI) and Quadrature Error (QE) are typical performance parameters of the modulator. The other parameters are also influenced by the transmission system and the receiver/demodulator.

It should be noted that the channel estimation/channel correction mechanism can have an impact on the measurement results. This is particularly true for measurements in the field or under simulated but realistic reception conditions.

For measurements taken at the output of a transmitter this impact of the channel estimation/channel correction mechanism is negligible.

For comparison of measurement results, information on the character of the channel estimation/channel correction mechanism should be provided.

9.18.2 Modulation Error Ratio (MER)

Purpose	To provide a single "figure of merit" analysis of the K carriers.
Interface	S, T and H
Method	The carrier frequency of the OFDM signal and the symbol timing are recovered. Origin offset of the centre carrier (e.g. caused by residual carrier or DC offset), Quadrature Error (QE) and Amplitude Imbalance are not corrected. A time record of N received symbol co-ordinate pairs $(\tilde{I}_j, \tilde{Q}_j)$ is captured. For each received symbol, a decision is made as to which symbol was transmitted. The error vector is defined as the distance from the ideal position of the chosen symbol (the centre of the decision box) to the actual position of the received symbol. This distance can be expressed as a vector $(\delta I_j, \delta Q_j)$. The sum of the squares of the magnitudes of the ideal symbol vectors is divided by the sum of the squares of the magnitudes of the symbol error vectors. The result, expressed as a power ratio in dB, is defined as the MER. $$MER = 10 \times \log_{10} \left\{ \frac{\sum_{j=1}^{N} \left(I_j^2 + Q_j^2 \right)}{\sum_{j=1}^{N} \left(\delta I_j^2 + \delta Q_j^2 \right)} \right\} dB$$ It should be reconsider that MER is just one way of computing a "figure of merit" for a vector modulated signal. Another "figure of merit" calculation is Error Vector Magnitude (EVM) defined in annex C of the present document. It is also shown in annex C that MER and EVM are closely related and that one can generally be computed from the other. MER is the preferred first choice for various reasons itemized in annex C of the present document.

9.18.3 System Target Error (STE) (void)

Figure 9.5: Definition of Target Error Vector (TEV) (void)

9.18.4 Carrier Suppression (CS)

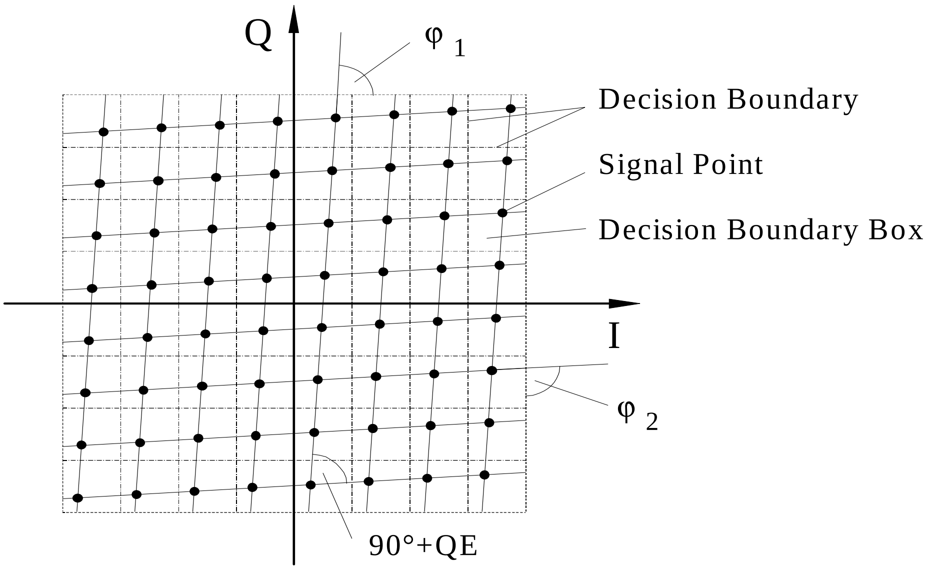
Purpose	A residual carrier is an unwanted coherent signal added to the centre carrier of the OFDM signal. It may have been produced by dc offset voltages of the modulating I and/or Q signal or by crosstalk from the modulating carrier within the modulator.
Interface	S and T.
Method	Search for systematic deviations of all constellation points of the centre carrier and isolate the residual carrier. Calculate the Carrier Suppression (CS) from the formula: $$CS = 10 \times \log_{10} \left(\frac{P_{sig}}{P_{RC}} \right)$$ where P_{RC} is the power of the residual carrier and P_{sig} is the power of the centre carrier of the OFDM signal (without residual carrier).

NOTE: Not applicable for 8k modes (see clause 9.18.1).

9.18.5 Amplitude Imbalance (AI)

Purpose	To separate the QAM distortions resulting from Amplitude Imbalance (AI) of the I and Q signal from all other kind of distortions.
Interface	S and T.
Method	Calculate the I and Q gain values v_I and v_Q from all points in a constellation diagram eliminating all other influences. Calculate Amplitude Imbalance (AI) from v_I and v_Q. NOTE 1: Since the allocation of I and Q to the axis in the complex plane is unambiguous for a DVB-T signal, the parameter AI can convey the information which component dominates. Therefore, this definition differs slightly from the one given in clause 6.9.5. $$AI = \begin{cases} \left(\frac{v_I}{v_Q} - 1\right) \times 100\ \% & if\ v_I \geq v_Q \\ \left(1 - \frac{v_Q}{v_I}\right) \times 100\ \% & if\ v_Q > v_I \end{cases}$$ $$v_I = \frac{1}{M}\sum_{i=1}^{M}\frac{I_i + \left(\bar{d}_i\right)_I}{I_i}$$ $$\left(\bar{d}_i\right)_I = \frac{1}{N}\sum_{j=1}^{N}\delta I_j \qquad \text{(I-component of } d_i \text{ as given in subclause 9.18.3)}$$ $$v_Q = \frac{1}{M}\sum_{i=1}^{M}\frac{Q_i + \left(\bar{d}_i\right)_Q}{Q_i}$$ $$\left(\bar{d}_i\right)_Q = \frac{1}{N}\sum_{j=1}^{N}\delta Q_j \qquad \text{(Q-component of } d_i \text{ as given in subclause 9.18.3)}$$ $$\left(\bar{d}_i\right)_I + \left(\bar{d}_i\right)_Q = \bar{d}_i$$ NOTE 2: Not applicable for 8k modes (see clause 9.18.1).

9.18.6 Quadrature Error (QE)

Purpose	The phases of the two carriers feeding the I and Q modulators have to be orthogonal. If their phase difference is not 90 a typical distortion of the constellation diagram results. *It is assumed that the value derived from the centre carrier is representative for the whole signal.*
Interface	S and T.
Method	Search for the constellation diagram error shown in figure 9.6 and calculate the value of the phase difference $\Delta\varphi = \varphi_1 - \varphi_2$ after having eliminated all other influences and convert this into degrees: $QE = \frac{180°}{\pi} \times (\varphi_1 - \varphi_2)\ \ [°]$  **Figure 9.6: Distortion of constellation diagram resulting from I/Q Quadrature Error (QE)**

NOTE: Not applicable for 8k modes (see clause 9.18.1).

9.18.7 Phase Jitter (PJ)

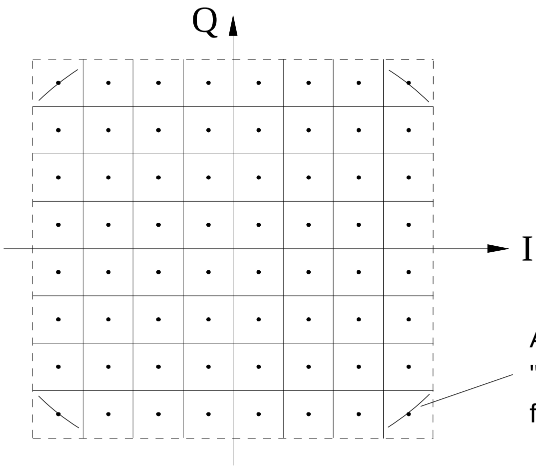
<table>
<tr><td>Purpose</td><td>The PJ of an oscillator is due to fluctuations of its phase or frequency. Using such an oscillator to modulate a digital signal results in a sampling uncertainty in the receiver, because the carrier regeneration cannot follow the phase fluctuations.
The signal points are arranged along a curved line crossing the centre of each decision boundary box as shown in figure 9.7 for the four "Corner Decision Boundary Boxes".


Figure 9.7: Position of "Arc section" in the constellation diagram to define PJ (example: 64-QAM)</td></tr>
<tr><td>Interface</td><td>S and T.</td></tr>
<tr><td>Method</td><td>Phase Jitter can be calculated theoretically using the following algorithm:
1) Calculate the angle between the I-axis of the constellation and the vector to the received symbol (I_{rcvd}, Q_{rcvd}):
$$\phi_1 = \arctan\frac{Q_{rcvd}}{I_{rcvd}}$$
2) Calculate the angle between the I-axis of the constellation and the vector to the corresponding ideal symbol (I_{ideal}, Q_{ideal}):
$$\phi_2 = \arctan\frac{Q_{ideal}}{I_{ideal}}$$
Phi 2 instead of Phi 1
3) Calculate the error angle:
$$\phi_E = \phi_1 - \phi_2$$
4) From these N error angles calculate the RMS phase jitter:
$$PJ = \sqrt{\frac{1}{N}\sum_{i=1}^{N}\phi_{E_i}^2 - \frac{1}{N^2}\left(\sum_{i=1}^{N}\phi_{E_i}\right)^2}$$
However, the following method may be more practical:
The first approximation of the "Arc Section" of a "Corner Decision Boundary Box" is a straight line parallel to the diagonal of the "Decision Boundary Box". Additionally the curvature of the Phase Jitter (PJ) trace has to be taken into account when calculating the standard deviation of the PJ. The mean value of the PJ is calculated in degrees.
$$PJ = \frac{180^\circ}{\pi} \times \arcsin\left(\frac{\sigma_{PJ}}{\sqrt{2}\times\left(\sqrt{M}-1\right)\times d}\right)[^\circ]$$
where M = Order of QAM
and 2d = Distance between two successive boundary lines
Within the argument of the arc sine function, the standard deviation of the Phase Jitter is referenced to the distance from the centre of the "Corner Decision Boundary Box" to the centre point of the QAM signal.</td></tr>
</table>

9.19 Overall signal delay

Purpose	To measure and adjust the signal delay of an OFDM transmitter to a given value so that the transmitters in an SFN can be synchronized.
Interface	A, M.
Method	(a) The total delay between the MPEG TS input of the transmitter under test and the MPEG TS output of a test receiver is established by measuring the time delay required to match the input and output data patterns. If the delay of the test receiver is known then the transmitter signal delay can be derived. Alternatively, the delay of the test receiver could be expressed relative to the delay of a reference receiver. This would avoid the need to measure the absolute delay of any receiver. (b) A more direct method may be to define a transmitter test mode in which the occurrence of a Mega-frame Initialization Packet (MIP) at the MPEG TS input causes a trigger pulse (see ETSI TS 101 191 [i.14]). The trigger pulse is made available for connection to an oscilloscope and also used to "arm" the modulator. At the start of the next mega-frame the modulator transmits a null symbol (or a defined pulse in the time domain) rather than the normal data. The delay between the trigger pulse and the RF null (or pulse) is measured. (c) The delay of a transmitter could be expressed relative to the delay of a reference transmitter. For the measurement a reduced amplitude sample is taken from both transmitters and adjusted to have similar level (< 3 dB difference), the samples are combined in a RF linear adder and the output is fed to a spectrum analyser. Typically the spectrum formed will have lobes due to the difference of delays in the two transmitters. The inverse of the frequency width of the lobes represents the relative delay between the transmitters. Two drawbacks has to be taken in account: 1) the delay is absolute, that is, it gives no indication of which transmitter has the longer delay; 2) the accuracy is related to the ability of identifying the minimal values of the lobes and the accuracy of the measurement. NOTE 1: The delay of a transmitter may be considered as the addition of various parts including the physical delays of the analogue part of the OFDM signal, including the path length to the antenna. Also the buffers used for signal conditioning (TS bitrate adaptation to the sampling frequency of the transmitter) and other intermediate buffers in the OFDM spectrum calculation (IFFT) may differ from manufacturer to manufacturer. NOTE 2: In cases of single frequency networks, the SFN adapter at the transmitter site may be considered as integral part of the modulator transmitter. It may calculate the delay, from the value of the STS (Synchronization Time Stamp) to the 1 pps used as reference, in different way from manufacturer to manufacturer and add differences in the delays that have to be included in the measurement result. It is recommended to use a test Transport Stream with embedded MIP data, and real-time calculation of the STS. See clause E.16 for test set-up, measurement description and example of results.

9.20 SFN synchronization

9.20.1 MIP_timing_error

Purpose	A necessary precondition for SFN synchronization is that the Synchronization Time Stamp (STS) values inserted in the Mega-frame Initialization Packet (MIP) are correct. This test checks that successive STS values are self-consistent. See ETSI TS 101 191 [i.14].
Interface	A, Z (especially Transport Stream between the "SFN adapter" and "SYNC system" as defined in [i.14]).
Method	Locate the MIP in three successive mega-frames numbered M, M+1 and M+2. Extract the synchronization_time_stamp field from each MIP (STS_M, STS_{M+1} and STS_{M+2}). In general, the difference between any two consecutive STS values will be the duration of one mega-frame minus some multiple (including zero) of the time between GPS pulses. Even without knowing the precise duration of the mega-frame, the duration is constant and the following can be derived: $STS_{M+2} - STS_{M+1} = STS_{M+1} - STS_M + nT$ where T is 1s and n is any integer. Calculate nT from the above formula and check it is an integral number of seconds to within a user defined accuracy. This test can be performed continually on each successive set of 3 mega-frames, {M+1, M+2, M+3}, {M+2, M+3, M+4} etc. The test result should be discarded if the mega-frame size changes over the set of three mega-frames. NOTE: The mega-frame size changes, for example, with the change of the DVB-T transmission mode. This would normally result in a resynchronization.

NOTE: Figure 9.8 is an illustration of the timing relationship between mega-frames and the GPS one second pulses. This shows how the synchronization_time_stamp (STS) is calculated.

Consider STS_{M+1} and STS_{M+2}. In this case it is quite clear that:

$$STS_{M+2} - STS_{M+1} = \text{duration of one mega-frame}$$

In the case of STS_M and STS_{M+1}, a 1s pulse has passed by and the equivalent equation is:

$$(STS_{M+1} + 1) - STS_M = \text{duration of one mega-frame}$$

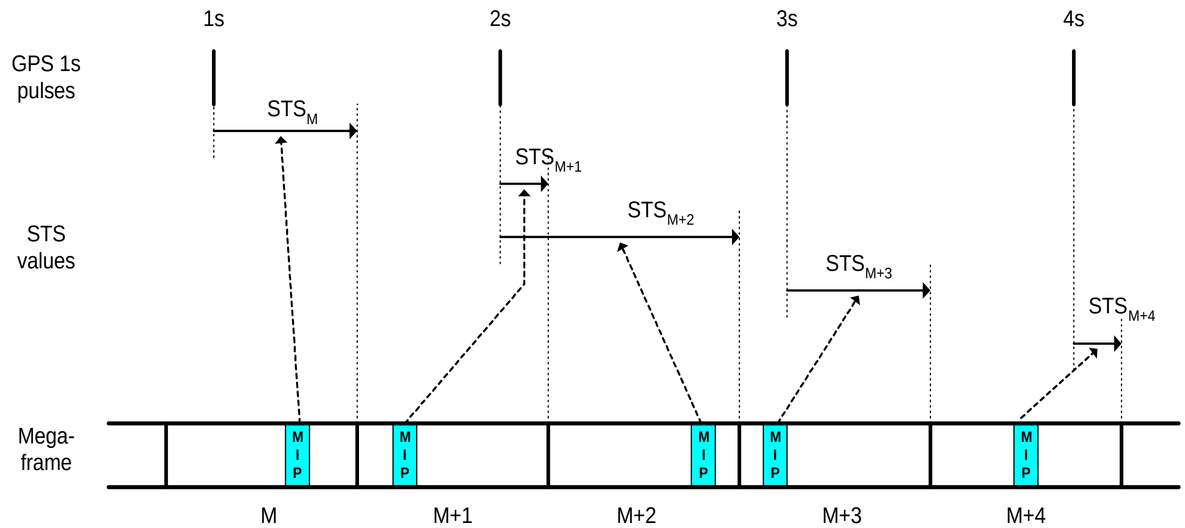


Figure 9.8: Megaframe/GPS pulse timing relationship

9.20.2 MIP_structure_error

Purpose	This test verifies that the syntax of the MIP complies with the specification in ETSI TS 101 191 [i.14].
Interface	A, Z
Method	For each transport packet carried on PID 0x15 in the transport stream, the following checks are performed: The transport_packet_header should comply with ETSI TS 101 191 [i.14] clause 6, table 1, and ISO/IEC 13818-1 [i.1] clause 2.4.3.2, tables 2 and 3. All length fields should be consistent to provide a proper length packet. This includes section_length (which also should not exceed 182), individual_addressing_length (which should match the length of the loops for each transmitter), function_loop_length (which should match the sum of the size of each of the functions), function_length (which should match the proper length of the function based upon the function tag). The synchronization_time_stamp and the maximum_delay should be in the range of 0x0 to 0x98967F. The CRC_32 field should match the CRC calculated for the MIP data.

9.20.3 MIP_presence_error

Purpose	This test verifies that the MIP is inserted into the transport stream only once per mega-frame.
Interface	A, Z
Method	The following checks are performed: Extra MIP – For every MIP_N (where N > 1), signal an error if it arrives within the number of packets indicated by the pointer field of MIP_{N-1}. Missing MIP - For each MIP received, calculate the mega-frame size from the parameters in the tps_mip. The latest two values of the mega-frame size are stored. After every MIP_N is received (where N > 1), signal an error if a MIP_{N+1} is not received before K + R packets are received after MIP_N, where K is the pointer value of MIP_N and R is mega-frame size in packets from the previous MIP_{N-1}.

9.20.4 MIP_pointer_error

Purpose	The MIP insertion can be at any location in the mega-frame. If the insertion is periodic as defined in the MIP, the MIP location in the mega-frame is constant over time. The MIP can be used to determine the mega-frame size and where each mega-frame starts and ends in the transport stream thanks to the pointer field verified by this test.
Interface	A, Z
Method	For each MIP received, calculate the mega-frame size from the parameters in the tps_mip. The latest three values of the mega-frame size are stored. For every MIP_N that is received (where N > 2), signal an error if the pointer value (P_N) of MIP_N does not hold in the following equation: $P_N = P_{N-1} + MF_{N-2} - (i_N - i_{N-1})$ Where MF_{N-2} is the size of the N^{th} mega-frame in packets but is calculated from MIP_{N-2}, and i_N is the packet index for MIP_N.

9.20.5 MIP_periodicity_error

Purpose	In the case of a periodic MIP insertion (as defined in ETSI TS 101 191 [i.14] clauses 5 and 6), the pointer value should remain constant, as well as the number of packets between each MIP.
Interface	A, Z
Method	The following checks are performed: Compare the current pointer field in MIP_N with the pointer field in the MIP_{N-1}. It is an error if they are different, unless the mega-frame size changed between N and N-1. The number of packets between each MIP (i_N - i_{N-1}) should also be constant unless the mega-frame size changes.

9.20.6 MIP_ts_rate_error

Purpose	In a SFN network the modulator settings are transmitted by the tps_mip (see ETSI TS 101 191 [i.14] clause 6, table 3). These settings determine the transmission mode and in this way the bit rate of the Transport Stream. This test verifies that the actual Transport Stream data rate is consistent with the DVB-T mode defined by the tps_mip.
Interface	A, Z
Method	For each MIP received, calculate the data rate of the transmission mode - given by tps_mip setting and compare it with the actual data rate of the Transport Stream. Signal an error if the following equation is correct: Max_deviation ≤ $\|$ TS_data_rate - [(IFFT_clock_freq × tpl /204 × c × m × (uc/tc))/(1 + g)] $\|$ Where: • Max_deviation e.g. 10 kb/s; maximum deviation between actual TS_data_rate and data rate of the transmission mode given by tps_mip. → The value results from the smallest difference of TS data rates which can be determined by two correct tps_mip settings for different modes. • TS_data_rate actual data rate of the Transport Stream → measured by a test instrument according to clause 5.3.3.2. • IFFT_clock_freq 64/7 MHz (for 8 MHz channel bandwidth), 64/8 MHz (for 7 MHz channel bandwidth) 48/7 MHz (for 6 MHz channel bandwidth) → given by tps_mip P_{12} and P_{13} • tpl transport packet length 188 or 204 byte • c code rate $^1/_2$, $^2/_3$, $^3/_4$, $^5/_6$ or $^7/_8$ → given by tps_mip P_5,P_6 and P_7 • m 2 (for QPSK), 4 (for 16 QAM) or 6 (for 64 QAM) → given by tps_mip P_0 and P_1 • uc useful_carriers 1512 (for 2k), 6 048 (for 8k) → given by tps_mip P_{10}, P_{11} (see note) • tc total_carriers 2 048 (for 2k), 8 192 (for 8k) → given by tps_mip P_{10}, P_{11} (see note) • g guard interval $^1/_4$, $^1/_8$, $^1/_{16}$ or $^1/_{32}$ → given by tps_mip P_8, P_9 NOTE: The term (uc/tc) can be replaced by a constant value since $uc_{2k}/tc_{2k} = uc_{8k}/tc_{8k}$.

9.21 System Error Performance

Purpose:	The System Error Performance describes the performance of the digital transmission from the input of the MPEG-2 TS signal into the DVB Baseline system to the MPEG-2 TS output of this Baseline system.
Interfaces:	A, Z, M: with reference receiver (e.g. Transmitter measurement). N: with reference receiver (e.g. coverage measurements).
Method:	The measurement of System Error Performance is based on a subset of the error events defined in clause 5.4: • Errored Second (ES) or Errored Time Interval (ETI), • Severely Errored Second (SES) or Severely Errored Time Interval (SETI). The used time interval T for identification of these events depends on the aim of the measurement. Time intervals longer or shorter than 1 second may be considered appropriate in certain circumstances. **Evaluation of Error Performance Parameters** Error performance should only be evaluated whilst the transmission is in the available state (see also clause 6.1). To evaluate error performance parameters from events, a certain measurement interval (MI) has to be used. This measurement interval depends on the specific aim of the measurement. Possible measurement intervals corresponding to special applications are proposed in table 9.4. In general the error performance is the ratio of number of true events to the total number of time intervals T during the measurement interval. Consequently derived performance parameters are: • Errored Second Ratio (ESR) or Errored Time Interval Ratio (ETIR); • Severely Errored Second Ratio (SESR) or Severely Errored Time Interval Ratio (SETIR).

Table 9.4: Examples of Measurement Intervals MI

Length of Measurement Interval (MI)	Application
5 s	- applicable for analysis of mobile reception
20 s	- Coverage Check - recommended minimum measurement interval for receiver comparison
5 minutes	- possible resolution for 1 hour analysis.
1 hour	- possible resolution for daily fluctuations analysis

10 Recommendations for the measurement of delays in DVB systems

Void.

11 Measurements for the second generation terrestrial (DVB-T2) system

11.1 Introduction

The DVB-T2 system as it is addressed in the following clauses, spans from the input Interface A to the output Interface D. Both these interfaces carry MPEG2 Transport Streams ("TS").

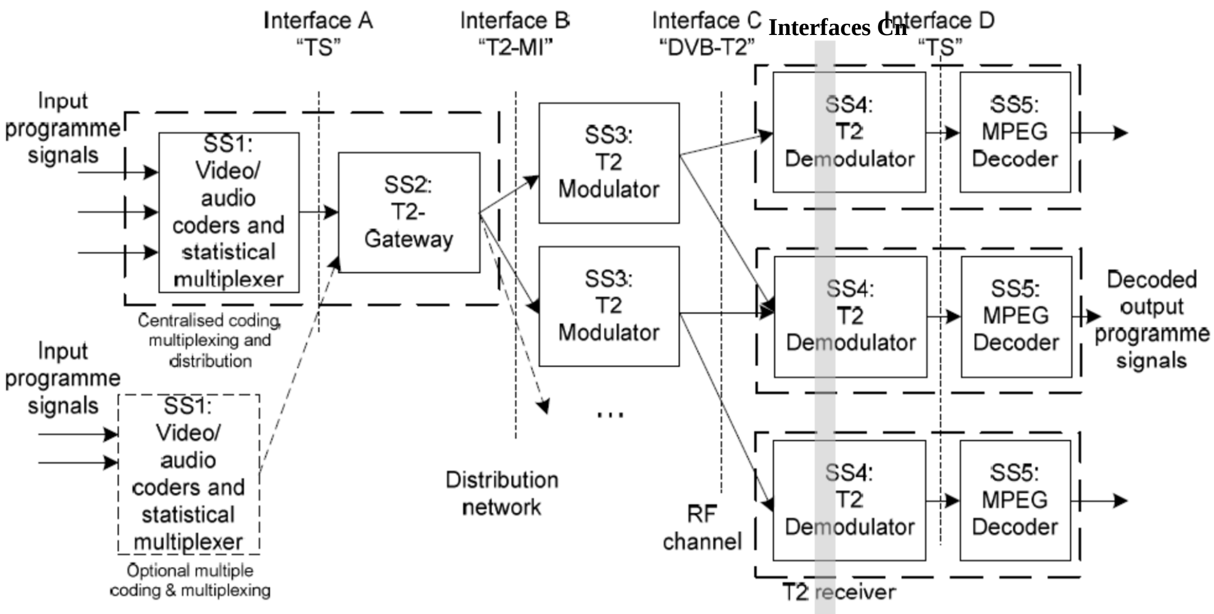


Figure 11.1: Block diagram of a typical DVB-T2 chain [i.27]

The following clauses specify a number of tests and measurements at the interfaces A, B, C, Cn and D.

For measurements at Interface A and D, see clause 5 for the parameters measured at the other interfaces see clauses 11.2 and 11.3. The Interfaces Cn are introduced in addition to the interfaces defined in ETSI EN 302 755 [i.27] to accommodate all specified measurements for DVB-T2.

With regard to the signalling, it is recommended that a measurement instrument should display the signalled information as readable text and abbreviations.

11.2 Measurements at the DVB-T2 Modulator Interface (T2-MI)

11.2.1 Introduction

The DVB-T2 Modulator Interface specification ETSI TS 102 773 [i.24] defines the format of T2-MI packets. The T2-MI format enables the operation of Single Frequency Networks SFNs by taking the scheduling decisions at the central point of the T2 gateway. The output from the T2 Gateway is forwarded to all modulators in the SFN so that all modulators receive identical information.

For more details refer to the T2-MI standard [i.24]

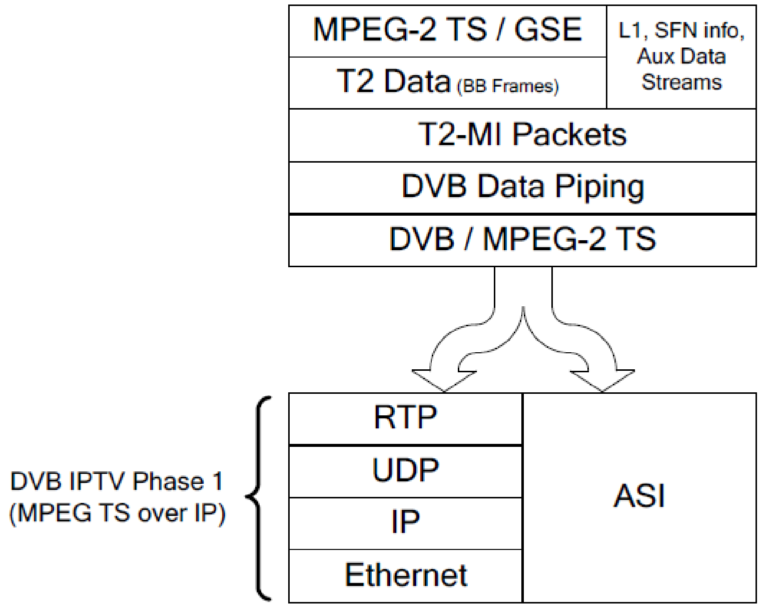


Figure 11.2: The T2-MI protocol stack [i.24]

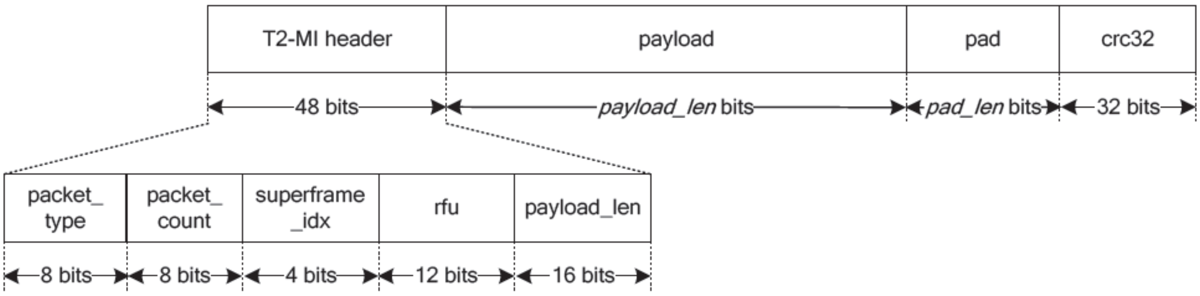


Figure 11.3 T2-MI packet format [i.24]

11.2.2 Measurements of the syntax of T2 MI packets

11.2.2.1 T2MI_packet_type_error_1

Purpose	Two of the various packet types are mandatory for each T2 frame: 10_{16} (L1-current data), 20_{16} (DVB-T2 Timestamp). If L1 repetition, in-band signalling (IBS) or Time-Frequency-Slicing (TFS) is indicated in the L1-current data, an L1-future packet (11_{16}) should also be present. If any of these packet_type is not present in each T2 frame, a T2MI_packet_type_error_1 is signalled.
Interface	Interface B "T2-MI"
Method	Comparison of the decoded packet_type value with the list of mandatory values.
Reference	Clauses 5.1 and 5.4 of ETSI TS 102 773 [i.24]

11.2.2.2 T2MI_packet_type_error_2

Purpose	The number of BB-frames (packet_type 00_{16}) relating to a PLP in a given T2 frame should match the value of PLP_NUM_BLOCKS signalled in the dynamic signalling of both the L1-current and L1-future (when present). The signalled values of frame_idx and superframe_idx for BB-frame packets should be consistent with the time interleaver parameters specified in the configurable signalling of the L1-current. NOTE: • BB-frames are only mandatory for a given PLP if they are signalled in the L1-current data. • In a case where the Interleaving Frame spans more than one T2-frame (P_I>1), there is a possibility that there will be no BB frames present with certain frame_idx values since the frame_idx always refers to the first T2 frame of the Interleaving Frame. • This could also occur if the bit-rate for a given PLP falls to zero in a given T2 frame.
Interface	Interface B "T2-MI"
Method	Comparison of the number of BB frames and their values of frame_idx and superframe_idx with the information from the L1 signalling.
Reference	Clauses 5.1 and 5.4 of ETSI TS 102 773 [i.24]

11.2.2.3 T2MI_packet_count_error

Purpose	This error indicates a discontinuity of T2-MI packets
Interface	Interface B "T2-MI"
Method	Comparison of the decoded packet_count value of the received T2-MI packet with the packet_count value of the previous packet. The case of receiving the first packet of a transmission, for which no specific packet_count value is required, needs to be considered.
Reference	Clause 5.1 of ETSI TS 102 773 [i.24]

11.2.2.4 T2MI_CRC_error

Purpose	The CRC32 check indicates if the content of the respective T2-MI packet is corrupted. It is calculated across all other bits in the packet (both header and payload plus any padding).
Interface	Interface B "T2-MI"
Method	According to annex A of ETSI TS 102 773 [i.24]
Reference	Clause 5.1 of ETSI TS 102 773 [i.24]

11.2.2.5 T2MI_payload_error

Purpose	The T2MI_payload_error is signalled if the decoded plp_id in T2-MI packets with packet_type 0016 is not included in the list (L1 post_signalling/configurable) of plp_id for the T2-MI signal.
Interface	Interface B "T2-MI"
Method	Comparison of the decoded plp_id of T2-MI packets with packet_type 0016 with list of possible values.
Reference	Clause 5.2.1 of ETSI TS 102 773 [i.24]

11.2.2.6 T2MI_plp_num_blocks_error

Purpose	The number of FEC blocks in an Interleaving Frame for a PLP as signalled in the Dynamic L1-post signalling should be consistent with the number of BB frame packets.
Interface	Interface B "T2-MI"
Method	This error indication is set if the number of received BB frame packets does not match the signalled value.
Reference	Clause 7.2.3.2 of ETSI EN 302 755 [i.27] and clause 5.2.1 of ETSI TS 102 773 [i.24]

11.2.2.7 T2MI_transmission_order_error

Purpose	The T2MI_transmission_order_error is signalled if the packet_types are in a wrong ordering and position inside a T2 frame.
Interface	Interface B "T2-MI"
Method	The required ordering is DVB-T2 Timestamp (20_{16}) -> P2 bias balancing cells (12_{16}, if present) -> L1_Current (10_{16}) -> L1_future (11_{16}, if present) -> change of frame_idx
Reference	Clause 5.4 of ETSI TS 102 773 [i.24]

11.2.2.8 T2MI_DVB-T2_Timestamp_error

Purpose	The T2MI_DVB-T2_Timestamp_error signals a wrong timestamp inside a superframe
Interface	Interface B "T2-MI"
Method	The T2MI_DVB-T2_Timestamp_error is signalled if at least one T2_timestamp has a different value than the other DVB-T2 timestamp inside a single superframe_(superframe_idx).
Reference	Clause 5.2.7.1 of ETSI TS 102 773 [i.24]

11.2.2.9 T2MI_DVB-T2_Timestamp_discontinuity

Purpose	The T2MI_DVB-T2_Timestamp_discontinuity signals a non-increasing timestamp (not relevant with null timestamps)
Interface	Interface B "T2-MI"
Method	The T2-MI timestamps are compared for subsequent superframes and the error is indicated if the difference is not equal to the duration of the superframe. NOTE: Timestamp_discontinuity should take into account a time window of what should be the next time stamp and the received timestamp.
Reference	Clause 5.7 of ETSI TS 102 773 [i.24]

11.2.2.10 T2MI_T2_frame_length_error

Purpose	The T2 frame length derived from parameters signalled in L1 is not longer than 250 ms.
Interface	Interface B "T2-MI"
Method	The error is indicated if the T2 frame length derived from L1 signalling parameters is over 250 ms.
Reference	Clause 5.2.4 of ETSI TS 102 773 [i.24] and clause 7.2 of ETSI EN 302 755 [i.27]

11.2.3 Checks on the T2-MI MIP (Modulator Information Packet)

11.2.3.1 T2MI_MIP_timestamp_error

Purpose	The value of the timestamp of the T2-MI MIP (t2_timestamp_mip) with PID = 15_{16} should not be lower than the value of the DVB-T2 timestamp.
Interface	Interface B "T2-MI"
Method	The error is indicated if the timestamp of the T2-MI MIP has a lower value than the value of the DVB-T2 timestamp.
Reference	Annex B of ETSI TS 102 773 [i.24]

11.2.3.2 T2MI_MIP_individual_addressing_error

Purpose	The consistency of the data contained in the individual_addressing_byte is checked against the bytes of the individual_addressing_data field of a T2-MI packet of type 21_{16}.
Interface	Interface B "T2-MI"
Method	The error is indicated if the comparison shows inconsistencies.
Reference	Clauses B.2.1 and 5.2.8 of ETSI TS 102 773 [i.24]

11.2.3.3 T2MI_MIP_continuity_error

Purpose	The insertion of at least one complete T2-MIP within a T2 superframe is required.
Interface	Interface B "T2-MI"
Method	The error is indicated if the usage of MIP is signalled, and a superframe without a complete T2-MIP packet is found.
Reference	Clause B.2.2 of ETSI TS 102 773 [i.24]

11.2.3.4 T2MI_MIP_CRC_error

Purpose	The CRC32 bits of the T2-MIP are checked to establish that the T2-MIP packet is uncorrupted.
Interface	Interface B "T2-MI"
Method	The error is indicated if the CRC check shows that the T2-MIP packet is corrupted.
Reference	Clause B.2.1 of ETSI TS 102 773 [i.24]

11.2.4 Check on consistency of T2-MI signalling information

11.2.4.1 T2MI_bandwidth_consistency_error

Purpose	The bandwidth signalled in the T2-MI DVB-T2 Timestamp and the determining parameters signalled in L1 (i.e. fft mode, guard interval, pilot pattern, number of OFDM data symbols and PLP specific code rate, modulation, fec type, num_blocks, baseband mode, issy information, null packet deletion, in-band signalling, num other plp inband) should be such that the stream can be transmitted in the pertaining channel.
Interface	Interface B "T2-MI"
Method	The error is indicated if the maximum possible bit rate which is calculated from the T2-MI DVB-T2 Timestamp and the determining L1 parameters is lower than the bit rate of the stream encapsulated in the T2-MI packets.
Reference	Clause 5.2.7 of ETSI TS 102 773 [i.24] and clause 7.2 of ETSI EN 302 755 [i.27]

11.2.4.2 T2MI_DVB-T2_Timestamp_leap_second_error

Purpose	The T2MI_DVB-T2_Timestamp_leap_second_error signals an error in leap second value signalled in utco field in T2 Timestamp.
Interface	Interface B "T2-MI"
Method	Leap second value in T2-MI timestamp is compared to known leap second value as published by IERS (*International Earth Rotation Service)*.
Reference	Clause 5.2.7 of ETSI TS 102 773 [i.24]

11.2.5 Measurements at T2-MI transport layer

11.2.5.1 Encapsulation of T2-MI packets into MPEG-2 TS (ASI) streams

11.2.5.1.0 General

T2-MI packets are encapsulated into 188-bytes MPEG-2 TS packets according to "*Data Piping*" mechanism (ETSI EN 301 192 [i.33], clause 4). This allows to reuse existing telecom links based on ASI.

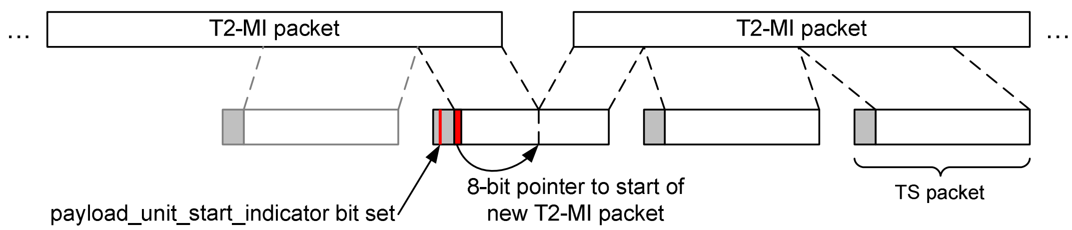


Figure 11.4: Encapsulation of T2-MI packets into MPEG-2 TS (ASI)

The addition of a minimum of PSI (PAT and PMT) is suggested, to help detecting alarms on the ASI link.

The "classic" Reed-Solomon (188,204) FEC can be (optionally) adopted.

11.2.5.1.1 Parameters relevant to transport of T2-MI packets over MPEG-2 TS

Measurements of the MPEG-2 TS stream carrying T2-MI packets can be done according to Clause 5 of the present document.

However, the TS in this case is only used as a container for the T2-MI packets according to data piping mechanism, and therefore only a subset of the parameters are relevant for the measurements.

11.2.5.1.2 Informative parameters

Parameter	Notes	Reference
Total TS bit-rate	The total bit-rate of the TS carrying the T2-MI data includes the TS header and PAT, PMT, stuffing, etc. The default bit-rate measurement profile should be MGB1. For special cases, MGB5 with a user-defined measurement interval should be used.	Clause 5.3.3.2
T2-MI bit-rate	Bit-rate corresponding to the PID carrying the T2-MI packets (data piping). The default bit-rate measurement profile should be the net TS bit-rate without header and without Adaptation Field.	
RS(188,204) FEC	Present/Absent	

11.2.5.1.3 Integrity parameters

Subset of relevant 1st priority parameters, with reference to clause 5.2.1:

No.	Indicator	Notes
1.1	TS_sync_loss	
1.2	Sync_byte_error	
1.3/1.3a	PAT_error	PAT is suggested but not mandatory in ETSI TS 102 773 [i.24]
1.4	Continuity_count_error	
1.5/1.5a	PMT_error	PMT is suggested but not mandatory in ETSI TS 102 773 [i.24]
1.6	PID_error	Since the T2-MI over TS is considered as a data service, a user-defined repetition rate of 5 s or more may be suitable.

Subset of relevant 2nd priority parameters, with reference to clause 5.2.2:

No.	Indicator	Notes
2.1	Transport_error	It should be possible to disable the indication of the Transport_error if the T2-MI transport system does not support this function.
2.3	PCR_error	If signalled according to ETSI TS 102 773 [i.24], annex G
2.4	PCR_accuracy_error	If signalled according to ETSI TS 102 773 [i.24], annex G

Subset of relevant 3rd priority parameters, with reference to clause 5.2.3:

No.	Indicator	Notes
3.4/3.4a	Unreferenced_PID	Relevant if PMT is present

11.2.5.2 Encapsulation of T2-MI packets into IP streams

11.2.5.2.1 Parameters relevant to transport of T2-MI packets over IP

11.2.5.2.1.0 General

MPEG-2 TS packets carrying the T2-MI data are encapsulated into IP using the RTP/UDP/IP protocol stack, according to DVB-IPTV standard (ETSI TS 102 034 [i.25]).

A number of TS packets can be allocated per IP packet, limited by the maximum size of the IP datagram. Care should be taken not to exceed the underlying maximum transmission unit (MTU) of the network, in order to avoid fragmentation.

11.2.5.2.1.1 Informative parameters

Parameter	Explanation and notes	Reference
Source IP address	Source address of the IP stream carrying the T2-MI packets	IETF RFC 791 [i.34]
Source IP port	Source port of the UDP stream carrying the T2-MI packets	IETF RFC 768 [i.35]
Destination IP address	Destination address of the IP stream carrying the T2-MI packets. It is normally a multicast group	IETF RFC 791 [i.34], IETF RFC 3171 [i.36]
Destination IP port	Destination port of the UDP stream carrying the T2-MI packets	IETF RFC 768 [i.35]
IP stream bit-rate	Total bit-rate associated to the IP stream carrying the TS with T2-MI packets	
No. of TS packets per IP frame	Number of TS packets encapsulated in one IP frame	Clause 7.1.1 of ETSI TS 102 034 [i.25]
FEC type	Type of FEC applied to protect the T2-MI stream over the IP linik. Possible values: 'None', 'SMPTE 2022-1', 'Raptor'	Annex E of ETSI TS 102 034 [i.25]
SMPTE FEC rows (L)	Number of rows of the SMPTE 2022-1 FEC matrix (if FEC is applied)	Clause 8.1 of SMPTE 2022-1 [i.26]
SMPTE FEC columns (D)	Number of columns of the SMPTE 2022-1 FEC matrix (if FEC is applied)	Clause 8.1 of SMPTE 2022-1 [i.26]
SMPTE FEC IP port	Destination port of the UDP stream carrying the SMPTE FEC packets (if FEC is applied). This port should be equal to destination IP port + 2	IETF RFC 768 [i.35], Clause 8.1 of SMPTE 2022-1 [i.26]
SMPTE FEC bit-rate	rate associated to the stream carrying SMPTE 2022-1 packets	

NOTE: Typical value for number of TS packets encapsulated in one IP frame over Ethernet is 7. In ISO/IEC 13818-1 [i.1] it is not required that this number keeps constant along the stream; however, a variation of this value is not compatible with SMPTE 2022-1 [i.26] FEC (and in fact such variation is not allowed by SMPTE 2022-2 [i.43]).

11.2.5.2.1.2 Integrity parameters

11.2.5.2.1.2.0 General

- Media Loss Rate (MLR) or Lost IP frames
- Corrected IP frames

- Delay Factor

11.2.5.2.1.2.1 Media Delivery Index - Media Loss Rate (MDI-MLR)

Purpose	Number of IP packets lost per second, after error recovery mechanisms, if any. This parameter is part of the MDI (Media Delivery Index) parameter. Error if > 0.
Interface	Interface B "T2-MI" – IP transport layer
Method	The Continuity Counter field of the inner MPEG-2 TS stream can be used. Under error free conditions, the sequence numbers increment. Counting any missing sequence numbers every second will produce the packets lost per second. Alternatively, the RTP layer information can be used as a supplement for the UDP protocol. RTP also uses sequence numbers and this number also increments with every packet. The same method as above, of counting missing sequence numbers may be used.
Reference	Clause 3.2 of IETF RFC 4445 [i.37]

11.2.5.2.1.2.2 Lost IP frames

Purpose	Alternative option to represent packet loss with respect to MLR. Cumulative number of IP packets lost per second, after error recovery mechanisms, if any. Error if > 0.
Interface	Interface B "T2-MI" – IP transport layer
Method	The Continuity Counter field of the inner MPEG-2 TS stream can be used. Under error free conditions, the sequence numbers increment. Counting any missing sequence numbers every second will produce the packets lost per second. Alternatively, the RTP layer information can be used as a supplement for the UDP protocol. RTP also uses sequence numbers and this number also increments with every packet. The same method as above, of counting missing sequence numbers may be used.
Reference	Clause 3.2 of IETF RFC 4445 [i.37]

11.2.5.2.1.2.3 Corrected IP frames

Purpose	The value of Lost IP frames + Corrected IP frames gives an indication of the capabilities of the FEC mechanism.
Interface	Interface B "T2-MI" – IP transport layer
Method	In case of FEC, number of IP packets corrected by the FEC mechanism.
Reference	

11.2.5.2.1.2.4 Media Delivery Index - Delay Factor (MDI-DF)

Purpose	Maximum difference, observed at the end of each media stream packet, between the arrival of media data and the drain of media data, over a calculation interval (typically 1 s). This assumes the drain rate is the nominal constant traffic rate for constant bit rate. The DF gives a good approximation of the needed buffer at receiving side to compensate for network jitter cumulated over the calculation interval. This parameter is part of the Media Delivery Index (MDI) parameter.
Interface	Interface B "T2-MI" – IP transport layer
Method	Consider a virtual buffer VB used to buffer received packets of a stream. When a packet P(i) arrives during a calculation interval, compute two VB values, VB(i,pre) and VB(i,post), defined as: • VB(i,pre) = sum (Sj) - MR × Ti; where j=1..i-1 • VB(i,post) = VB(i,pre) + Si where Sj is the media payload size of the jth packet, Ti is the relative time at which packet i arrives in the interval, and MR is the nominal media rate. VB(i,pre) is the Virtual Buffer size just before the arrival of P(i). VB(i,post) is the Virtual Buffer size just after the arrival of P(i). The initial condition of VB(0) = 0 is used at the beginning of each measurement interval. A measurement interval is defined from just after the time of arrival of the last packet during a nominal period (typically 1 second) to the time just after the arrival of the last packet of the next nominal period. During a measurement interval, if k packets are received, then there are 2 × k + 1 VB values used in deriving VB(max) and VB(min). After determining VB(max) and VB(min) from the 2k + 1 VB samples, DF for the measurement interval is computed and displayed as: • DF = [VB(max) - VB(min)]/MR
Reference	Clause 3.1 of IETF RFC 4445 [i.37]

11.3 Measurements for DVB-T2 baseline system

11.3.0 General

This clause lists a number of measurements for the DVB-T2 baseline system. The parameters are mainly measured at interface C "DVB-T2" or at interface C1 "T2-IQ" as in figure 11.1.

A list of the main application area of the DVB-T2 measurement parameters described in this clause is given in table 11.1.

The measurements in clause 6.1 "System availability" (Interface D in a DVB-T2 system, figures 9.2) and clause 6.2 "Link availability" (Interface C5 in a DVB-T2 system, figure 9.2) are also applicable.

Table 11.1: DVB-T2 measurement parameters and their application area

Measurement parameter	Transmitter	Network	Receiver	In-service measure-ment
11.3.1 RF measurements	X			X
11.3.1.1 RF frequency accuracy	X	X		X
11.3.1.2 RF occupied bandwidth	X		X	X
11.3.2 Selectivity			X	
11.3.3 AFC capture range			X	
11.3.4 Phase noise of Local Oscillators (LO)	X		X	
11.3.5 RF/IF signal power	X	X	X	X
11.3.6 MISO Group Power Ratio	X	X	X	X
11.3.7 Noise Power	X	X	X	X
11.3.8 RF and IF spectrum	X		X	X
11.3.9 Receiver sensitivity/dynamic range for a Gaussian channel			X	
11.3.10 Linearity characterization (shoulder attenuation)	X			X
11.3.11 Power efficiency	X			X
11.3.12 PAPR effect	X			
11.3.13 P1 Symbol Error Rate			X	X
11.3.14 BER before LDPC (inner) decoder			X	X
11.3.15 Number of LDPC iterations			X	X

Measurement parameter	Transmitter	Network	Receiver	In-service measure-ment
11.3.16 BER before BCH (outer) decoder			X	X
11.3.17 Baseband Frame Error Rate BBFER			X	X
11.3.18 Errored Second Ratio ESR			X	X
11.3.19 IQ signal analysis	X		X	X
11.3.19.2 Modulation Error Ratio (MER)	X	X	X	X
11.3.19.3 Signal to Interference Noise Ratio (SINR)	X	X	X	X
11.3.19.4 Carrier Suppression (CS)	X			X
11.3.19.5 Carrier Phase	X			X
11.3.19.6 Amplitude Imbalance (AI)	X			X
11.3.19.7 Quadrature Error (QE)	X			X
11.3.20 SFN synchronization		X		X
11.3.21 L1 signalling error	X		X	X
11.3.22 RMS Delay-Spread (RMS-DS)		X	X	X
11.3.23 Maximum Excess Delay (MED)		X	X	X
11.3.24 Receiver Buffer Model (RBM) validation test			X	
11.3.25 Relative power Level during the non-P1 part of the FEF (RLF_non_P1)		X		X

NOTE 1: The term 'In-service measurement' is understood as a measurement that does not require a specific test signal but can be carried out with a normal DVB-T2 signal.

NOTE 2: As an In-service measurement in an unoccupied channel, the measurement of Noise Power can provide an overview of the man-made noise conditions in a certain channel or frequency band.

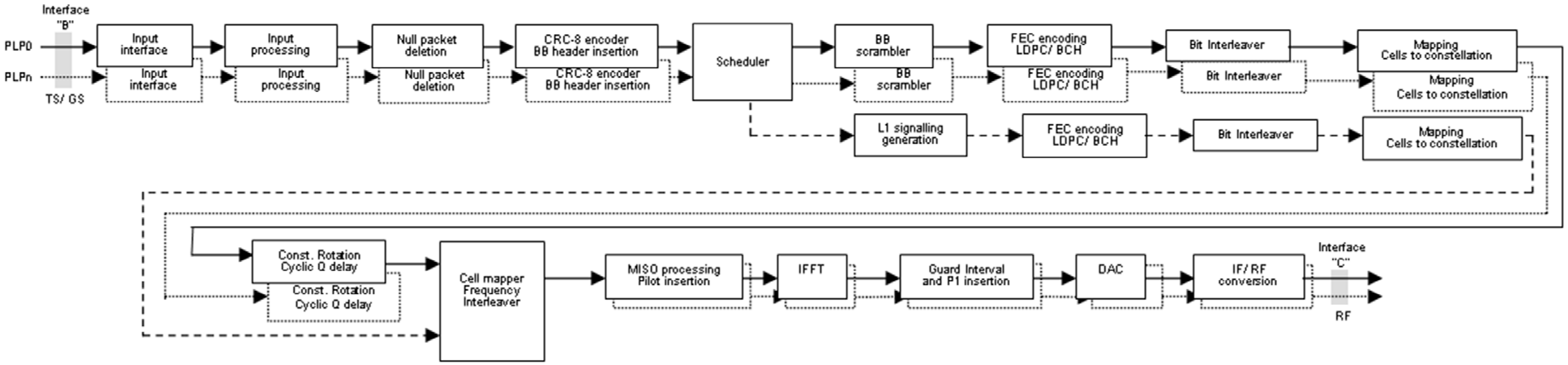


Figure 11.5: Simplified block diagram of a DVB-T2 transmitter

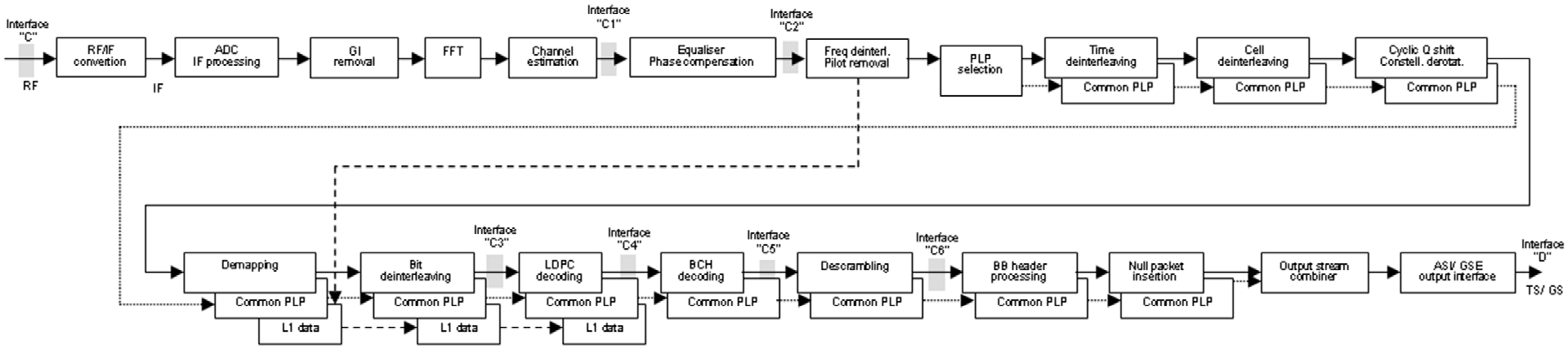


Figure 11.6: Simplified block diagram of a DVB-T2 receiver

11.3.1 RF measurements

11.3.1.0 General

The measurement of some basic parameters of the DVB-T2 OFDM signal may be carried out at the RF layer with a test receiver, a spectrum analyser or similar instruments.

11.3.1.1 RF frequency accuracy

Purpose	Successful processing of OFDM signals requires that certain carrier frequency accuracy be maintained at the transmitter. Specific network operations modes such as SFN require high accuracy of the carrier frequency.
Interface	C
Method	The measurement of the RF frequency accuracy determines the centre frequency of the signal, i.e. the positioning of the signal in the RF channel. a) Spectrum analyser method: the centre frequency is derived from the frequencies measured for the continual pilots and/or the edge pilots. b) Test receiver method: the centre frequency is derived from the digital samples after the test receiver has synchronized to the incoming DVB-T2 signal. In this case, the accuracy is typically expressed as 'Carrier Offset' and given in Hz or ppm.
Reference	Clauses 9.2.4, 9.2.5 of ETSI EN 302 755 [i.27].

NOTE: In any case, the usage of a high-precision reference frequency, e.g. 10 MHz, may be helpful.

11.3.1.2 RF occupied bandwidth

Purpose	The measurement of the occupied bandwidth allows the verification of the correct sampling frequency at the modulator.
Interface	C
Method	The occupied bandwidth is calculated from the measurements of the frequencies of the edge pilots and/or continual pilots of the DVB-T2 signal. a) Spectrum analyser method: if the frequency of the edge carriers is known then the related values for the occupied bandwidth may be calculated. Denoting the edge pilot frequencies as F_{min} and F_{max} the occupied bandwidth is appropriately OB = F_{max} - F_{min}+ $1/T_U$. b) Test receiver method: the occupied bandwidth is derived from the digital samples after the test receiver has synchronized to the incoming DVB-T2 signal.
Reference	Clause 9.2.5 of ETSI EN 302 755 [i.27].

11.3.2 Selectivity

Purpose	To identify the capability of the receiver to reject out-of-channel interference.
Interface	The measurement of the signal input level and the interferer should be carried out at the interface C, using interface C3 or C4 for the BER monitoring.
Method	The input power is adjusted to 10 dB above the minimum input power as defined in "Receiver sensitivity" (see clause 11.3.9). The C/I threshold needed for QEF operation should be measured as a function of the frequency of a CW interferer. The failure point is defined as ESR5. ESR (Errored Second Ratio) is defined in clause 11.3.18. One errored second during a time interval of 20 seconds is given as ESR5 (5 % of the seconds are errored). Alternatively, the QEF point is used that is defined as BER after LDPC = 10^{-7}. Since this measurement is time-consuming, the BER after LDPC = 10^{-4} is measured and the associated C/N value is increased by 0,2 dB. This corresponds to a BER after BCH = 10^{-11}.
Reference	Clause 6.1.2 of ETSI EN 302 755 [i.27].

11.3.3 AFC capture range

Purpose	To determine the frequency range over which the receiver will acquire overall lock.
Interface	C for the application of the test signal; C1 or C2 for the test of receiver synchronization
Method	A signal is applied to the input of the receiver, at a level 10 dB above the minimum input power as defined in "Receiver sensitivity" *(*see clause 11.3.9*)*. The signal is frequency shifted in steps (from below and above) towards a nominal value, whilst forcing the receiver to re-acquire after each step. Correct reception is assumed if after each step: a) the receiver can synchronize to the applied DVB-T2 signal, or if this is not indicated, b) the failure point is defined as ESR5. ESR (Errored Second Ratio) is defined in clause 11.3.18. One errored second during a time interval of 20 seconds is given as ESR5 (5 % of the seconds are errored). Alternatively, the QEF point is used that is defined as BER after LDPC = 10^{-7}. Since this measurement is time-consuming, the BER after LDPC = 10^{-4} is measured and the associated C/N value is increased by 0,2 dB. This corresponds to a BER after BCH = 10^{-11}.
Reference	Clause 5.1.7 of ETSI EN 302 755 [i.27].

11.3.4 Phase noise of Local Oscillators (LO)

Purpose	Phase noise can be introduced at the transmitter, at any frequency converter or by the receiver. In an OFDM system the phase noise can cause Common Phase Error (CPE) which affects all carriers simultaneously, and which can be minimized or corrected by using the continual pilots. However the noise-like Inter-Carrier Interference (ICI) cannot be corrected. This measurement may be useful for manufacturing, incoming inspection and maintenance of modulators, transmitters, up/down converters and receivers, either professional or consumer type.
Interface	Any access to Local Oscillators (LO), in transmitters, converters and receivers.
Method	Phase noise can be measured with a spectrum analyser, a vector analyser or a phase noise test set.
Reference	n/a

11.3.5 RF/IF signal power

Purpose	Signal power, or wanted power, measurement is required to set and check signal levels at the transmitter and receiver sites.
Interface	C
Method	The signal power of a DVB-T2 signal is defined as the mean power of the signal as would be measured with a thermal power sensor. In the case of received signals care should be taken to limit the measurement to the bandwidth at the wanted signal. When using a spectrum analyser or a calibrated receiver, it should integrate the signal power within the nominal bandwidth of the signal ($n \times f_{SPACING}$) where n is the number of carriers. Note that some spectrum analyser may not automatically compensate for the applied Resolution Bandwidth so that a manual correction may be applicable.
Reference	n/a

11.3.6 MISO Group Power Ratio

Purpose	The MISO Group Power Ratio (MGPR) is required to check the presence of both MISO Groups within a network.
Interface	C
Method	The MGPR is defined as the RF signal power for MISO Group 1 divided by the RF signal power for MISO Group 2. NOTE: The value should be given in dB. The MGPR may be negative if the signals of MISO Group 2 are received with higher input level than the signals of MISO Group 1. The MGPR takes the value of $\pm\infty$ in the presence of one MISO Group. For the display of the power levels of the paths of both MISO groups, the use of different colours for the figures or the diagram components (e.g. power level over delay) is recommended.
Reference	Clause 9.1 of ETSI EN 302 755 [i.27].

11.3.7 Noise power

Purpose	Noise is a significant impairment in a transmission network.
Interface	C (RF or IF)
Method	The noise power (mean power), or unwanted power, can be measured with a spectrum analyser (out of service). The noise power is specified using the occupied bandwidth of the OFDM signal ($n \times f_{SPACING}$) where n is the number of carriers. NOTE: The Carrier-to-Noise ratio C/N should be calculated as the ratio of the signal power, measured as described in clause 11.3.5, to the noise power, measured as described in this clause.
Reference	n/a

11.3.8 RF and IF spectrum

Purpose	To avoid interfering with other channels, the transmitted RF spectrum should comply with a spectrum mask, which is defined for the terrestrial network. If the spectrum at the modulator output is defined by a spectrum mask, the same procedure can be applied to the IF signal (with no pre-correction active).
Interface	C (RF or IF)
Method	This measurement is usually carried out with a spectrum analyser. The spectral density of a terrestrial DVB signal is defined as the long-term average of the time-varying signal power per unity bandwidth (i.e. 1 Hz). Values for other bandwidths can be achieved by proportional increase of the values for unity bandwidth. To avoid regular structures in the modulated signal a non-regular, e.g. a Pseudo-Random Binary Sequence (PRBS) is applied as an input signal to the modulator. For the resolution bandwidth, the recommended values should not exceed 30 kHz. Preferred values are approximately 3 kHz. The measurement should be Noise-normalized to 3 kHz.
Reference	Clause 10 of ETSI EN 302 755 [i.27].

11.3.9 Receiver sensitivity/dynamic range for a Gaussian channel

Purpose	For network planning purposes, the minimum and maximum input powers for normal operation of a receiver have to be determined.
Interface	Test signals are applied and measured at interface C; interfaces C3 or C4 are used for the monitoring of BER or PER.
Method	The minimum and maximum input power thresholds for QEF reception should be measured. The failure point is defined as ESR5. ESR (Errored Second Ratio) is defined in clause 11.3.18. One errored second during a time interval of 20 seconds is given as ESR5 (5 % of the seconds are errored). Alternatively, the QEF point is used that is defined as BER after LDPC = 10^{-7}. Since this measurement is time-consuming, the BER after LDPC = 10^{-4} is measured and the associated C/N value is increased by 0,2 dB. This corresponds to a BER after BCH = 10^{-11}. The dynamic range is the difference of the input power between the measured values in dB.
Reference	Clause 6.1.2 of ETSI EN 302 755 [i.27].

11.3.10 Linearity characterization (shoulder attenuation)

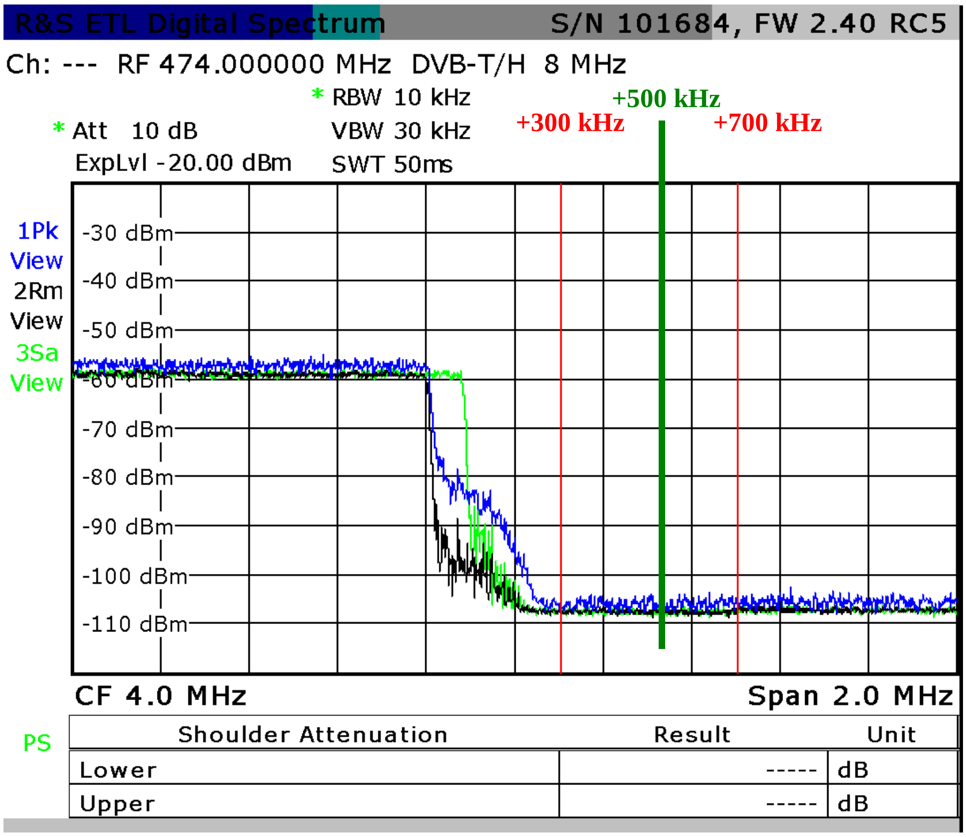
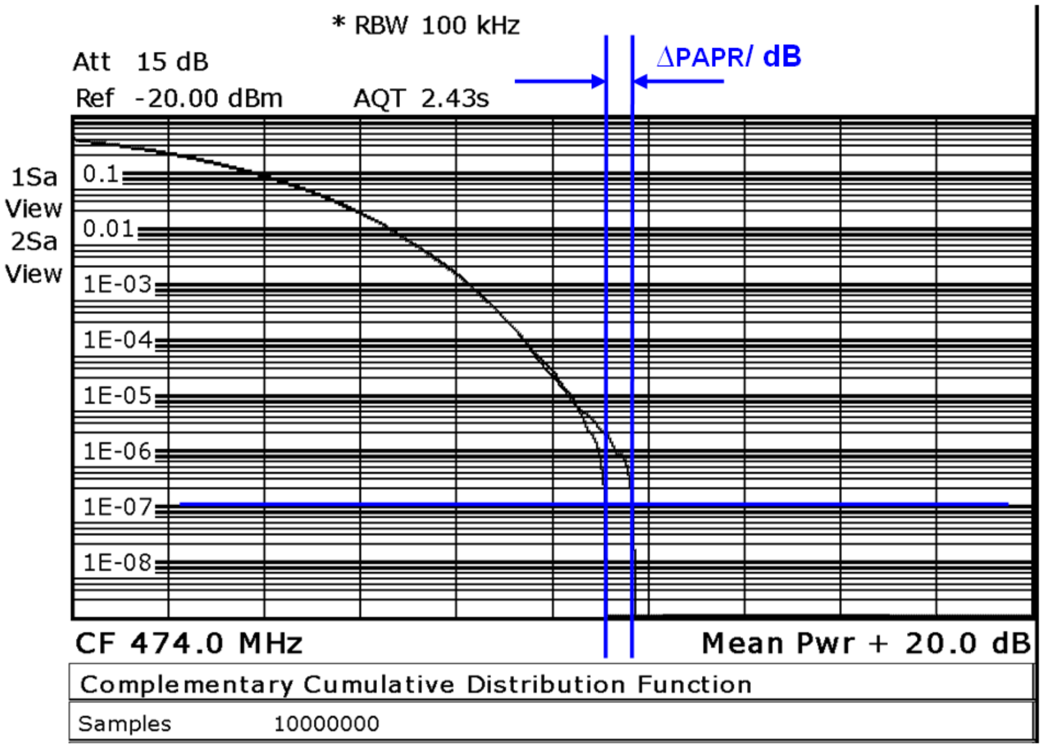
<table>
<tr><td>Purpose</td><td>The shoulder attenuation can be used to characterize the linearity of an OFDM signal.</td></tr>
<tr><td>Interface</td><td>C</td></tr>
<tr><td>Method</td><td>The following procedure is applied to the measured RF spectrum of the transmitter output signal:
a) Identify the maximum value of the spectrum by using a resolution bandwidth at approximately 10 times the carrier spacing.
b) Place declined, straight lines connecting the measurement points at 300 kHz and 700 kHz from each of the upper and lower edges of the spectrum. Draw additional lines parallel to these, so that the highest spectrum value within the respective range lies on the line.
c) Subtract the power value of the centre of the line (500 kHz away from the upper and lower edge) from the maximum spectrum value of (a) and note the difference as the "shoulder attenuation" at the upper and lower edge.
d) Take the worst case value of the upper and lower results from (c) as the overall "shoulder attenuation".
NOTE: For a quick overview the value at e.g. 500 kHz can be measured directly provided that coherent interferers are not present.
The method described here is used for DVB-T2 signals with a nominal channel width of 5, 6, 7, 8 or 10 MHz. The reference frequency points are the same for non-extended and extended modes (for reasons of protection of adjacent channels).
The recommended method is the same as for DVB-T for reason of comparability.
If a spectrum mask is used, it should be always the spectrum mask of the non-extended mode, even if the signal uses extended bandwidth. For DVB-T2 signals with a nominal channel width of 1,7 MHz an appropriate spectrum masks has to be defined. If such a spectrum mask is not specified, the measurement points should be set 150 kHz below or above the respective edge carrier.


Figure 11.7: Measured spectrum of different DVB-T2 transmission modes</td></tr>
<tr><td>Reference</td><td>Clause 10 of ETSI EN 302 755 [i.27].</td></tr>
</table>

11.3.11 Power efficiency

Purpose	To compare the overall efficiency of DVB transmitters.
Interface	C
Method	Power efficiency is defined as the ratio of the DVB output power to the total power consumption of the chain from TS input to the RF signal output including all necessary equipment for operation such as blowers, transformers etc. (and is usually quoted in % terms). The operational channel and the environmental conditions need to be specified.
Reference	n/a

11.3.12 PAPR effect

Purpose	To measure the PAPR effect in the OFDM signal by applying one of the PAPR methods (ACE Active Constellation Extension or TR Tone Reservation). The method is to measure the performance difference without the PAPR technique and with the PAPR. The parameters of interest are the power increase, CCDF, MER performance increase.
Interface	C (modulator output, transmitter output)
Method	**CCDF effect (**modulator output) Note that the CCDF impact should be measured before the amplification in order to measure the effect of amplifier saturation. Comparison of CCDF charts (cumulative complementary distribution function) and indicating the difference in dB in the CCDF chart for a specified probability (e.g. 10^{-7}). Special attention should be given to the requirement that the instrument settings should not be changed between the two measurements (with and without PAPR) and no overload or clipping is effecting the OFDM signal. The number of samples for such a comparison should be >= 10^7.  **Figure 11.8: Example of CCDF measurement of PAPR effect**

Reference	Clause 9.6 of ETSI EN 302 755 [i.27]. At transmitter output: **Power increase**: Signal Mean power is first measured before and after the PAPR is applied. For the tone reserved carriers, it means that it is measured with empty tone reserved carriers (before PAPR is applied) and after PAPR is applied. The power increase should be kept as low as possible. **MER improvement**: MER is first measured before and after the PAPR is applied. For the tone reserved carriers, it means that it is measured with empty tone reserved carriers (before PAPR is applied) and after PAPR is applied. For ACE case, MER measurement method should follow the procedure described in annex L.

11.3.13 P1 Symbol Error Rate

Purpose	This measurement gives an indication of the P1 Symbol Error rate.
Interface	C5
Method	The data transmitted within the S1 and S2 field are decoded and compared to the correct data. A P1 symbol is erroneous, if at least one bit error within the S1 or the S2 field occurred. The measurement of this variable allows an estimation of the T2 signal level if the actual payload data is not decodable. The correct values of the S1 and S2 field can e.g. be obtained by the S1/S2 field transmitted within the L1-Pre data.
Reference	Clause 6.1.2 of ETSI EN 302 755 [i.27].

11.3.14 BER before LDPC (inner) decoder

Purpose	This measurement gives an in-service indication of the un-coded performance of the transmitter. Since the residual BER contributions from transmitter and (test) receiver contribute to the result of this measurement, the contribution of the (test) receiver should be negligible when validating a transmitter.
Interface	C3
Method	The BER before LDPC is measured separately for each PLP. It allows the identification of sporadic bit errors in a transmitter output signal. The averaging period for the calculation of the BER before LDPC should be set so that sporadic bit errors are not averaged out.
Reference	Clause 6.1.2 of ETSI EN 302 755 [i.27].

11.3.15 Number of LDPC iterations

Purpose	This measurement gives an in-service indication of the quality of the received signal and the computational resources activated for the LDPC decoder. Since the result of this measurement is largely dependent on the actual LDPC decoder implementation, results can only be compared when taken from the same test instrument.
Interface	C3
Method	The number of LDPC iterations is measured separately for each PLP. The end of the iterations is reached when the number of remaining errors is lower or equal than the error correction capability of the following BCH decoder, or when the maximum number of LDPC iterations is reached. An error-free signal requires a minimum of one iteration of the LDPC decoder. The average of the Number of LDPC iterations should be calculated over 1 second, and the maximum value during 1 second should also be displayed together with the average value. In case the data rate is very low and frames of the respective PLP are received at longer intervals, periods for averaging and display of maxima should be set accordingly.
Reference	Clause 6.1.2 of ETSI EN 302 755 [i.27].

NOTE: It is recommended to provide an indication if the LDPC decoder does not converge.

11.3.16 BER before BCH (outer) decoder

Purpose	The BER is the primary parameter which describes the quality of the digital transmission link. It provides a quick indication of potential problems, especially in cases were sporadic errors would lead only to small increases of BER after BCH.
Interface	C4
Method	The BER before BCH is measured separately for each PLP. The calculation can be based on the re-encoded signal that is available after the BCH decoder. The BER is defined as the ratio between erroneous bits and the total number of transmitted bits. The time interval for this calculation should be definable.
Reference	Clause 6.1.1 of ETSI EN 302 755 [i.27].

11.3.17 Baseband Frame Error Rate BBFER

Purpose	To gain information about the number of baseband frames which are effected by bit errors.
Interface	C5
Method	The BBFER is measured separately for each PLP. A Baseband Frame is erroneous, if an uncorrectable error has been discovered and indicated by the error flag by the BCH decoder. The parameter is either given as a ratio or as the number of erroneous BB frames per second.
Reference	Clause 5.1.7 of ETSI EN 302 755 [i.27].

11.3.18 Errored Second Ratio ESR

Purpose	To obtain statistical information about the link quality.
Interface	C5
Method	Time intervals of the length of 1 second are defined according to the system clock reference. Those seconds during which a Baseband Frame Error occurred are marked as 'errored'. The ratio of Errored Seconds to the total time elapsed or to a period set by the test receiver is indicated as ESR. In practice, a time interval of 20 seconds is used.
Reference	Clause 5.1.7 of ETSI EN 302 755 [i.27].

11.3.19 IQ signal analysis

11.3.19.1 Introduction

The IQ analysis can be applied on single carriers of the OFDM signal as well as on groups of carriers. If a group of carriers is evaluated all received symbols of this group can be superimposed in order to obtain the respective measurement parameters or the constellation diagram.

The definitions for the I/Q related parameters are based on the following assumptions:

- a constellation diagram of M symbol points and K carriers under consideration with $0 < K \leq K_{MAX} + 1$ and $K_{MAX} + 1$ is the total number of active OFDM carriers;
- a measurement sample of N data points, where N is sufficiently larger than $M \times K$ to deliver the wanted measurement accuracy; and
- the co-ordinates of each received data point j being $I_j + \delta I_j$, $Q_j + \delta Q_j$ where I and Q are the co-ordinates of the ideal symbol point and δI and δQ are the offsets forming the error vector of the data point (as long as the respective carrier is a "useful" one).

11.3.19.2 Modulation Error Ratio (MER)

Purpose	To provide a "figure of merit" analysis for L1 Signalling data and each PLP of the DVB-T2 signal, typically at a transmitter output (for assessing the quality of the transmitted signal) or in a fixed location in a SFN (for identifying severe distortions in the set-up of the transmitters forming the SFN).
Interface	C2 (i.e. after equalization)
Method	The carrier frequency of the OFDM signal and the symbol timing are recovered. Origin offset of the centre carrier (e.g. caused by residual carrier or DC offset), Quadrature Error (QE) and Amplitude Imbalance are not corrected. A time record of N received symbol co-ordinate pairs $(\tilde{I}_j, \tilde{Q}_j)$ is captured. For each received symbol, a decision is made as to which symbol was transmitted. The error vector is defined as the distance from the ideal position of the chosen symbol (the centre of the decision box) to the actual position of the received symbol. This distance can be expressed as a vector $(\delta I_j, \delta Q_j)$. The sum of the squares of the magnitudes of the ideal symbol vectors is divided by the sum of the squares of the magnitudes of the symbol error vectors. The result, expressed as a power ratio in dB, is defined as the MER. $$MER = 10 \times \log_{10} \left\{ \frac{\sum_{j=1}^{N} \left(I_j^2 + Q_j^2 \right)}{\sum_{j=1}^{N} \left(\delta I_j^2 + \delta Q_j^2 \right)} \right\} dB$$ It should be reconsider that MER is just one way of computing a "figure of merit" for a vector modulated signal. Another "figure of merit" calculation is Error Vector Magnitude (EVM). MER and EVM are closely related and one can generally be computed from the other. Error Vector Magnitude (EVM) is defined as: $$EVM_{RMS} = \sqrt{\frac{\frac{1}{N}\sum_{j=1}^{N}\left(\delta I_j^2 + \delta Q_j^2\right)}{S_{\max}^2}} \times 100\%$$ Where I and Q are the ideal co-ordinates, δI and δQ are the errors in the received data points. N is the number of data points in the measurement sample. S_{max} is the magnitude of the vector to the outermost state of the constellation. For both parameters, MER and EVM, the display of the values as a function of frequency or carrier number can be very helpful.

11.3.19.3 Signal to Interference Noise Ratio (SINR)

Purpose	To provide a "figure of merit" analysis for L1 Signalling data and each PLP of the DVB-T2 signal with special focus on field measurements where the MER is not applicable.
Interface	C1 (i.e. before equalization)
Method	The carrier frequency of the OFDM signal and the symbol timing are recovered. Origin offset of the centre carrier (e.g. caused by residual carrier or DC offset), Quadrature Error (QE) and Amplitude Imbalance are not corrected. A time record of N received complex QAM cells $\tilde{d}_j$ is captured at the input of the cell de-interleaver without equalization. For each received QAM cell, a decision is made which QAM symbol $\hat{d}_j$ was transmitted. This value is then multiplied by the estimated channel transfer function $\hat{H}_j$, which defines the ideal position of the received QAM cell. The absolute squared value $\left\|\hat{H}_j \cdot \hat{d}_j\right\|^2$ defines its signal power. In the absence of any disturbance, the value $\hat{H}_j \cdot \hat{d}_j$ should ideally match $\tilde{d}_j$. The absolute squared value of the delta equals the interference plus noise power. The result, expressed as a power ratio in dB: $$SINR = 10 \times \log_{10} \left\{ \frac{\sum_{j=1}^{N} \left\|\hat{H}_j \cdot \hat{d}_j\right\|^2}{\sum_{j=1}^{N} \left\|\hat{H}_j \cdot \hat{d}_j - \tilde{d}_j\right\|^2} \right\} dB$$ It should be mentioned that the SINR is very similar to the MER, except that no equalization to the input data is performed. This avoids the effect of noise amplification in frequency selective channels. The display of the SINR values as a function of frequency or carrier number can be very helpful.

11.3.19.4 Carrier Suppression (CS)

Purpose	A residual carrier is an unwanted coherent signal added to the centre carrier of the OFDM signal. It may have been produced by DC offset voltages of the modulating I and/or Q signal or by crosstalk from the modulating carrier within the modulator.
Interface	C2
Method	Search for systematic deviations of all constellation points of the centre carrier and isolate the residual carrier. Calculate the Carrier Suppression (CS) from the formula: $$CS = 10 \times \log_{10} \left(\frac{P_{sig}}{P_{RC}} \right)$$ where P_{RC} is the power of the residual carrier and P_{sig} is the power of the centre carrier of the OFDM signal (without residual carrier).

11.3.19.5 Carrier Phase (CPh)

Purpose	The measurement of the phase of the residual carrier which is an unwanted coherent signal added to the centre carrier of the OFDM signal, allows for an automatic and efficient calibration of a modulator which optimizes the Carrier Suppression (CS).
Interface	C2
Method	This parameter describes the phase of a not perfectly suppressed carrier (see Carrier Suppression). When the Carrier Suppression is infinite, a carrier phase cannot be measured. The reference for the Carrier Phase is the I axis. A residual carrier that is directed towards positive I values has the carrier phase of 0°. A residual carrier that is directed towards positive Q values has the carrier phase of 90°.

11.3.19.6 Amplitude Imbalance (AI)

Purpose	To separate the QAM distortions resulting from Amplitude Imbalance (AI) of the I and Q signal from all other kind of distortions.
Interface	C2
Method	Calculate the I and Q gain values v_I and v_Q from all points in a constellation diagram eliminating all other influences. Calculate Amplitude Imbalance (AI) from v_I and v_Q $$AI = \begin{cases} \left(\frac{v_I}{v_Q} - 1\right) \times 100\% & if\ v_I \geq v_Q \\ \left(1 - \frac{v_Q}{v_I}\right) \times 100\% & if\ v_Q > v_I \end{cases}$$ $$v_I = \frac{1}{M}\sum_{i=1}^{M}\frac{I_i + \left(\bar{d}_i\right)_I}{I_i}$$ $$\left(\bar{d}_i\right)_I = \frac{1}{N}\sum_{j=1}^{N}\delta I_j \qquad \text{(I - component of } d_i \text{ as given in clause 9.18.3 System Target Error)}$$ $$v_Q = \frac{1}{M}\sum_{i=1}^{M}\frac{Q_i + \left(\bar{d}_i\right)_Q}{Q_i}$$ $$\left(\bar{d}_i\right)_Q = \frac{1}{N}\sum_{j=1}^{N}\delta Q_j \qquad \text{(Q - component of } d_i \text{ as given in clause 9.18.3 System Target Error)}$$ $$\left(\bar{d}_i\right)_I + \left(\bar{d}_i\right)_Q = \bar{d}_i$$

11.3.19.7 Quadrature Error (QE)

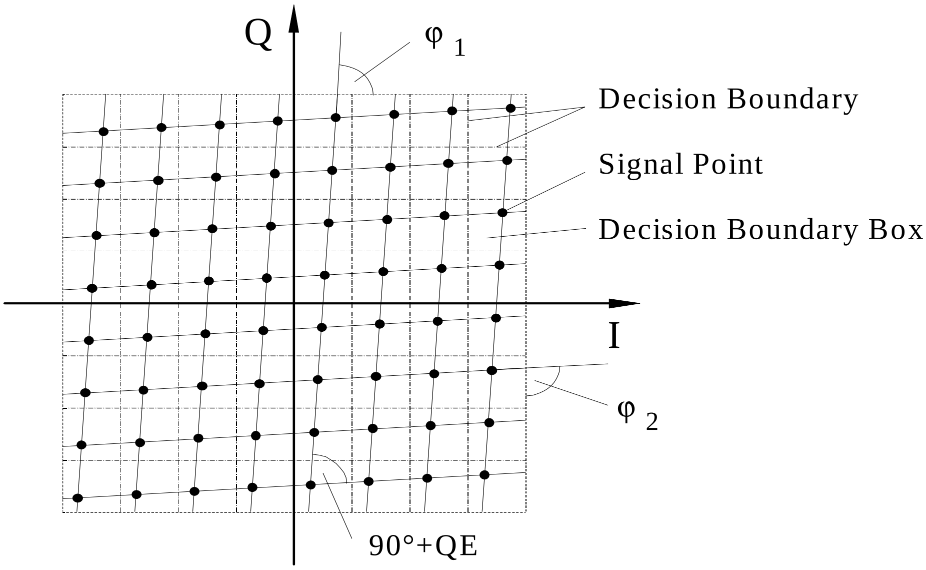
<table>
<tr><td>Purpose</td><td>The phases of the two carriers feeding the I and Q modulators have to be orthogonal. If their phase difference is not 90° a typical distortion of the constellation diagram results.</td></tr>
<tr><td>Interface</td><td>C2</td></tr>
<tr><td>Method</td><td>Search for the constellation diagram error shown in Figure 11.9 and calculate the value of the phase difference $\Delta\varphi = \varphi_1 - \varphi_2$ after having eliminated all other influences and convert this into degrees:
$$QE = \frac{180°}{\pi} \times (\varphi_1 - \varphi_2) \quad [°]$$


Figure 11.9: Distortion of constellation diagram resulting from I/Q Quadrature Error (QE)</td></tr>
</table>

11.3.20 SFN synchronization

Purpose	To measure and adjust the signal delay of an OFDM transmitter to a given value so that the transmitters in an SFN can be synchronized.
Interface	C1
Method	Based on a sufficient number of I/Q samples, the channel response of the received DVB-T2 signal is calculated. The measurement of the SFN synchronization depends on the location where this measurement is executed. The SFN synchronization is related to two independent parameters: a) Time synchronization: All signals received from the certain transmitters at the location of interest should be received within the Guard Interval. It is the task of the network planning that also echoes from these signals will not infringe this rule. Method: The complex channel impulse response is measured. In addition, the Guard Interval is indicated. Also pre-echoes are taken into account correctly, according to the positioning of the FFT window. The indication of the Guard Interval should always start with the start of the FFT window position. b) Frequency synchronization: The center frequency of all signals received from the certain transmitters should be within a certain limit. In practice, 1 Hz has been established as a good and practicable target value for this limit in case of an 8 MHz channel and an 8k FFT size. For other channel bandwidth and other FFT sizes, e.g.16k and 32k signals, a proportionally higher accuracy may be required. The SFN frequency offset can be calculated as a relative value. It should be referenced to the signal path that provides the maximum power. Method: The complex channel impulse response is measured from which the frequency offset of each echo is derived and indicated. Also pre-echoes have to be taken into account correctly. A positive SFN frequency offset value means that the local center frequency of the related echo is higher than the local center frequency of the reference signal.

11.3.21 L1 signalling error

Purpose	The consistency of the L1 data is essential for the successful decoding of the DVB-T2 signal by the receiver.
Interface	C3
Method	The receiver checks the CRC within the corresponding L1-pre and L1-post data.

11.3.22 RMS Delay-Spread (RMS-DS)

Purpose	The performance of OFDM based systems depends on the characteristics of the transmission channel. The extent of the multipath propagation of the transmission channel is characterized by means of the root means square (RMS) of the channel delay spread.
Interface	C1
Method	The RMS Delay-Spread is directly measured from the channel impulse response. It is calculated from all i multipath components with power P_i and excess delay τ_i that exceed a specified power threshold (e.g. -15 dB, -20 dB, -25 dB) with respect to the strongest component. The excess delay (or delta) τ_i is defined as the delta of the multipath component i with respect to the strongest signal component. The calculation should be limited to the Guard Interval to avoid ambiguities. $$\tau_{RMS} = \sqrt{\frac{1}{\sum_{i=1}^{n} P_i} \cdot \sum_{i=1}^{n} (\tau_i^2 P_i) - \tau_d^2} \; with \; \tau_d = \frac{\sum_{i=1}^{n} (\tau_i P_i)}{\sum_{i=1}^{n} P_i}$$

11.3.23 Maximum Excess Delay (MED)

Purpose	The performance of OFDM based systems depends on the characteristics of the transmission channel. The largest multipath delay of the transmission channel is characterized by means of the Maximum Excess Delay (MED).
Interface	C1
Method	The Maximum Excess Delay is directly measured from the channel impulse response. It is defined as the maximum delay $\tau_{\max}$ of all multipath components that exceed a specified power threshold of e.g. -15 dB, -20 dB, -25 dB with respect to the strongest component. The maximum excess delay $\tau_{\max}$ is defined with regard to the first arrived component above the specified power threshold.

11.3.24 Receiver Buffer Model (RBM) validation test

Purpose	This test applies a test stream for an out-of-service test to the receiver to validate the proper functioning of the receiver buffer.
Interface	C
Method	A DVB-T2 signal is applied to the input of a DVB-T2 receiver. This DVB-T2 signal contains a test stream (e.g. V&V number 7xx). The test stream is generated (typically in the T2 gateway) and includes a modified Common PLP. The modified Common PLP for this test does not contain the tables and service information as usual, but an externally applied video/audio test stream. The failure point is defined as ESR5. ESR (Errored Second Ratio) is defined in clause 11.3.18. One errored second during a time interval of 20 seconds is given as ESR5 (5 % of the seconds are errored).

11.3.25 Relative power level during the non-P1 part of the FEF (RLF_non_P1)

Purpose	The purpose of the test is to measure the average power level of the non-P1 signal (excluding the rise time/fall time between the P1 and adjacent waveforms) of the FEF relative to the power of the T2 signal, excluding the FEF. This procedure allows for in-service measurements of co-channel interference (CCI) when empty FEFs are applied.
Interface	C
Method	The DVB-T2 system makes it possible to broadcast a FEF part carrying a non-T2 waveform at potentially another power level, including an empty FEF. Empty FEF means null power during non-P1 part of the FEF. A DVB-T2 signal is generated including a FEF part, which contains some other waveform than DVB-T2, e.g. T2 TX-SIG or an empty FEF. During the FEF period the power level of the non-P1 part of the FEF is measured. In addition the power level of the DVB-T2 signal (excluding the FEF part) is measured. A fast enough power sensor gated properly, or a spectrum analyser capable for channel power measurements, or a DVB-T2 receiver can be used for this measurement.

NOTE: Note that DVB-T2 receivers of the 1st generation should ignore FEFs.

12 Measurements for the second generation cable system (DVB-C2)

12.1 Introduction

The DVB-C2 system as it is addressed in the following clauses, spans from the input interface (MPEG TS or GSE stream) to the output interface (RF signal) of the DVB-C2 transmitter.

NOTE: The development of DVB-C2 modulator interface is not finalized at the time of completion of these Measurement Guidelines.

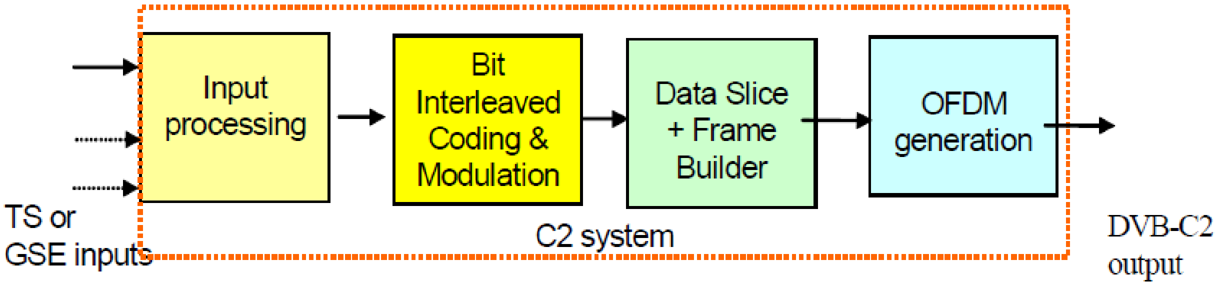


Figure 12.1: High level C2 modulator block diagram [i.28]

DVB-C2 supports TS, any packetized and continuous input formats as well as the so called Generic Stream Encapsulation (GSE). All input streams are multiplexed into a Baseband Frame format. The Forward Error Correction (FEC) scheme is applied to these Baseband Frames. In line with the other DVB-X2 systems, DVB-C2 uses a combination of LDPC and BCH codes, which is a very powerful FEC providing about 6 dB improvement of signal-to-noise ratio (SNR) with reference to DVB-C.

Appropriate Bit-Interleaving schemes optimize the overall robustness of the FEC system. Extended by a FEC Frame header, those frames compose the payload of Physical Layer Pipes (PLP). One or several of such PLPs are multiplexed into a Data Slice. A two-dimensional interleaving (in time and frequency domain) is applied to each slice enabling the receiver to eliminate impacts of burst impairments and frequency selective interference such as single frequency ingress. One or several Data Slices compose the payload of a C2-frame.

The Frame Building process includes inter alia the insertion of Continual and Scattered Pilots. The first symbol of a DVB-C2 frame, the so-called "Preamble", carries the signalling data. A DVB-C2 receiver will find all relevant configuration data about the structure and the technical parameters of the DVB-C2 signal in the signalling data block in the Preamble as well as in the headers of the PLPs. In the following step the OFDM symbols are generated by means of an Inverse Fast Fourier Transformation (IFFT). A 4K-IFFT algorithm is applied generating a total of 4 096 sub-carriers, 3 409 of which are actively used for the transmission of data and pilots within a frequency band of 8 MHz. The guard interval used between the OFDM symbols has a relative length of either 1/128 or 1/64 in reference to the symbol length (448 µs).

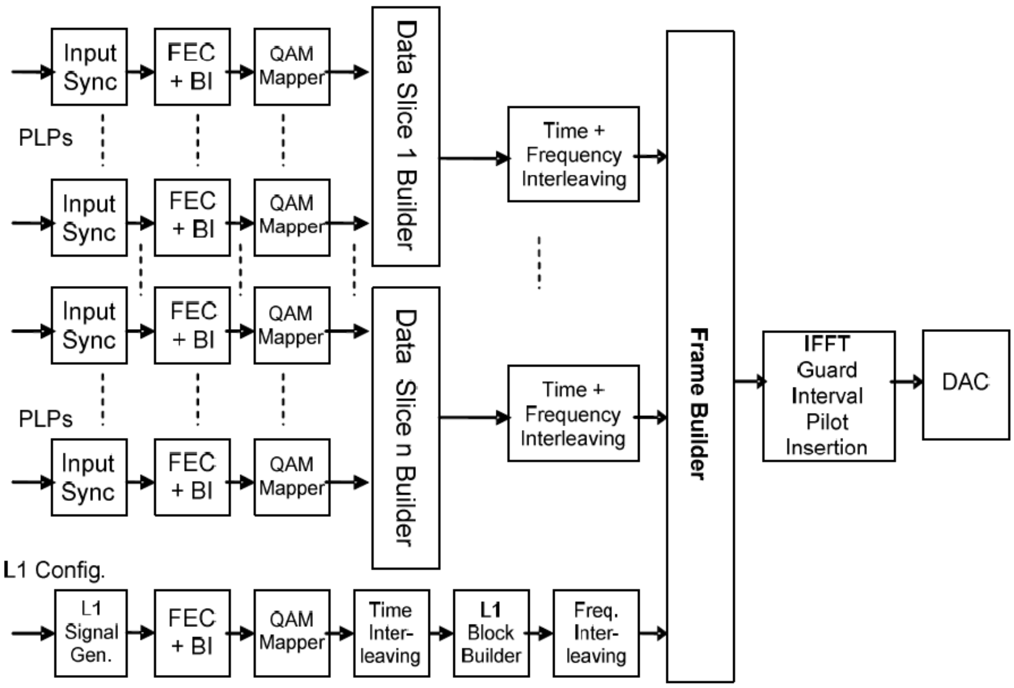


Figure 12.2: High-level block diagram of the signal processing defined for the DVB-C2 transmitting end [i.29]

The following clauses specify a number of tests and measurements at the modulator output interface and the corresponding input interface at the receiver side.

With regard to the signalling, it is recommended that a measurement instrument should display the signalled information (DVB-C2 Layer 1 and Layer 2 signalling) as readable text and abbreviations.

12.2 Measurements for DVB-C2 baseline system

12.2.0 General

This clause lists a number of measurements for the DVB-C2 baseline system.

A list of the main application area of the DVB-C2 measurement parameters described in this clause is given in table 11.1.

The measurements in clause 6.1 "System availability" and clause 6.2 "Link availability" are also applicable at the TS interfaces of transmitter and receiver.

Table 12.1 DVB-C2 measurement parameters and their application area

Measurement parameter				
	Transmitter	**Network**	**Receiver**	**In-service measurement**
11.3.1 RF measurements				
11.3.1.1 RF frequency accuracy	X	X		X
11.3.1.2 RF occupied bandwidth	X			X
11.3.3 AFC capture range			X	
11.3.4 Phase noise of Local Oscillators (LO)	X		X	
11.3.5 RF/IF signal power	X	X	X	X
11.3.7 Noise Power	X	X	X	X
11.3.8 RF and IF spectrum	X		X	X
11.3.9 Receiver sensitivity/dynamic range for a Gaussian channel			X	
11.3.10 Linearity characterization (shoulder attenuation)	X			
11.3.14 BER before LDPC (inner) decoder			X	X
11.3.15 Number of LDPC iterations			X	X
11.3.16 BER before BCH (outer) decoder			X	X
11.3.17 Baseband Frame Error Rate BBFER			X	X
11.3.19 IQ signal analysis				
11.3.19.2 Modulation Error Ratio (MER)	X		X	X
11.3.19.6 Amplitude Imbalance (AI)	X			X
11.3.19.7 Quadrature Error (QE)	X			X
11.3.21 L1 signalling error	X		X	X
11.3.24 Receiver Buffer Model (RBM) validation test			X	

NOTE: As an In-service measurement in an unoccupied channel, the measurement of Noise Power can provide an overview of the conditions in a certain frequency band.

12.2.1 RF measurements

12.2.1.0 General

The measurement of some basic parameters of the DVB-C2 OFDM signal may be carried out at the RF layer with a test receiver, a spectrum analyser or similar instruments.

12.2.1.1 RF frequency accuracy

Purpose	Successful processing of OFDM signals requires that certain carrier frequency accuracy needs to be maintained at the transmitter with respect to the relative and absolute frequency position of the signal/carriers.
Interface	RF
Method	The measurement of the RF frequency accuracy determines the relative carrier offset regarding its carrier spacing. Furthermore, the measurement determines the absolute frequency/carrier number offset of the signal with respect to its configured frequency location (absolute carrier position). a) Test receiver method: relative carrier offset is derived from the digital samples after the test receiver has synchronized to the incoming DVB-C2 signal. In this case, the accuracy is typically expressed as 'Carrier Offset' and given in Hz or ppm. Furthermore, the absolute frequency offset is typically given in number of carriers. b) Spectrum analyser method: the relative carrier offset is derived from the frequencies measured for the continual pilots and/or the edge pilots which are continuously (per OFDM symbol) transmitted at a boosted amplitude in contrary to the scattered pilots. The absolute frequency offset can only be determined in knowledge of the DVB-C2 signal configuration, i.e. the Layer 1 Signalling and the START_FREQUENCY parameter, or by pattern matching of the continual pilot positions. NOTE: The assumption is that the frequency offset is not greater than the frequency difference between adjacent pilots.
Reference	Clauses 9.6.3, 9.6.4 of ETSI EN 302 769 [i.28].

12.2.1.2 RF sampling frequency

Purpose	The measurement of the occupied bandwidth allows the verification of the correct sampling frequency at the modulator.
Interface	RF
Method	The occupied bandwidth (7,61 MHz or 5,71 MHz respectively) is calculated from the measurements of the frequencies of the edge pilots and/or continual pilots of the DVB-C2 signal. a) Spectrum analyser method: if the frequencies of the edge carriers of the DVB-C2 spectrum boundary are known then the related values for the occupied bandwidth may be calculated. Denoting the edge pilot frequencies as Fmin and Fmax the occupied bandwidth is appropriately OB = Fmax - Fmin + 1/TU. NOTE: If the measurement is based on pilots that are closer to each other than 7,61 MHz or 5,71 MHz in the frequency domain, a reduction of the accuracy of the calculated occupied bandwidth should be considered. Notches in the channel where the measurement is carried out need to be taken into account. b) Test receiver method: the occupied bandwidth is derived from the digital samples after the test receiver has synchronized to the incoming DVB-C2 signal.
Reference	Clauses 9.6.3, 9.6.4 of ETSI EN 302 769 [i.28].

12.2.2 AFC capture range

Purpose	To determine the frequency range over which the receiver will acquire overall lock.
Interface	RF input for the application of the test signal
Method	• Set RF tuning frequency in the C2 signalling to L1 START_FREQUENCY + OFDM bandwidth/2 in the modulator AND demodulator (C2 'Absolute OFDM' requirement) as a first step; • then repeat the procedure as described hereafter: A signal is applied to the input of the receiver, at a level 10 dB above the minimum input power as defined in "Receiver sensitivity" *(*see clause 11.3.9*)*. The signal is frequency shifted in steps (from below and above) towards a nominal value, whilst forcing the receiver to re-acquire after each step. Correct reception is assumed if after each step: a) the receiver can synchronize to the applied DVB-C2 signal, b) the Frame Error Rate or the rate of visible errors is below the failure point which is defined as 1 frame error or visible error in 20 seconds. • note the signal could be frequency shifted at the modulator or the receiver RF tuner, if shifted at the modulator then the RF frequency needs to be decoupled from the L1 signalling (L1 signalling SHOULD NOT change while the LO frequency is changed otherwise the demodulator will not acquire to the signal even if within the AFC capture range) • OFDM bandwidth/2 = 1 704 × 64/(7 × 4 096) MHz for 8 MHz tuning window (or OFDM bandwidth/2 = 1 704 × 48/(7 × 4 096) MHz for 6 MHz tuning window) • for tuning window > 8 MHz or 6 MHz respectively adjust tuning frequency to centre point between the two outermost OFDM subcarriers of the C2 system
Reference	Clause 10.1.1.1 of ETSI TS 102 991 [i.29].

12.2.3 Phase noise of Local Oscillators (LO)

Purpose	Phase noise can be introduced by the transmitter, by any frequency converter or by the receiver. In an OFDM system the phase noise can cause Common Phase Error (CPE) which affects all carriers simultaneously, and which can be minimized or corrected by using the continual pilots. However the noise-like Inter-Carrier Interference (ICI) cannot be corrected. This measurement may be useful for manufacturing, incoming inspection and maintenance of modulators, transmitters, up/down converters and receivers, either professional or consumer type.
Interface	Any access to Local Oscillators (LO), in transmitters, converters and receivers.
Method	Phase noise can be measured with a spectrum analyser, a vector analyser or a phase noise test set.
Reference	n/a

12.2.4 RF/IF signal power

Purpose	Signal power, or wanted power, measurement is required to set and check signal levels at the transmitter and receiver sites.
Interface	RF
Method	The signal power of a DVB-C2 signal is defined as the mean power of the signal between the outermost pilots of the 7,61 MHz (or of the 5,71 MHz) signal. The measurement is typically carried out with a spectrum analyser or a test receiver that integrates the signal power between a lower and an upper limit on the frequency axis. NOTE: The same procedure can be applied to wider DVB-C2 signals, e.g. 32 MHz at the output of a modulator/transmitter. If the signal includes notches, they have to be empty, i.e. not filled with other signals.
Reference	n/a

12.2.5 Noise power

Purpose	Noise is a significant impairment in a transmission network.
Interface	RF
Method	The noise power , or unwanted power, can be measured with a spectrum analyser (out of service). The noise power is specified using the occupied bandwidth of the C2 signal, i.e. 7,61 MHz or 5,71 MHz. The measurement can also be carried out as an in-service measurement by inserting DVB-C2-compliant notches where the noise power is measured by a spectrum analyser or a test receiver. NOTE: The Carrier-to-Noise ratio C/N should be calculated as the ratio of the signal power, measured as described in clause 11.3.5, to the noise power, measured as described in this clause.
Reference	n/a

12.2.6 RF and IF spectrum

Purpose	To avoid interfering with other channels, the transmitted RF spectrum should comply with a spectrum mask, which is defined for the cable network, typically by the cable network operator.
Interface	RF
Method	This measurement is usually carried out with a spectrum analyser or a test receiver. The spectral density of a DVB-C2 signal is defined as the long-term average of the time-varying signal power per unity bandwidth (i.e. 1 Hz). Values for other bandwidths can be achieved by proportional increase of the values for unity bandwidth. For the resolution bandwidth, the recommended values should not exceed 30 kHz. Preferred values are 3 kHz or lower. The measurement should be Noise-normalized to the same bandwidth.
Reference	Clause 10.3 of ETSI EN 302 769 [i.28].

12.2.7 Receiver sensitivity/dynamic range for a Gaussian channel

Purpose	For network planning purposes, the minimum and maximum input powers for normal operation of a receiver have to be determined.
Interface	Test signals are applied at the RF input and measured at Baseband level (or by expert viewing on a screen).
Method	The minimum and maximum input power thresholds for good reception should be measured by stepwise changes of the input power. For 'good reception', the Frame Error Rate or the rate of visible errors is below the failure point which is defined as 1 frame error or visible error in 20 seconds. Both measurement should provide results that are very close to the cut-off point as determined by the chosen FEC parameters (typically within 0,1 dB). The dynamic range is the difference between the maximum and minimum input power in dB.
Reference	n/a

12.2.8 Linearity characterization (shoulder attenuation)

Purpose	The linearity of an OFDM signal characterizes the level of interference in the adjacent bands. The requirements are normally defined by the cable network operator who also is likely to define acceptable levels for ACLR (Adjacent Channel Leakage Ratio), out-of-band emissions and spurious emissions.
Interface	RF (Modulator/transmitter output)
Method	Spectrum analyser or test receiver measurement
Reference	n/a

12.2.9 BER before LDPC (inner) decoder

Purpose	This measurement gives an in-service indication of the un-coded performance of the transmitter. Since the residual BER contributions from transmitter and (test) receiver contribute to the result of this measurement, the contribution of the (test) receiver should be negligible when validating a transmitter.
Interface	Baseband
Method	The BER before LDPC is measured separately for each PLP. It allows the identification of sporadic bit errors in a transmitter output signal if the averaging period for the calculation of the BER before LDPC is set so that sporadic bit errors are not averaged out.
Reference	Clause 6.1 of ETSI EN 302 769 [i.28].

12.2.10 Number of LDPC iterations

Purpose	This measurement gives an in-service indication of the quality of the received signal and the computational resources activated for the LDPC decoder. Since the result of this measurement is largely dependent on the actual LDPC decoder implementation, results can only be compared when taken from the same test instrument.
Interface	Baseband
Method	The number of LDPC iterations is measured separately for each PLP. The end of the iterations is reached when the number of remaining errors is lower or equal than the error correction capability of the following BCH decoder, or when the maximum number of LDPC iterations is reached. An error-free signal requires a minimum of one iteration of the LDPC decoder. The average of the number of LDPC iterations should be calculated over 1 second, and the maximum value during 1 second should also be displayed together with the average value. In case the data rate is very low and frames of the respective PLP are received at longer intervals, periods for averaging and display of maxima should be set accordingly.
Reference	Clauses 10.9, 11.2 of ETSI TS 102 991 [i.29].

NOTE: It is recommended to provide an indication if the LDPC decoder does not converge.

12.2.11 BER before BCH (outer) decoder

Purpose	The BER is the primary parameter which describes the quality of the digital transmission link. It provides a quick indication of potential problems, especially in cases where sporadic errors would lead only to small increases of BER after BCH.
Interface	Baseband
Method	The BER before BCH is measured separately for each PLP. The calculation can be based on the re-encoded signal that is available after the BCH decoder. The BER is defined as the ratio between erroneous bits and the total number of transmitted bits. The time interval for this calculation should be definable.
Reference	Clause 8.4.3.2 of ETSI EN 302 769 [i.28].

12.2.12 Baseband Frame Error Rate BBFER

Purpose	To gain information about the number of baseband frames which are affected by bit errors.
Interface	Baseband
Method	The BBFER is measured separately for each PLP. A Baseband Frame is erroneous, if an uncorrectable error has been discovered and indicated by the error flag by the BCH decoder. The parameter is either given as a ratio or as the number of erroneous BB frames per second.
Reference	Clause 5.1.6 of ETSI EN 302 769 [i.28].

12.2.13 IQ signal analysis

12.2.13.1 Introduction

The IQ analysis can be applied to single carriers of the OFDM signal as well as to groups of carriers. If a group of carriers is evaluated all received symbols of this group can be superimposed in order to obtain the respective measurement parameters or the constellation diagram.

The definitions for the I/Q related parameters are based on the following assumptions:

- a constellation diagram of M symbol points and K carriers under consideration with $0 < K \leq K_{MAX} +1$ and $K_{MAX} + 1$ is the total number of active OFDM carriers;
- a measurement sample of N data points, where N is sufficiently larger than $M \times K$ to deliver the wanted measurement accuracy; and
- the co-ordinates of each received data point j being $I_j + \delta I_j$, $Q_j + \delta Q_j$ where I and Q are the co-ordinates of the ideal symbol point and δI and δQ are the offsets forming the error vector of the data point (as long as the respective carrier is a "useful" one).

12.2.13.2 Modulation Error Ratio (MER)

Purpose	To provide a "figure of merit" analysis for L1 Signalling data [MER(L1p2)] and each PLP [MER(PLP), MER(Common-PLP)] of the DVB-C2 signal, typically at a transmitter output (for assessing the quality of the transmitted signal) or in a fixed location in a network.
Interface	Internal I/Q interface (test receiver or modulation analyser)
Method	The carrier frequency of the OFDM signal and the symbol timing are recovered. Origin offset of the centre carrier (e.g. caused by residual carrier or DC offset), Quadrature Error (QE) and Amplitude Imbalance are not corrected. A time record of N received symbol co-ordinate pairs $(\tilde{I}_j, \tilde{Q}_j)$ is captured. For each received symbol, a decision is made as to which symbol was transmitted. For the calculation of MER(PLP), the information derived from the LDPC decoder should be used to calculate the ideal symbol position transmitted (I_j, Q_j). The error vector is defined as the distance from the ideal position of the chosen symbol to the actual position of the received symbol. This distance can be expressed as a vector $(\delta I_j, \delta Q_j)$. The sum of the squares of the magnitudes of the ideal symbol vectors is divided by the sum of the squares of the magnitudes of the symbol error vectors. The result, expressed as a power ratio in dB, is defined as the MER. $$MER = 10 \times \log_{10} \left\{ \frac{\sum_{j=1}^{N} \left(I_j^2 + Q_j^2 \right)}{\sum_{j=1}^{N} \left(\delta I_j^2 + \delta Q_j^2 \right)} \right\} dB$$ NOTE: In OFDM systems, a symbol consists of many OFDM subcarriers. The calculation for one symbol means, that all subcarriers should be considered that are included in this symbol. Thus, a single symbol contains a plurality of error vectors. It should be reconsidered that MER is just one way of computing a "figure of merit" for a vector modulated signal. Another "figure of merit" calculation is Error Vector Magnitude (EVM). MER and EVM are closely related and one can generally be computed from the other. Error Vector Magnitude (EVM) is defined as: $$EVM_{RMS} = \sqrt{\left\{ \frac{\sum_{j=1}^{N} \left(\delta I_j^2 + \delta Q_j^2 \right)}{\sum_{j=1}^{N} \left(I_j^2 + Q_j^2 \right)} \right\}} \cdot 100\%$$ For both parameters, MER and EVM, the display of the values as a function of frequency or carrier number can be very helpful.

12.2.13.3 Amplitude Imbalance (AI)

Purpose	To separate the QAM distortions resulting from Amplitude Imbalance (AI) of the I and Q signal from all other kind of distortions.
Interface	Internal I/Q interface (test receiver or modulation analyser)
Method	Calculate the I and Q gain values v_I and v_Q from all points in a constellation diagram eliminating all other influences. Calculate Amplitude Imbalance (AI) from v_I and v_Q $$AI = \begin{cases} \left(\frac{v_I}{v_Q} - 1\right) \times 100\% & if\ v_I \geq v_Q \\ \left(1 - \frac{v_Q}{v_I}\right) \times 100\% & if\ v_Q > v_I \end{cases}$$ $$v_I = \frac{1}{M}\sum_{i=1}^{M}\frac{I_i + (\bar{d}_i)_I}{I_i}$$ $$(\bar{d}_i)_I = \frac{1}{N}\sum_{j=1}^{N}\delta I_j \qquad \text{(I - component of } d_i \text{ as given in clause 9.18.3 System Target Error)}$$ $$v_Q = \frac{1}{M}\sum_{i=1}^{M}\frac{Q_i + (\bar{d}_i)_Q}{Q_i}$$ $$(\bar{d}_i)_Q = \frac{1}{N}\sum_{j=1}^{N}\delta Q_j \qquad \text{(Q - component of } d_i \text{ as given in clause 9.18.3 System Target Error)}$$ $$(\bar{d}_i)_I + (\bar{d}_i)_Q = \bar{d}_i$$

12.2.13.4 Quadrature Error (QE)

Purpose	The phases of the two carriers feeding the I and Q modulators have to be orthogonal. If their phase difference is not 90° a typical distortion of the constellation diagram results.
Interface	Internal I/Q interface (test receiver or modulation analyser)
Method	Search for the constellation diagram error shown in Figure 12.3 and calculate the value of the phase difference $\Delta\varphi = \varphi_1 - \varphi_2$ after having eliminated all other influences and convert this into degrees: $QE = \frac{180°}{\pi} \times (\varphi_1 - \varphi_2) \quad [°]$

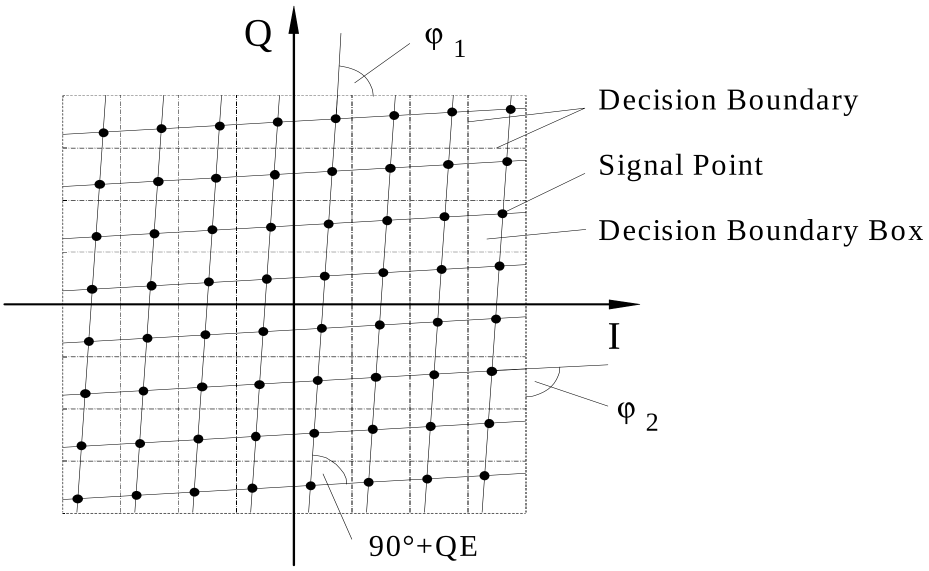


Figure 12.3: Distortion of constellation diagram resulting from I/Q Quadrature Error (QE)

12.2.14 L1 signalling error

Purpose	The consistency of the L1 data is essential for the successful decoding of the DVB-C2 signal by the receiver.
Interface	Baseband
Method	The receiver checks the CRC within the corresponding L1 signal.
Reference	Clause 8.3.2 of ETSI EN 302 769 [i.28]

12.2.15 Receiver Buffer Assumptions (RBA) validation test

Purpose	This test applies a test stream for an out-of-service test to the receiver to validate the proper functioning of the receiver buffer.
Interface	Baseband
Method	A DVB-C2 signal is applied to the input of a DVB-C2 receiver. This DVB-C2 signal contains a test stream (e.g. VV017,018,019 (M-PLP)). & ISSY The failure point is defined as: the Frame Error Rate or the rate of visible errors is below 1 frame error or visible error in 20 seconds..
Reference	n/a

NOTE: These test streams exercise the buffer but there are no specific tests defined or proposed yet to test buffer limits. Therefore, this clause is kept in the present document as a placeholder and can be revised if such test streams become available.

Annex A: General measurement methods

A.1 Introduction

It is recommended that manufacturers add the test mode described in this annex to certain professional grade cable and satellite broadcast equipment. This recommendation is relevant to equipment that implements the channel encoding schemes defined in ETSI EN 300 429 [i.6] (cable) and ETSI EN 300 421 [i.5] (satellite).

The purpose of the recommended test mode is to simplify out of service testing of systems and system components by making the channel encoder able to generate a known, fixed, repeating bit sequence of an essentially pseudo-random nature.

The central requirement is that when the channel encoder is in the test mode, the data entering the sync inversion/randomization function is a continuous repetition of one fixed TS packet. The fixed packet is defined as the four byte sequence 0x47, 0x1f, 0xff, 0x10, followed by 184 zero bytes (0 x 00). This form of data is a refinement of the *null TS packet* definition in ISO/IEC 13818-1 [i.1].

A.2 Null packet definition

This clause summarizes the null packet definition from ISO/IEC 13818-1 [i.1] and then describes how the definition has been extended for the purpose of the recommended test mode.

ISO/IEC 13818-1 [i.1] defines a null TS packet for the purposes of data rate stuffing.

Table A.1 shows the structure of a null TS packet using the method of describing bit stream syntax defined in clause 2.4.3.3 of ISO/IEC 13818-1 [i.1].

This description is derived from tables 2-3 Transport Header (TH) in ISO/IEC 13818-1 [i.1]. The abbreviation "bslbf" means "bit string, left bit first", and "uimsbf" means "unsigned integer, most significant bit first".

The column titled "Value", gives the bit sequence for the recommended null packet.

A null packet is defined by ISO/IEC 13818-1 [i.1] as having:

- payload_unit_start_indicator = "0";
- **PID** = 0x1FFF;
- **transport_scrambling_control** = "00";
- **adaptation_field_control** value = "01". This corresponds to the case "*no adaptation field, payload only*".

The remaining fields in the null packet that should be defined for testing purposes are:

- **transport_error_indicator** which is "0" unless the packet is corrupted. For testing purposes this bit is defined as "0" when the packet is generated;
- **transport_priority** which is not defined by ISO/IEC 13818-1 [i.1] for a null packet. For testing purposes this bit is defined as "0";
- **continuity_counter** which ISO/IEC 13818-1 [i.1] states is undefined for a null packet. For testing purposes this bit field is defined as "0000";
- d**ata_byte** which ISO/IEC 13818-1 [i.1] states may have any value in a null packet. For testing purposes this bit field is defined as "00000000".

Table A.1: Null TS packet definition

Syntax	No. of bits	Identifier	Value
null_transport_packet(){			
sync_byte	8	bslbf	"01000111"
transport_error_indicator	1	bslbf	"0"
payload_unit_start_indicator	1	bslbf	"0"
transport_priority	1	bslbf	"0"
PID	13	uimsbf	"1111111111111"
transport_scrambling_control	2	bslbf	"00"
adaptation_field_control	2	bslbf	"01"
continuity_counter	4	uimsbf	"0000"
for (I = 0;i<N;i++) {			
data_byte	8	bslbf	"00000000"
}			
}			

A.3 Description of the procedure for "Estimated Noise Margin" by applying statistical analysis on the constellation data

Instead of adding real noise to the received signal this method uses statistical analysis and an iterative search algorithm to estimate the added noise power to reach the critical BER.

1) Demodulate the signal to produce a statistically significant sequence of data records. Each record represents the state of the demodulated I and Q components at a decision instant.

2) Compute the average noise power as the mean square of the error vectors and calculate the estimated $S_{avg/}N_{avg}$ ratio.

$$SNR = 10 \times \log_{10} \left(\frac{\frac{1}{N} \sum_{j=1}^{N} \left(I_j^2 + Q_j^2 \right)}{\frac{1}{N} \sum_{j=1}^{N} \left(\sigma I_j^2 + \sigma Q_j^2 \right)} \right)$$

The σI_j and σQ_j are the error vector co-ordinates which represent the offset from the co-ordinates of the centre (mean value) of the actual received data for a specific constellation point, to the actual received data point j.

If only Gaussian noise is present as an impairment the "centre (mean value) of the actual received data for a specific constellation point" is identical to the ideal symbol point.

N is the number of data points in the measurement sample.

3) Compute the additional noise power N_{step} required to degrade the computed SNR by a certain amount. The value N_{step} is usually determined by the iterative optimization procedure which is used.

4) For each data record in the sample compute the distances d from the true position of the signal at the decision instant to each of the decision boundaries with adjacent cells. For each of the directions +I, -I, +Q, -Q that would cause a symbol error, convert the distance to the decision boundary into the number of standard deviations (k) of a normal distribution with a variance corresponding to the added noise power. The variance of the added noise power is:

$$\sigma^2 = N_{step}$$

and the normalized standard deviation corresponding to the distance d_{I+} is for example:

$$k_{I+} = \frac{\sigma}{d_{I+}}$$

5) Compute the probability Q_S of a symbol error for each distribution tail due to an erroneous state transition in the relevant direction.

$$Q_s(k) = \frac{1}{\sqrt{2\pi}} \int_k^{\infty} \exp\left(-\frac{x^2}{2}\right) dx$$

or

$$Q_s(k) = \frac{1}{2} erfc\left(\frac{k}{\sqrt{2}}\right)$$

6) Compute the number of bit errors that the erroneous state transition would cause and calculate the bit error probability Q_B. One symbol error may result in more than one bit error for transitions across either the I or Q axis. Sum the individual Q_B values and divide by the number of points in the sample to get the average probability of a bit error.

7) Repeat the steps 4 to 6 for incremental values of noise power until the critical BER is found and calculate the noise margin:

$$\text{Noise Margin } (dB) = 10 \times \log_{10}\left(1 + \frac{N_{added}}{N_{avg}}\right)$$

A.4 Set-up for RF phase noise measurements using a spectrum analyser

The noise performance of the carrier can be characterized as the ratio of the measured power in one noise sideband component, on a per hertz of bandwidth spectral density basis, to the total signal power:

$$\alpha(f_m) = 10 \times \log_{10}\left(\frac{power_density(one_sideband, phase_only)}{power_of_total_signal}\right)$$

in (dBc/Hz) and f_m is the frequency distance away from the carrier.

For this measurement it is assumed that contributions from amplitude modulation to the noise spectrum are negligible compared to those from frequency modulation and that ΔB, the measurement bandwidth, is much smaller than f_m. A spectrum analyser with a noise measurement option is able to measure the power within 1 Hz bandwidth. If this is not available the resolution bandwidth should be as small as possible and the video bandwidth has to be 10 or 20 times smaller in order to get sufficient averaging of the noise over time.

For example: carrier frequency: 36 MHz

f_m = 10 kHz
ΔB = Equivalent Noise Bandwidth (ENB) of the resolution bandwidth filter: 270 Hz
video bandwidth: 10 Hz or 30 Hz

NOTE 1: Spectrum analysers typically use near Gaussian filters for the resolution bandwidth with a 20 % tolerance. The Equivalent Noise Bandwidth (ENB) is equal to the bandwidth of the filter measured at -3,4 dB, (by actually measuring the filter of the spectrum analyser, the 20 % tolerance factor is eliminated).

Then the following conversion to 1 Hz bandwidth can be applied:

$$\alpha(f_m) \cong 10 \times \log_{10}\left(\frac{\text{noise_power_in_DB}}{\text{signal_power}}\right) - 10 \times \log_{10}\Delta B + 2{,}5 \text{ dB in [dBc/Hz]}$$

NOTE 2: The 2,5 term accounts for the correction of 1,05 dB due to narrowband envelope detection and the 1,45 dB due to the logarithmic amplifier.

Having measured $\alpha(f_m)$ for various values of f_m an estimation of equivalent peak phase deviation and frequency deviation is possible by using sinusoidal analogy:

$$\alpha(f_m) \cong 20 \times \log_{10}(\Delta\varphi_{rms}/\sqrt{2}) \qquad \text{in [dB/Hz]}$$

with $\Delta\varphi$ in [rad/Hz]

The square root of the sum of all noise densities within the frequency range of interest will give the equivalent RMS phase noise error vector in the I/Q plane.

An estimation can be done if the phase noise power slope may be approximated by the density function:

$$Y = a\frac{1}{f^b}[W/Hz]$$

with $b = \frac{slope[dB]_per_decade}{10}$ (b > 0) and

$$a = N_0 \times f_1^b \qquad \text{where } N_0 = 10^{\left(\frac{\alpha(f_1)}{10}\right)}$$

Then the total double-side-band phase noise power within the frequency range of interest (f_1,f_2) can be approximated by:

$$DSB-Phase-Noise = 2a\int_{f_1}^{f_2}\frac{1}{f^b}df = \frac{2a}{(b-1)}\left(\frac{1}{f_1^{(b-1)}} - \frac{1}{f_2^{(b-1)}}\right)$$

For the normalized RMS error vector (carrier = 1) it follows:

$$\text{RMS Quadrature Error Vector } = \sqrt{\frac{2a}{(b-1)}\left(\frac{1}{f_1^{(b-1)}} - \frac{1}{f_2^{(b-1)}}\right)} = \sigma_{ph}$$

$$\Delta\varphi_\sigma \cong \arctan \sigma_{ph}[rad] \qquad \text{(for carrier = 1)}$$

A.5 Amplitude, phase and impulse response of the channel

The amplitude, phase and impulse response can be derived from the equalizer tap coefficients. The use of a good equalizer that is designed to cope with the echo profile defined in clause B.14 is recommended to get accurate results in case of high linear distortions.

The capabilities to derive the channel response from the equalizer tap coefficients depend on the structure of the equalizer. Especially the channel response in the Nyquist slope of the signal cannot be measured exactly with a T-spaced equalizer.

A.6 Out of band emissions

The out of band emissions can be measured using a spectrum analyser. The resolution bandwidth should be low enough to detect peaks in the out of band spectrum. The video filter should be at least 10 times lower than the resolution bandwidth for sufficient averaging of the noise-like signal.

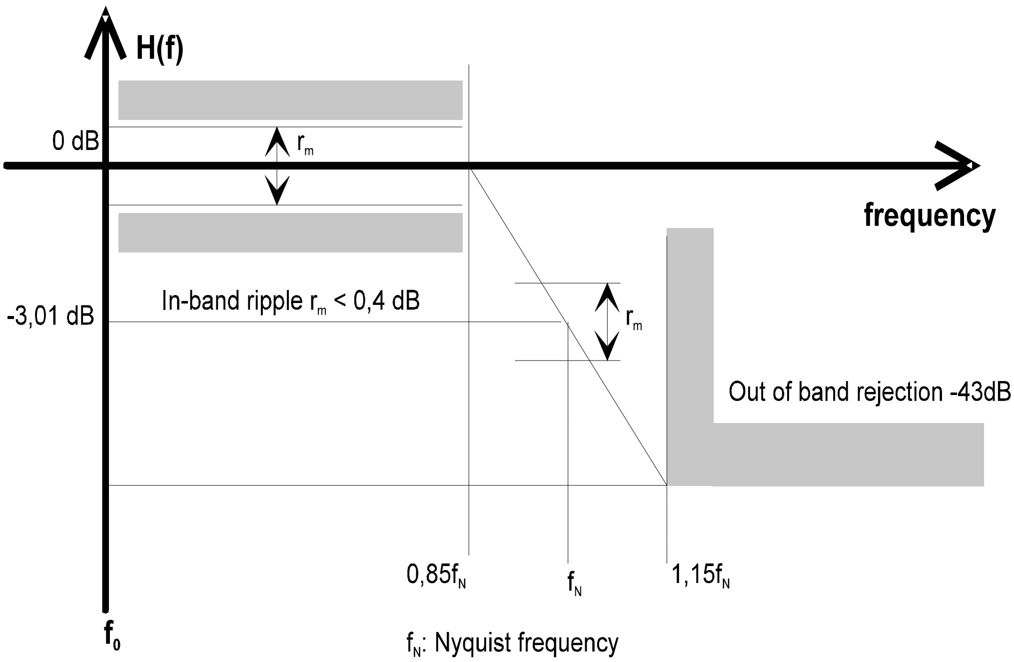


Figure A.1: Spectrum mask as defined in ETSI EN 300 429 [i.6]

Annex B:
Examples for test set-ups for satellite and cable systems

B.0 Introduction

Even if not demonstrated in the diagrams of this clause and also not mentioned in the explanations the receiver may be a part of the measurement device. In this case all the interfaces defined in figure 4.2 are internal ones, where the measurement device has access to.

B.1 System availability

See clause 6.1.

Because this measurement is based on the error_indicator_flag in the TS header set in any previous stage including the last stage of the transmission chain the signal at interface Z should be used.

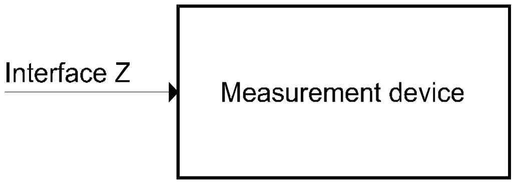


Figure B.1: Test set-up for system availability

B.2 Link availability

See clause 6.2.

This measurement monitors the performance of an individual link. Therefore the RS information should be created and be correct at the start point of the link. The measurement set-up may rely on the overload information coming from the RS decoder in the receiver at interface X or on the transport_error_indicator in the header of the TS packets at interface Z.

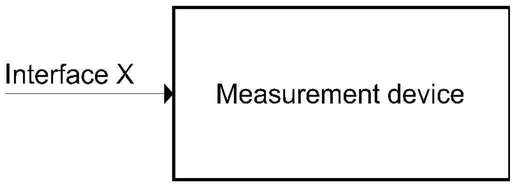


Figure B.2: Test set-up for link availability

B.3 BER before RS

B.3.0 General

See clause 6.3.

The measurement can be done as out service measurement or as in service measurement. In both cases the measurement time is an important parameter. This parameter should be selectable within a wide range by the user. Preferably the measurement should display the BER as a function of measurement time.

B.3.1 Out of service measurement

See clause 6.3.1.

When the BER is measured out of service Null packets as defined in clause A.2 should be created and transmitted to the receiving site. At the receiving site the signal at the interface W is compared against the pre-calculated values. The time window for the BER measurement should be selectable by the user.

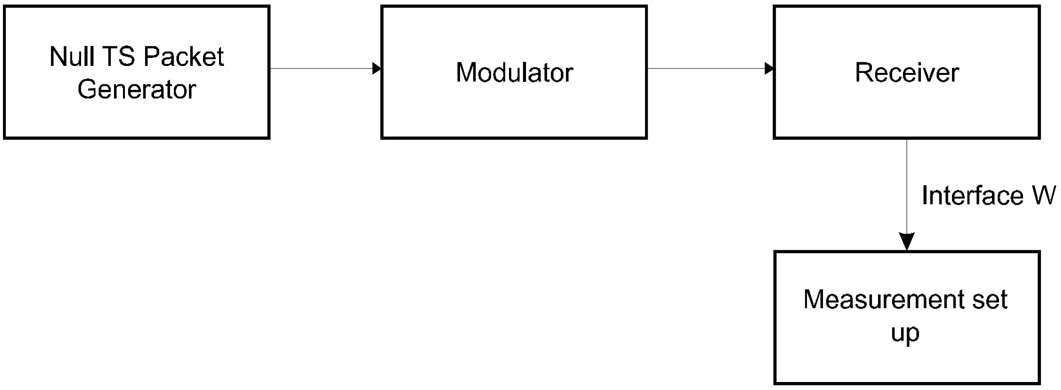


Figure B.3: Test set-up for out of service BER measurement before RS decoding

B.3.2 In service measurement

See clause 6.3.2.

In this case no special signal should be inserted in the transmitter. The measurement only relies on the results of the RS decoder. The measurement can be done by using the signals at the interfaces W and Z.

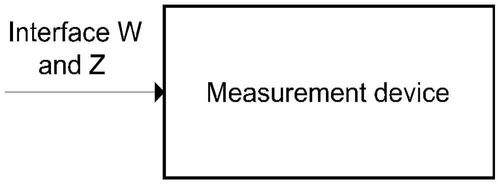


Figure B.4: Test set-up for out of service BER measurement before RS decoding

B.4 Event error logging

See clause 6.4.

This measurement relies on information coming from different parts of the receiver like tuner, RS decoder or a demultiplexer. Typically the receiver will be a part of the measurement device because it is not expected that all this information will be available at a standard receiver.

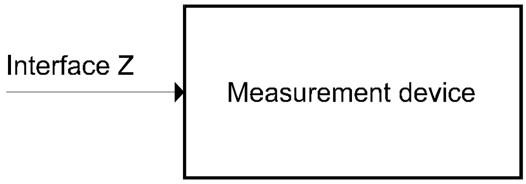


Figure B.5: Test set-up for event error logging

B.5 Transmitter symbol clock jitter and accuracy

See clause 6.5.

This measurement requires a symbol clock output at the modulator. To this interface an appropriate frequency counter and/or jitter and wander analyser can be connected.

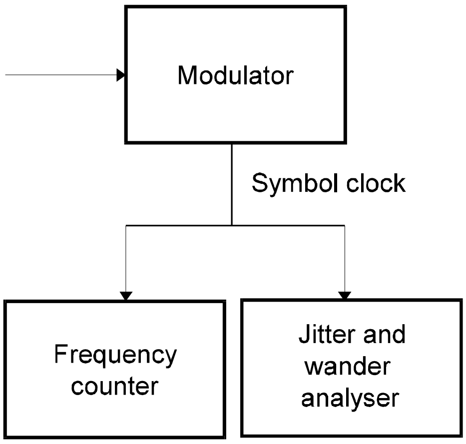


Figure B.6: Test set-up for transmitter symbol clock measurement

B.6 RF/IF signal power

See clause 6.6.

The signal power can be measured directly at the interfaces N or P or by using a calibrated splitter. If needed an appropriate filter should be used.

Figure B.7: Test set-up for RF/IF level measurement

B.7 Noise power

B.7.0 General

See clause 6.7.

Typically all the power present in a channel which is not part of the signal can be regarded as unwanted noise. It can be produced from different origination and be of the form of random noise (thermal), pseudo-random (digitally modulated interfering carriers) or periodic (Continuous Waves CW or narrowband interferences), the first two are called non-coherent and the periodic ones are termed as coherent.

B.7.1 Out of service measurement

For doing this measurement the carrier should be switched off. The measurements can be done at interface N (RF level) or at interface P (IF level). Noise level can be measured with a spectrum analyser or any other appropriate device. If a power metre is used the equivalent noise bandwidth should be taken into account. In this case of out-of-service measurement, all different types of noise are measured simultaneously, and the measured result can be termed as unwanted power.

Figure B.8: Test set-up or out-of-service noise level measurement

B.7.2 In service measurement

For the "in service measurement" eye diagrams or IQ constellation diagram derived from I and Q signals available at interface T should be employed. In this case of "in service measurement", it is possible to determine the type of the noise by applying the I/Q signal analysis (see clause 6.9).

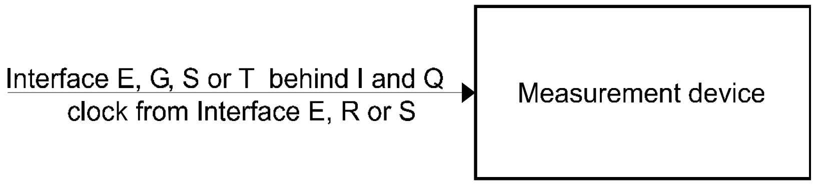


Figure B.9: Test set-up for in-service noise level measurement

B.8 BER after RS

See clause 6.8.

The set-up is equivalent to clause 6.3 BER before RS. The comparison is done after RS at interface Y.

B.9 I/Q signal analysis

See clause 6.9.

For this measurement eye diagrams or IQ constellation diagram derived from I and Q signals available at interface T should be employed.

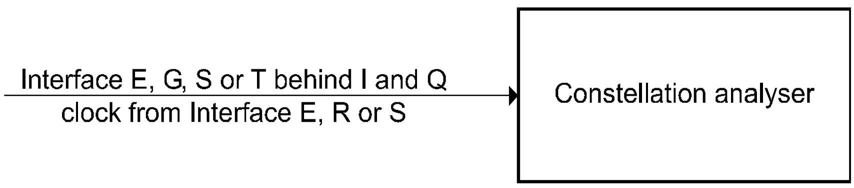


Figure B.10: Test set-up for I/Q signal analysis

B.10 Service data rate measurement

The set-up is equivalent to B.1 The measurement is based on the TS only.

B.11 Noise margin

B.11.0 General

See clause 7.1.

Purpose	To provide an indication of the reliability of the transmission channel (i.e. cable network), the noise margin measurement is a more useful measure of system operating margin than a direct BER measurement due to the steepness of the BER curve.
Interface	The reference interface for the noise injection is the RF interface (N, input of receiver). For practical implementation, other interfaces can be used, provided equivalence to the described set-up is ensured.
Test set-up	Figure B.11 shows the recommended test **set-up** for the measurement of noise margin.

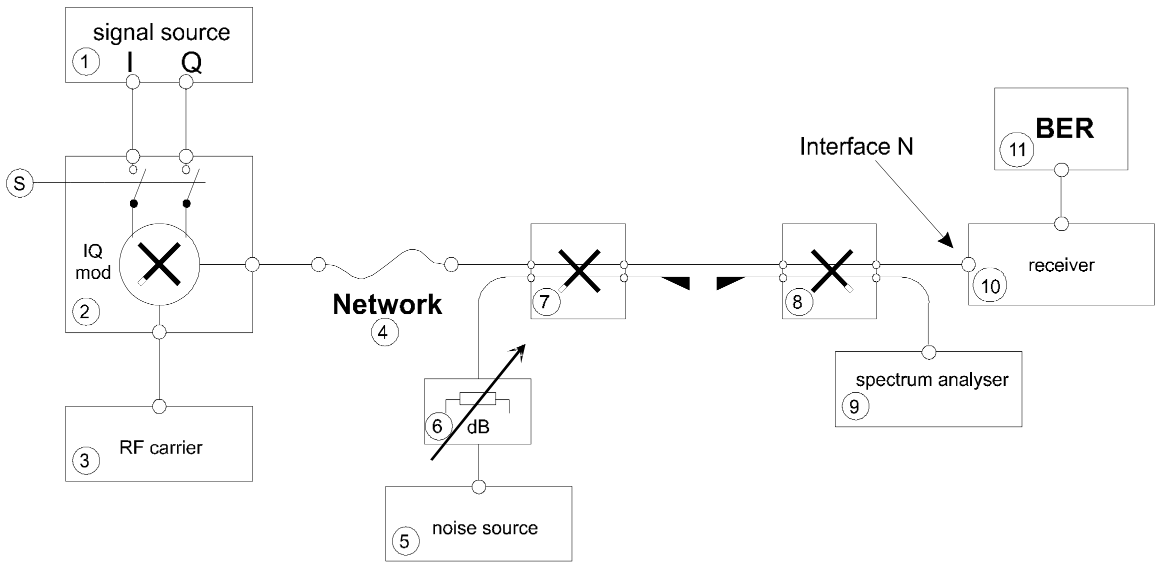


Figure B.11: Test Set-up for noise margin measurement

B.11.1 Recommended equipment

1 I/Q baseband signal source for 64 QAM;

S Switch (to switch off modulation);

2 I/Q modulator;

3 RF generator (see clause B.11.2 below, remark 1) (level and frequency adjustable);

4 cable network (see clause B.11.2, remark 2);

5 noise source (flat within the required measurement range) (see clause B.11.2, remark 3);

6 adjustable attenuator in 0,1 dB (max. 0,5 dB) steps;

7, 8 directive couplers (see clause B.11.2, remark 4);

9 spectrum analyser;

10 reference receiver with good equalizer (see clause B.11.2, remark 5);

11 counter of BER.

B.11.2 Remarks and precautions

1) Adjust RF carrier level so that non-linear distortion (i.e. CW, CSO, CTB) has no impact to BER measurement.

2) Pay attention to the amplitude response of the noise spectrum. If it is not white Gaussian spectrum (flat amplitude response) figure B.12 takes care to measure:

   a) If the effect produced by the thermal random noise is the wanted measurement, then take the measurement at the lowest level found in the wanted band (P4 in figure B.12), because it is the closest approximation to the random white thermal noise, then normalize the result to the full bandwidth of the channel, defined by the symbol rate x(1 + α).

   b) If the mean unwanted power is to be reported in the measurement, then integrate the spectrum with a suitable spectrum analyser or use a power metre with the appropriate filter as per clause B.7.1.

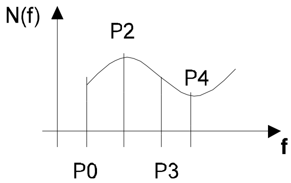


Figure B.12: Amplitude response of the noise spectrum

3) If a noise source with broadband output spectrum is used, avoid any affect to BER measurement by non-linear distortion due to an overload of the reference receiver's input amplifier stage.

4) Usual power splitters are allowed if sufficient matching at all ports is ensured for all measurement conditions (i.e. high attenuation in adjustable attenuator).

5) Influence of linear distortion of the cable network to the BER measurement should be negligible.

B.11.3 Measurement procedure

Step 1: Add noise to modulated cable network output until BER is 10^{-4}.

Step 2: Switch off modulation with (S);

Measure Noise power N1 (dBm) beside carrier (Δ f $\geq$ 0,5 MHz).

Step 3: Switch off noise source (5);

Measure Noise power N2 (dBm) beside carrier.

Step 4: Compute Noise Margin (NM):

$$NM = N1 - N2 \text{ (dB)}$$

NOTE: Due to step 2, the measurement of noise margin is to be done under out of service conditions.

B.12 Equivalent Noise Degradation (END)

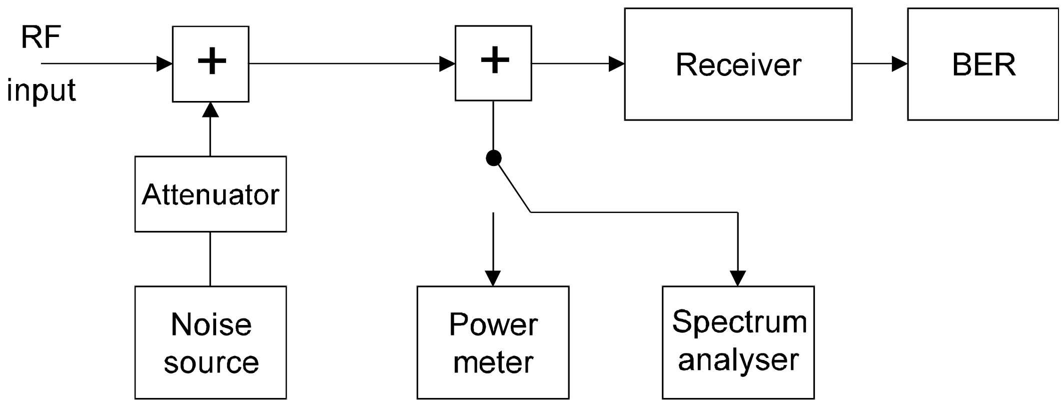


Figure B.13: Test set-up for END measurement

Procedure for the measurement of one point in figure B.13:

1) Measure the power of the DVB signal with a power metre. If this is not possible due to signals in the neighbouring channels, use a calibrated spectrum analyser.

2) Remove the wanted input signal and terminate the input.

3) Add noise to obtain the same level on the spectrum analyser. Now C/N = 0 dB.

4) Add the wanted input signal and increase the attenuation of the noise until a BER of 10^{-4} is measured. The value, for which the attenuation was increased, is the C/N for the given BER.

5) END is the difference between the measured C/N and the theoretical value of C/N for a BER of 10^{-4}.

Proposed settings for the spectrum analyser: RBW = 30 kHz, VBW < 300 Hz.

B.13 BER vs. E_b/N_0

The BER versus E_b/N_0 will be measured using the test set-up described above.

C/N measurements can be converted to E_b/N_0 using the following formula:

$$E_b/N_0 = C/N - 10\log_{10}(m)$$

NOTE: For consideration of FEC overhead, see also clauses 7.5, 8.2, G.5, G.6 and G.7.

B.14 Equalizer specification

High order modulations such as 64 QAM are very sensitive to distortions. The eye aperture is so small that any perturbation can seriously disturb the reception of the signal. In the case of the DVB modulation formats, this problem is increased by the low value of the roll-off factor (0,15). In a real network, if no special processing is carried out in the receiver, the eyes appear completely closed, and no synchronization is possible. This is why all cable receivers, professional or not, are equipped with equalizers.

Some of the most common impairments met on cable networks are echoes due to equipment impedance mismatching, or filtering effects. These impairments appear as perturbations of the frequency response (or impulse response) of the channel, and are corrected by the equalizer which is a form of adaptive filter. Equalizers are very efficient for linear distortions, but cannot combat those of a non-linear nature. They combat fixed frequency interference, which is equivalent to intermodulation products of analogue television signals. Equalizers have a large influence on the clock or carrier recovery systems, since these can use equalized signals. Thus the overall behaviour of the receiver depends on the performance of the equalizer.

Most of the measurements specified in the present document are carried out after equalization. The first reason is that the signal is too impaired before equalization to obtain meaningful measurement results. Moreover, as most of the distortion at that point would be removed in any practical receiver, such measurements may not be relevant. The consequence of this is that measurement results are dependent on the equalizer response. This also means that equipment with different equalizer architectures will have different performance characteristics. This situation is not acceptable, and has led to the specification of the equalizer.

The specification of an equalizer is a difficult task, because there is a large number of types of equalizer, due to the range of algorithms for the updating of coefficients, and the different filter architectures (time based, frequency based, recursive or non-recursive). In addition, the performance of future equipment should not be limited by any specification here. This is why a convenient solution is to specify the overall performance of the receiver as regards a perturbation typically corrected by the equalizer, specifically - echoes.

The specification has to be defined so that the reference perturbation does not affect the measurement. The minimum level of perturbation that the equalizer will have to correct can then be defined. A solution is to set the minimum level of an echo that will not degrade the equivalent noise degradation of the incoming signal of more than 1 dB. This measurement is carried out for the worst case phase shift of the echo.

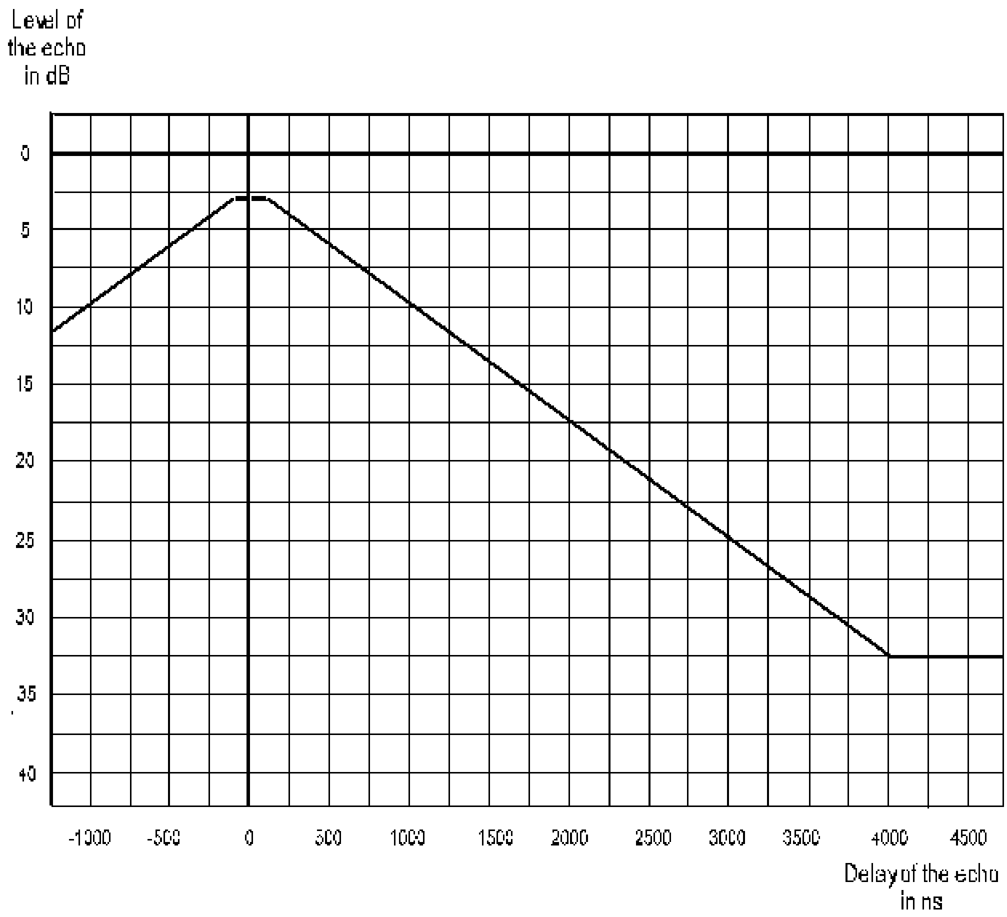


Figure B.14: Specification of an equalizer

Figure B.14 gives a possible equalizer specification which is subject to verifications in real systems.

In some cases, when the likely response of a consumer receiver to network signals is studied, it is appropriate to have an equalizer in the measurement equipment whose performance is close to that of the consumer receiver.

B.15 BER before Viterbi decoding

This measurement should be based on the I and Q signals at interface T. If an external measurement device is used the signals at interfaces T and V are needed. The set-up is equivalent to figure B.9.

B.16 Receive BER vs. E_b/N_0

The measurement is based on transmission of Null packets as defined in clause A.2. At the receiving site noise is added at one of the interfaces N, P or R. The spectrum analyser is used for checking that the normal noise level is well below the added noise. The measurement itself is done either within the receiver or at one of the interfaces T, V or Y depending whether BER before Viterbi, after Viterbi or after RS should be evaluated. In case of interface Y, RS decoding should be deactivated in order to reduce the duration of the measurement.

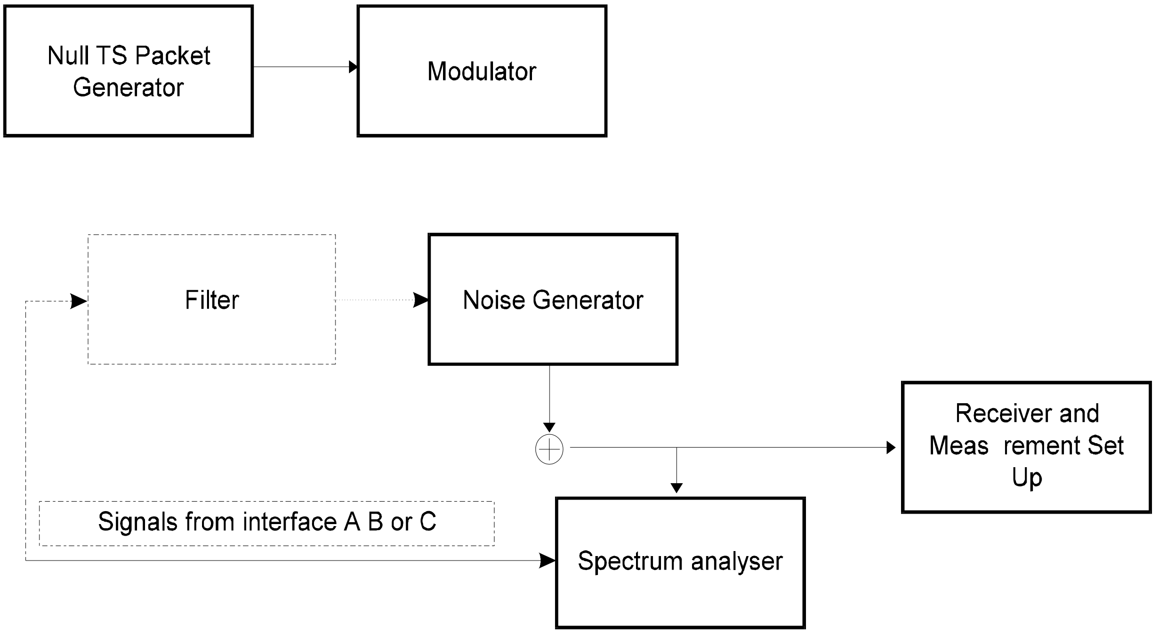


Figure B.15: Test set-up for BER vs. E_b/N_0 measurement

B.17 IF spectrum

The output of the modulator should be directly connected to the spectrum analyser. In addition it is also possible to use a (calibrated) splitter.

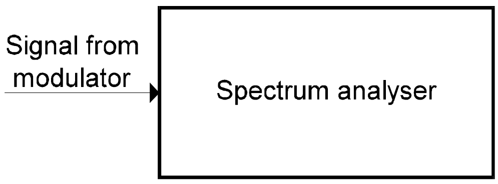


Figure B.16: Test set-up for IF spectrum measurement

Annex C: Measurement parameter definition

C.1 Definition of Vector Error Measures

Modulation Error Ratio (MER) is defined as:

$$MER = 10 \times \log_{10}\left\{\frac{\sum_{j=1}^{N}\left(I_j^2 + Q_j^2\right)}{\sum_{j=1}^{N}\left(\delta I_j^2 + \delta Q_j^2\right)}\right\} dB = 20 \times \log_{10}\left\{\frac{\sqrt{\sum_{j=1}^{N}\left(I_j^2 + Q_j^2\right)}}{\sqrt{\sum_{j=1}^{N}\left(\delta I_j^2 + \delta Q_j^2\right)}}\right\} dB$$

Error Vector Magnitude (EVM) is defined as:

$$EVM_{RMS} = \sqrt{\frac{\frac{1}{N}\sum_{j=1}^{N}\left(\delta I_j^2 + \delta Q_j^2\right)}{S_{\max}^2}} \times 100\%$$

Where I and Q are the ideal co-ordinates, δI and δQ are the errors in the received data points. N is the number of data points in the measurement sample. S_{max} is the magnitude of the vector to the outermost state of the constellation.

C.2 Comparison between MER and EVM

To compare the two measures it is easier to write them both as simple ratios, clearly the use of decibels and percentages is not central to the definition. Taking MER first, the simple voltage ratio (MER_V) is:

$$MER_V = \left\{\frac{\sqrt{\sum_{j=1}^{N}\left(I_j^2 + Q_j^2\right)}}{\sqrt{\sum_{j=1}^{N}\left(\delta I_j^2 + \delta Q_j^2\right)}}\right\}$$

and multiplying both numerator and denominator by $\sqrt{1/N}$ gives:

$$MER_V = \left\{\frac{\sqrt{\frac{1}{N}\sum_{j=1}^{N}\left(I_j^2 + Q_j^2\right)}}{\sqrt{\frac{1}{N}\sum_{j=1}^{N}\left(\delta I_j^2 + \delta Q_j^2\right)}}\right\}$$

Now looking at EVM as a simple voltage ratio (EVM_V), the following can be written:

$$EVM_V = \frac{\sqrt{\frac{1}{N}\sum_{j=1}^{N}\left(\delta I_j^2 + \delta Q_j^2\right)}}{S_{\max}}$$

EVM and MER are related such that:

$$MER_V \times EVM_V = \frac{\sqrt{\frac{1}{N}\sum_{j=1}^{N}\left(I_j^2 + Q_j^2\right)}}{S_{\max}} = \frac{1}{V} = S_{rms} / S_{\max}$$

or

$$EVM_V = \frac{1}{MER_V \times V}$$

If the peak to mean voltage ratio, V, is calculated over a large number of symbols (10 times the number of points in the constellation is adequate if the modulation is random) and each symbol has the same probability of occurrence then it is a constant for a given transmission system. The value tends to a limit which can be calculated by considering the peak to mean of all the constellation points. Table C.2 lists the peak- to-mean voltage ratios for the DVB constellation sizes.

Table C.1: Peak-to-mean ratios for the DVB constellation sizes

QAM format	Peak-to-mean voltage ratio (V)
16	1 341
32	1 303
64	1 527

C.3 Conclusions regarding MER and EVM

MER and EVM measure essentially the same quantity and easy conversion is possible between the two measures if the constellation is known. When expressed as simple voltage ratios MER_V is equal to the reciprocal of the product of EVM_V and the peak-to-mean voltage ratio for the constellation.

MER is the preferred measurement for the following reasons:

- The sensitivity of the measurement, the typical magnitude of measured values, and the units of measurement combine to give MER an immediate familiarity for those who have previous experience of C/N or SNR measurement.
- MER can be regarded as a form of Signal-to-Noise ratio measurement that will give an accurate indication of a receiver's ability to demodulate the signal, because it includes, not just Gaussian noise, but all other uncorrectable impairments of the received constellation as well.
- If the only significant impairment present in the signal is Gaussian noise then MER and SNR are equivalent.

Annex D: Exact values of BER vs. E_b/N_0 for DVB-C systems

Exact values of BER vs. E_b/N_0 for DVB-C systems (see figure 7.2).

Table D.1: Exact values of BER vs. E_b/N_0 for DVB-C systems

E_b/N_0	P_b
10	0,025 48
10,5	0,020 72
11	0,016 46
11,5	0,012 74
12	0,009 582
12,5	0,006 981
13	0,004 909
13,5	0,003 319
14	0,002 147
14,5	0,001 323
15	0,000 771 6
15,5	0,000 423 5
16	0,000 217 1
16,5	0,000 103 1
17	4,499e-005
17,5	1,783e-005
18	6,351e-006
18,5	2,006e-006
19	5,537e-007
19,5	1,314e-007
20	2,634e-008
20,5	4,365e-009
21	5,846e-010
21,5	6,166e-011
22	4,974e-012

This assumes that the relationship between BER and Symbol Error Rate (SER) is given by the formula:

$$BER = \frac{1}{m} \times SER$$

Annex E:
Examples for the terrestrial system test set-ups

E.0 Introduction

Due to the essential differences in the modulation method used for the terrestrial system some of the test methods are different with respect to those used for cable and/or satellite.

Even if not demonstrated in the diagrams of this clause and also not mentioned in the explanations, the receiver may be a part of the measurement device. In this case all the interfaces defined in figure 9.2 are internal ones, which the measurement device has access to.

E.1 RF frequency accuracy

E.1.0 General

See clause 9.1.

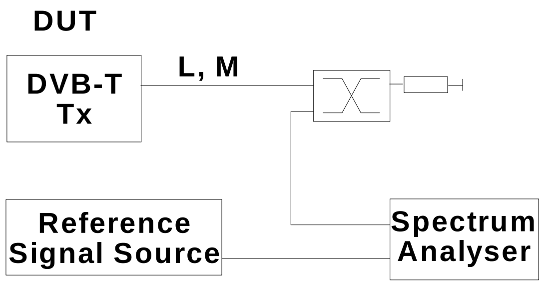


Figure E.1: RF frequency accuracy set-up

The measurement is to be done with a spectrum analyser. The signal can be picked up at interface L (IF) or M (RF), eventually by means of an aerial, or at interface N, if the received signal can be maintained stable enough for the measurement purposes, and applied to a spectrum analyser. Care should be taken at interfaces L or M not to overdrive the maximum allowed input signal for the spectrum analyser.

E.1.1 Frequency measurements in DVB-T (void)

Table E.1 (void)

Figure E.2: Examples of 8k centre channel measurement… (void)

E.1.2 Measurement in other cases (void)

Table E.2 (void)

Figure E.3: Examples of 8k carrier k = 48… (void)

E.1.3 Calculation of the external pilots frequency when they do not have continual phase. (void)

Table E.3 (void)

Figure E.4: Examples of 2k carrier k = 1 704… (void)

Figure E.5: Examples of 8k carrier k = 6 816… (void)

Figure E.6: Examples of 2k carrier k = 0… (void)

Figure E.7: Examples of 2k carrier k = 0… (void)

Table E.4 (void)

E.1.4 Measuring the symbol length and verifying the Guard Interval (void)

Table E.5 (void)

Figure E.8: Examples of 8k carrier k = 6 813… (void)

Table E.6 (void)

Table E.7 (void)

Table E.8 (void)

Figure E.9: Examples of 2k carrier k = 1 701… (void)

E.1.5 Measuring the occupied bandwidth, and calculation of the frequency spacing and sampling frequency

The occupied bandwidth depends directly from the frequency spacing and this from the sampling frequency.

If the frequency of the external pilots is known, see above on how to measure them, then the related values may be calculated as per table below. Denoting the outermost pilot frequencies as F_L and F_H appropriately the occupied bandwidth is OB = F_H _ F_L. The number of carriers K, and for 2k mode K-1 = 1 704 while for 8k mode K-1 = 6 816.

Table E.9

	Calculated value		Nominal value (8 MHz Channels)	
	8k mode	2k mode	8k mode	2k mode
Occupied bandwidth	$F_H - F_L$		*7,607142857142857142857142857142857142857… MHz*	
Frequency Spacing	$(F_H - F_L)/6\,816$	$(F_H - F_L)/1\,704$	*1 116,0714285…Hz*	*4 464,2857142…Hz*
Useful duration	$6\,816/(F_H - F_L)$	$1\,704/(F_H - F_L)$	896 µs	224 µs
Centre channel 1st IF	$(F_H - F_L) \times 4\,096/(K-1)$	$(F_H - F_L) \times 1\,024/(K-1)$	*4,571428571428571428571428571428571428571…MHz*	
Sampling Frequency	$(F_H - F_L) \times 16\,384/(K-1)$	$(F_H - F_L) \times 4\,096/(K-1)$	*18,285714285714285714285714285714285714285…MHz*	

NOTE: The long periodic decimals have been calculated using the Calculator facility from Windows, and have been left here as resulted from copying through the clipboard, as a matter of curiosity only.

Values in italics are approximate values.

Table E.10

	Nominal value (7 MHz Channels)		Nominal value (6 MHz Channels)	
	8k mode	2k mode	8k mode	2k mode
Occupied bandwidth	6,656250 MHz		*5,70535714285714285714285714285842… MHz*	
Frequency Spacing	976,5625 Hz	3 906,25 Hz	*837,053571428571…Hz*	*3 348,2142857142…Hz*
Useful duration	1 024 µs	256 µs	*1 194,666666… µs*	*298,666666… µs*
Centre channel 1st IF	4 MHz		*3,42857142857142857142857142857334…MHz*	
Sampling Frequency	16 MHz		*13,71428571428571428571428571142934…MHz*	

E.2 Selectivity

See clause 9.2.

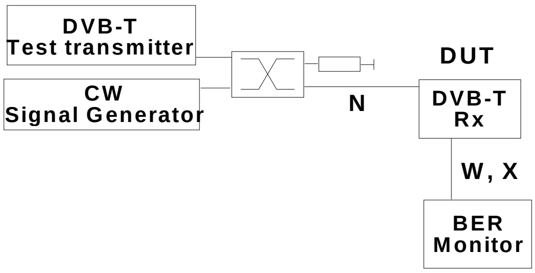


Figure E.10: Selectivity

E.3 AFC capture range

See clause 9.3.

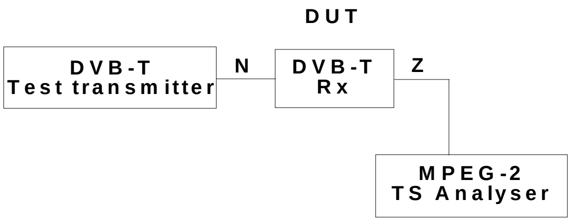


Figure E.11: AFC capture range

E.4 Phase noise of Local Oscillators (LO)

E.4.0 General

See clause 9.4.

The measurement can be done with a spectrum analyser. As the spectrum shape of the phase noise sidebands of any Local Oscillator (LO) used in the process of up/down conversion could be very different depending on factors such as the type of crystal cut, the filter of the PLL, the noise of the active devices involved, etc. it is not convenient to integrate the spectrum of a sideband to reflect a single measured number which could not have meaning at all.

However, samples at certain offsets of the oscillator signal could have more meaning, as indicated in clause 9.4. In each case of Common Phase Error (CPE) and Inter-Carrier Interference (ICI), 3 frequencies at each side of the oscillator signal should be measured. In order to make the measurement as accurate in frequency as possible, the spectrum analyser should be set to the minimum resolution filter available, and should be, at least, as low as 1 kHz for the 2 k system and 300 Hz for the 8 k system. In order to average the noise, the video filter should be activated with a value of at least 100 times narrower than the resolution filter used. The measured values should be normalized to a 1 Hz bandwidth.

Should the spectrum analyser used not have the 1 Hz normalization capability, it can be done manually with the following criterion:

For example: carrier frequency: 36 MHz
f_m = 10 kHz (represents any of the required offsets f_a, f_b or f_c)
ΔB = Equivalent Noise Bandwidth (ENB) of the resolution bandwidth filter: 270 Hz
video bandwidth: 10 Hz or 30 Hz

NOTE 1: The spectrum analysers typically use near Gaussian filters for the resolution bandwidth with a 20 % tolerance. The Equivalent Noise Bandwidth (ENB) is equal to the bandwidth of the filter measured at -3,4 dB, (by actually measuring the filter of the spectrum analyser, the 20 % tolerance factor is eliminated).

Then the following conversion to 1 Hz bandwidth can be applied:

$$Pn \cong (noise_power_in_\Delta B)dBm - 10\log_{10}\Delta B + 2{,}5dB \qquad \text{in [dBm/Hz]}$$

NOTE 2: The 2,5 dB term accounts for the correction of 1,05 dB due to narrowband envelope detection and the 1,45 dB due to the logarithmic amplifier.

E.4.1 Practical information on phase noise measurements

This example from the works of AC106 VALIDATE Project and taken from the DTG D book, shows a recommended mask for phase noise measurements that is valid for local oscillators and is considered to cover safe limits for both CPE and ICI phase errors in the **2k mode** of DVB-T. The following values are recommended.

Table E.11: Frequency offsets for phase noise measurements

	f_a	f_b	f_c	f_d
Frequency	10 Hz	100 Hz	3 kHz	1 MHz
Limits L_a to L_d	-55 dBc/Hz	-85 dBc/Hz	-85 dBc/Hz	-130 dBc/Hz

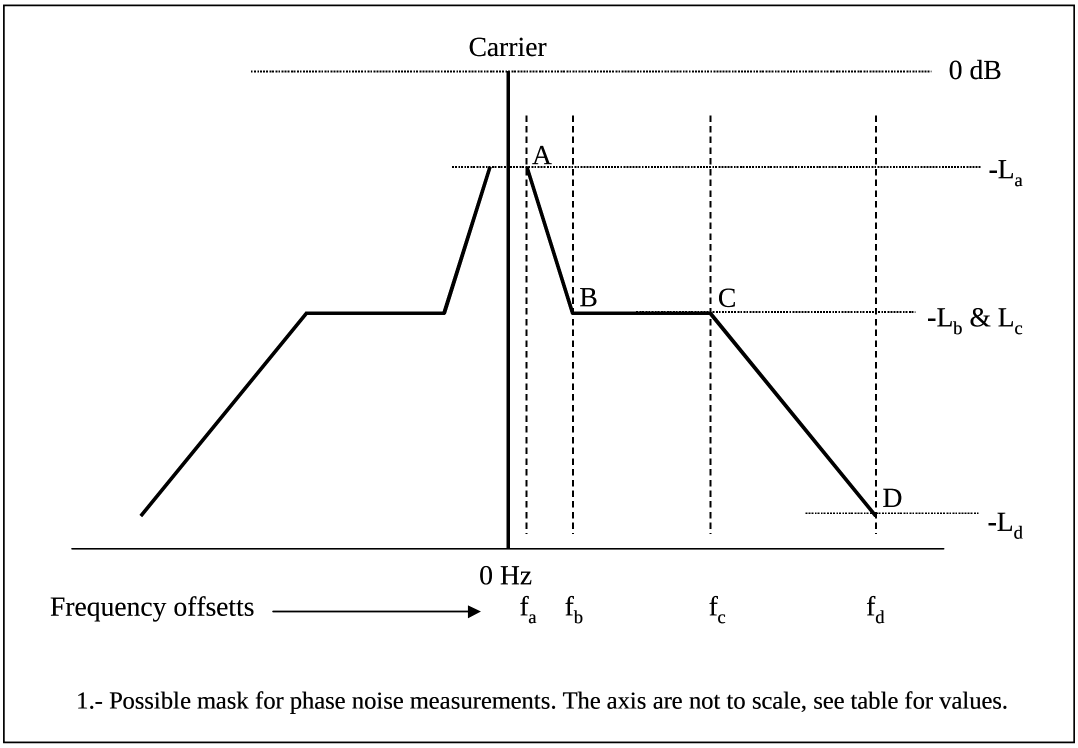


Figure E.12: Example for phase noise mask

The total phase noise in the signal is the cumulative effect of all local oscillators (L.O.) that are used in the signal path.

Clause A.4 can be seen for additional information on phase noise measurements.

E.5 RF/IF signal power

E.5.0 General

See clause 9.5.

The signal power can be measured directly at the interfaces K, L, M, N or P or by using a calibrated splitter. Care should be taken at interfaces L or M not to overdrive the maximum allowed input signal for the spectrum analyser or power metre.

The shoulders of the spectrum should not be accounted for in the measurement of power because they do not represent any useful power conveying information. The shoulders are unwanted results of the FFT process and also due mainly to non-linearity of the practical implementations.

E.5.1 Procedure 1 (power metre)

An spectrum analyser is used with an integrating routine which can measure the mean power along frequency slots covering the overall part of the spectrum to be measured (this capability is currently available in several spectrum analyser on the market). In this case the values to be supplied to such a routine or to be used if manual undertaken of the measurement is wanted are:

1) Centre frequency of the spectrum: if possible as calculated under measurement E.2;

2) Spectrum bandwidth of the signal: 7,61 MHz for an 8 MHz channel system.

E.5.2 Procedure 2 (spectrum analyser)

With the above considerations in mind, it would be very difficult to use an exact square filter for the measurement with a power sensor, however a good approximation should be obtained if a filter is used which can even take in account part of the shoulders in the measurement.

For measuring with a thermal power sensor such an appropriate filter should be used.

Figure E.13: Test set-up for RF/IF power measurement

E.6 Noise power

E.6.0 General

See clause 9.6.

Typically all the power present in a channel which is not part of the signal can be regarded as unwanted noise. It can be produced from different origination and be of the form of random noise (thermal), pseudo-random (digitally modulated interfering carriers) or periodic (Continuous Waves CW or narrowband interference), the first two are called non-coherent and the periodic ones are termed as coherent. In this measurement, all different types of noise are measured simultaneously, and the measured result can be termed as unwanted power.

For doing this measurement the signal should be switched off. The measurements can be done at interface N (RF level) or at interface P (IF level).

Noise level can be measured with a spectrum analyser or any other appropriate device. The same bandwidth considerations and methodology used in clause E.6 apply to this measurement in both cases, using a power metre and a spectrum analyser.

Figure E.14: Test set-up for out-of-service noise power measurement

E.6.1 Procedure 1

Exactly equal to the above preferred procedure for signal power, clause E.6, but understanding that the signal for this channel under measurement has been switched off.

E.6.2 Procedure 2

Using a power metre as in the alternate procedure above in clause E.6, using the same filter and with the channel signal off.

E.6.3 Procedure 3

If the noise floor in all bandwidth of interest is flat, it would be possible to measure the noise power at any frequency point inside the channel bandwidth and normalize the value to the nominal bandwidth of (n-1) × $f_{SPACING}$ (7,61 MHz for 8 MHz channels 6,66 MHz for 7 MHz channels).

If the spectrum analyser does not have normalization routine to the wanted bandwidth the following procedure can be used.

In order to average the noise, the video filter should be activated with a value of at least 100 times narrower than the resolution filter used, this resolution bandwidth filter should be chosen to be as wide as possible in order to average as much spectrum of the channel as possible, but not exceeding such bandwidth (e.g. 7,61 MHz), the equivalent noise bandwidth ΔB (MHz) of the filter should be known by the specifications given by the manufacturer, or measured following manufacturer indications. The noise power measured can be normalized to the wanted bandwidth using the following formulae:

$$\text{Noise power (dB)} = \text{Measured level (dB)} + 10 \log_{10} (7{,}61/\Delta B) + 2{,}5 \text{ dB} \qquad \text{(for 8 MHz channels)}$$

If the spectrum analyser has a routine to normalize to 1 Hz, (this use to include the 2,5 dB correction) but not able to normalize to the wanted bandwidth, the following conversion can be applied:

$$\text{Noise power (dB)} = \text{Measured level (dB/Hz)} + 10 \log_{10} (7{,}61 \times 10^6) =$$

$$\text{Measured level (dB/Hz)} + 68{,}8 \text{ dB} \qquad \text{(for 8 MHz channels)}$$

E.6.4 Measurement of noise with a spectrum analyser

Care should be taken when the measured noise has a display level close to the display level of instrument noise, (less than 10 dB), because an additional proximity factor should be applied. This is typically done automatically in some instruments available in the market.

If this is not available in the instrument, it is necessary to subtract a correction factor CF from the noise level measured, the following correction table can be used.

Table E.12: Correction Factor (CF) for measured noise level

D (dB)	CF (dB)
0,5	8,63
1	6,87
1,5	5,35
2	4,33
3,01	3,01
4	2,2
5	1,65
6	1,26
7	0,98
8	0,75
9	0,58
10	0,46

D is the distance in display level between the instrument noise (no signal applied to the input) and measured noise level (with no change in the settings).

Notice that below 2 dB of D, the reliability of the result after applying the CF is under question due to the uncertainty of the measurement and the corresponding big value of CF to be subtracted.

E.7 RF and IF spectrum

See clause 9.7.

To be defined after some practical experience is achieved.

E.8 Receiver sensitivity/dynamic range for a Gaussian channel

See clause 9.8.

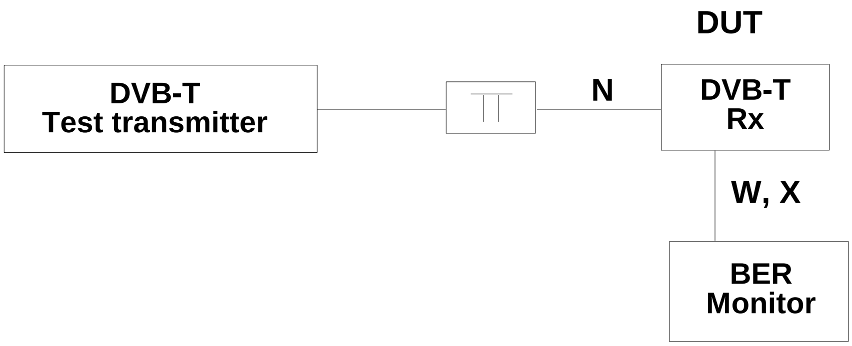


Figure E.15: Receiver sensitivity/dynamic range for a Gaussian channel

E.9 Equivalent Noise Degradation (END)

E.9.0 General

See clause 9.9.

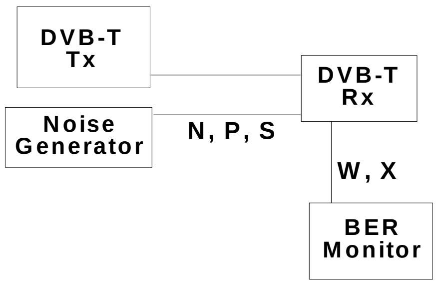


Figure E.16: Equivalent Noise Degradation (END)

All measurements of performance parameters are carried out by using a dummy load which provides a return loss for the wanted channel which is low enough not to influence the measurement.

E.9.1 Description of the measurement method for END

To improve the accuracy of the measurement, two independent noise sources are used. By this, the influence of the tolerance of the first attenuator is eliminated which could well be in the same magnitude as the wanted measurement result.

The following steps should be carried out to arrive at an accurate ENF value:

1) Connect the real DVB-T transmitter to the DVB-T receiver and add Gaussian noise, N_{cal}, to the point where the BER reaches a pre-determined value (e.g. 2 x 10^{-4} after Viterbi decoding). N_{cal} does not have to be measured. No channel noise, N_{ch}, should be added. The C/N at the input to the receiver (Interface C) is therefore
$C/(N_{tx} + N_{cal})$.

2) Replace the real DVB-T transmitter by the ideal one (disconnect N_{tx} in figure E.17). The C/N at Interface C is now somewhat higher (C/N_{cal}), since N_{tx} is no longer present. The BER is therefore now lower than the predetermined value.

3) Add Gaussian channel noise, N_{ch}, to the point where the BER has reached its predetermined value again. The C/N at interface C is now $C/(N_{ch} + N_{cal})$.

4) Measure the value of C/N_{ch} at Interface B.

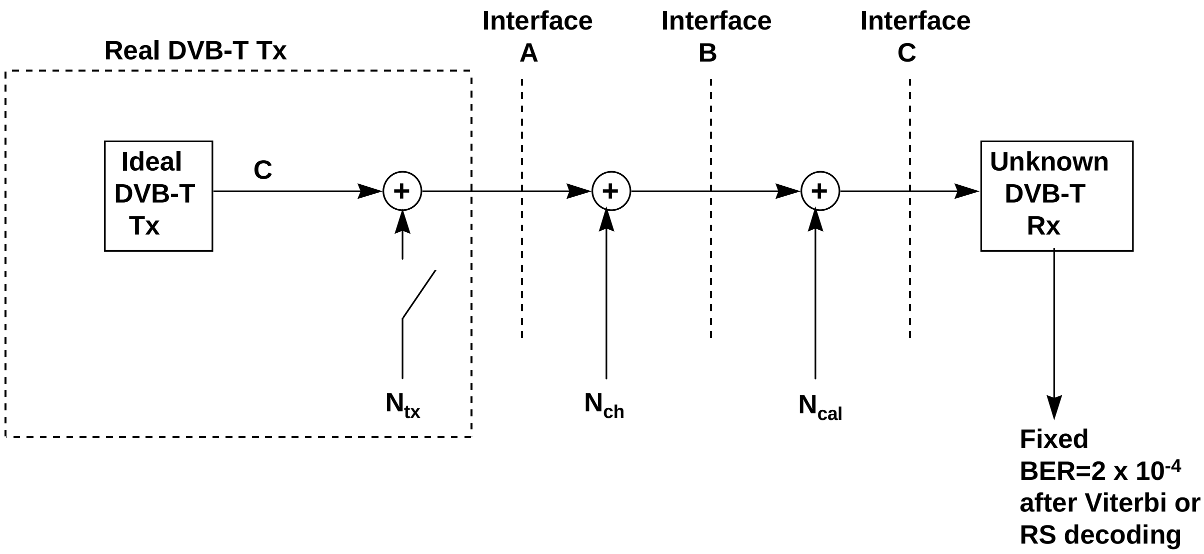


Figure E.17: ENF measurement scheme

Since both $C/(N_{tx} + N_{cal})$ and $C/(N_{ch} + N_{cal})$ lead to the same BER, N_{ch} can be identified with N_{tx} and be regarded as an estimate of N_{tx}.

The ENF is defined to be 10 $^{10}\log(N_{tx}/C)$. The estimated ENF value is similarly 10 $^{10}\log(N_{ch}/C)$

As long as all distortions of a DVB-T transmitter can be well approximated by the Gaussian noise, N_{tx}, the ENF measurement, as described above, should be completely independent of both the DVB-T mode and the receiver characteristics. For highest measurement accuracy the measurement should however preferably be done using the (non-hierarchical) mode requiring the highest C/N, i.e. 64-QAM R=7/8.

In practice, there might however be selective effects such as amplitude ripple and spurious signals within the useful bandwidth. In these cases the ENF will in principle be better (= a more negative value) when stronger code rates are used (such as R = 1/2 or 2/3) than when weaker codes are used (such as R = 5/6 or 7/8). Whether this difference is measurable or not remains to be seen. It is therefore recommendable to measure the ENF also for the other code rates. If there is negligible difference between the ENF figures for the different code rates, this will imply that there are few selective effects and/or that these effects can be well approximated by Gaussian noise. If however there is a significant difference in ENF figures this implies that the ENF (and hence END) is code rate dependent. In such a case the ENF value to be used (either by itself or for the calculated END) should preferably be the one measured with the same code rate as the DVB-T transmitter will be used with by the network operator.

E.9.2 Conversion method between ENF and END

Let $(C/N)_{min,\ theory}$ be the minimum C/N requirement for a DVB-T mode given by ETSI EN 300 744 [i.9].
Assume an implementation loss of 3,0 dB for all modes.
Let $X = (C/N)_{min,\ real}$ be the corresponding minimum required C/N for a DVB-T mode.

$$X = (C/N)_{min,\ real} = (C/N)_{min,\ theory} + 3{,}0\ \text{dB}$$

END can be calculated from ENF by the formula:

$$END = -10\ {}^{10}\log(10^{-X/10} - 10^{ENF/10}) - X$$

Example:

$$X = 19{,}5\ \text{dB (64QAM, R= 2/3)} \qquad ENF = -30{,}0\ \text{dB}$$

$$END = -10\ {}^{10}\log(10^{-19{,}5/10} - 10^{-30{,}0/10}) - 19{,}5\ \text{dB} = 0{,}41\ \text{dB}$$

E.10 Linearity characterization (shoulder attenuation)

E.10.0 General

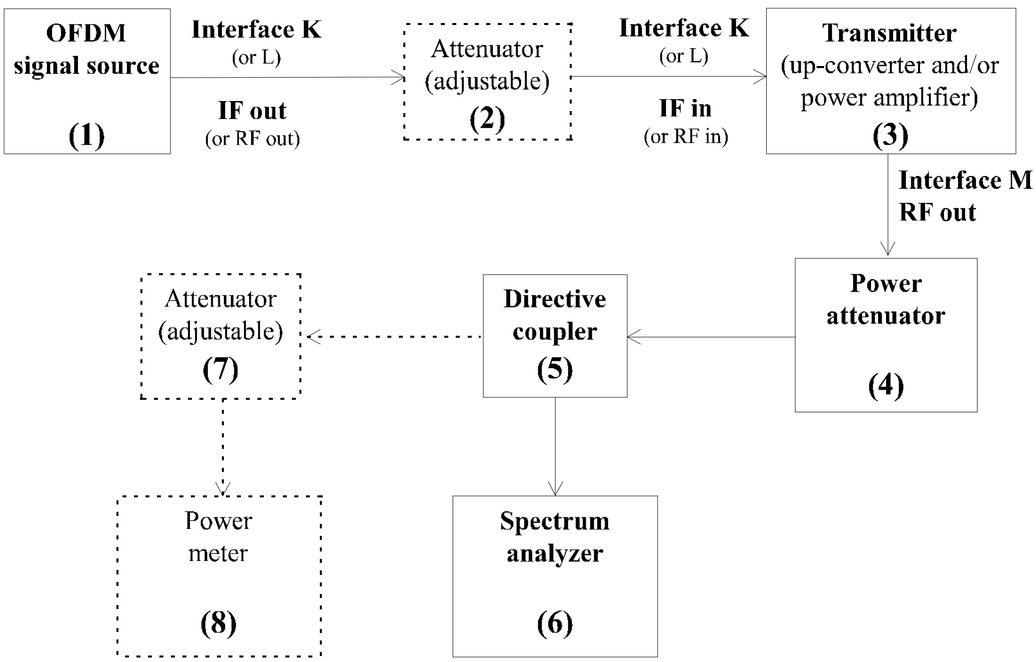


Figure E.18: Test set-up for "linearity characterization"

E.10.1 Equipment

(1) OFDM signal source (interface K or L of DVB-T transmitter);

(2) attenuator, possibly adjustable in 0,1 dB (max. 0,5 dB) steps. Optional, see clause E.10.2, remark (d);

(3) transmitter under measurement;

(4) power attenuator;

(5) directive coupler or attenuator, see clause E.10.2, remark (a);

(6) spectrum analyser;

(7) attenuator, possibly adjustable. Optional, see clause E.10.2, remark (c);

(8) power metre. Optional, see clause E.10.2, remark (a).

E.10.2 Remarks and precautions

a) Power metre (8) can be useful to verify and monitoring the output power of the transmitter (3) and for the calibration process. If power metre (8) is not available, the directive coupler (5) can be replaced by an opportune attenuator connected to the spectrum analyser (6).

b) Care should be taken in the choice of the power attenuator (4) in terms of max. admitted power.

c) Care should be taken in the choice of all attenuators (and directive coupler) to prevent damage to test-set equipment. For example, the function of the optional attenuator (7) is to protect the probe of the power metre. The attenuator (7) can also be useful for other measurements and, for example, be connected in a chain to the receiver.

d) Pay attention to the admitted power at the IF (or RF) input of the transmitter, in order to obtain a proper working point. Optional attenuator (2) can be used for this purpose.

E.10.3 Measurement procedure (example for UHF channel 47)

Step 1: Select the centre frequency of spectrum analyser in the middle of the UHF channel (i.e. 682 MHz for channel 47). Verify the output power level using an high resolution BW (3 MHz or 5 MHz) and compare with the value obtained by the power metre (if available).

Step 2: Select the centre frequency of spectrum analyser at the end of the UHF channel (i.e. 686 MHz for channel 47).

Step 3: Select an adequate span (for example 2 MHz).

Step 4: Select the resolution BW (10 kHz is adequate for 2 k and 8 k mode) and adjust levels. Video BW is of the same order.

Step 5: Measure the power level at 300 kHz and 700 kHz from upper edge of the DVB-T spectrum and proceed as indicated in clause 9.10. Last DVB-T carrier is at approximately +3,8 MHz from the centre of the UHF channel: then, for channel 47, the two measurement points are at 686,1 MHz and 686,5 MHz.

Step 6: Repeat steps from 2 to 5 for the lower edge of the spectrum.

Step 7: The worst case value of the upper and lower results is the "shoulder attenuation" (dB).

NOTE: The value obtained should be joined up with the used mode (2 k or 8 k) of the OFDM source.

If available, the "maximum-hold" function of the spectrum analyser can help to carry out the measurement.

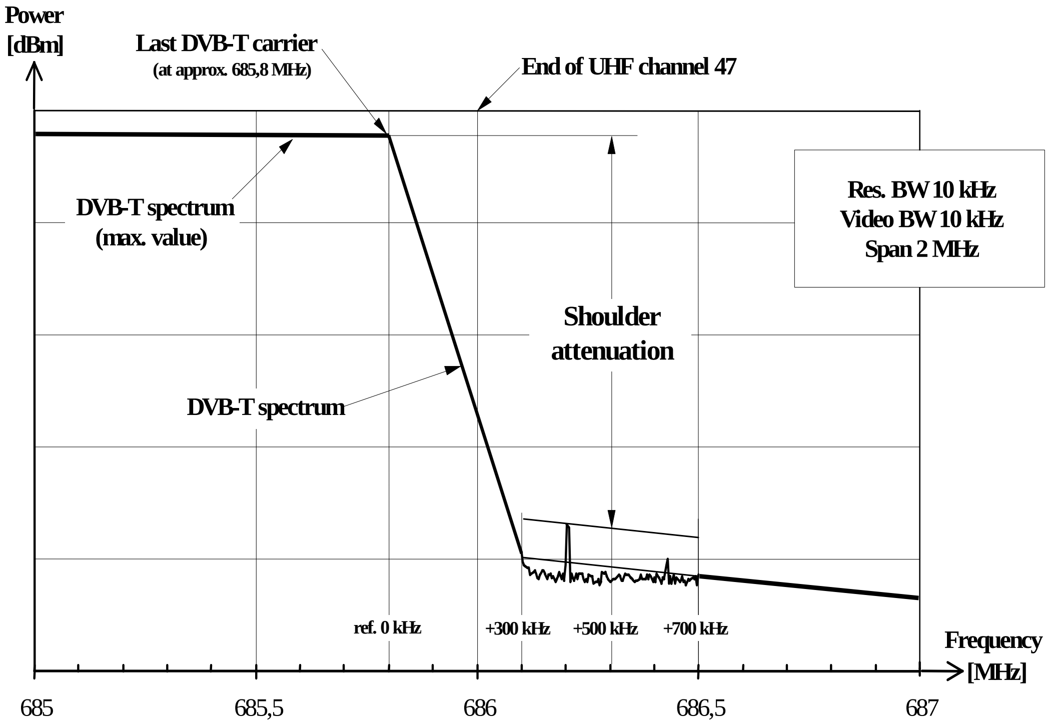


Figure E.19: Example with the upper edge of the DVB-T spectrum in UHF channel 47

E.11 Power efficiency

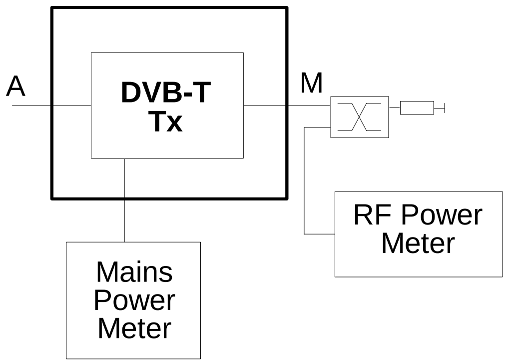


Figure E.20: Power efficiency

E.12 Coherent interferer

Connect a suitable spectrum analyser to interface N.

E.13 BER vs. C/N by variation of transmitter power

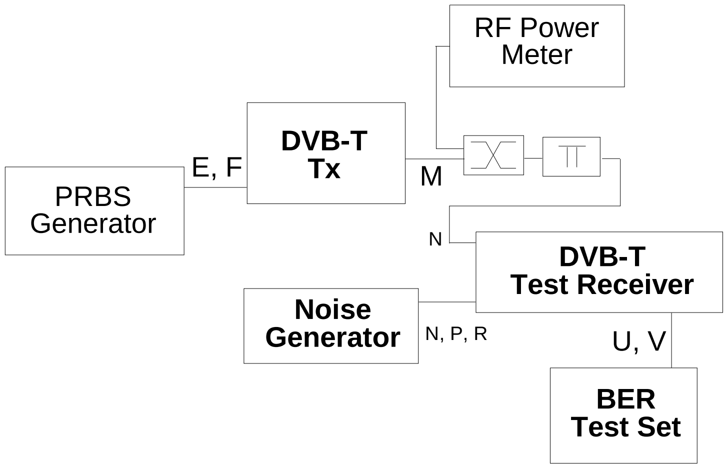


Figure E.21: BER vs. C/N by variation of transmitter power

Adjust signal level at receiver input to the same value for different Tx output power values by attenuator.

The results of this measurement can be put in diagrams, such as:

- BER vs. C/N for constant P_{out};
- BER vs. P_{out} for constant C/N;
- BER vs. P_{out} for constant noise power.

E.14 BER vs. C/N by variation of Gaussian noise power

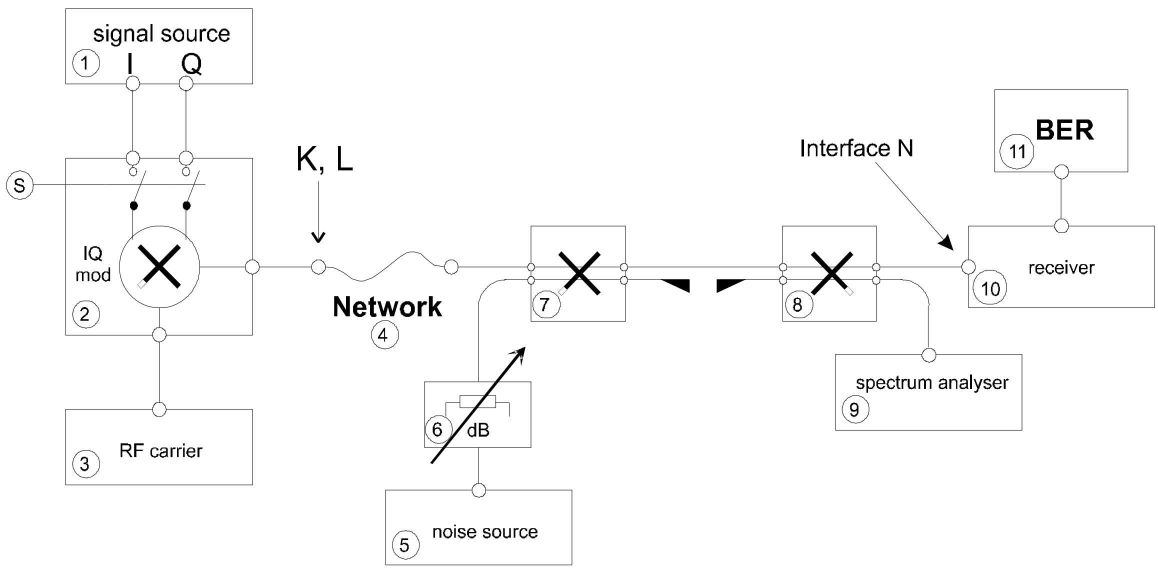


Figure E.22: BER vs. C/N by variation of Gaussian noise power

E.15 BER before Viterbi (inner) decoder

See clause 9.15.

NOTE: For the measurements described in clauses 9.15, 9.16, 9.17, 9.18 and 9.19 dedicated measurement instruments are envisaged.

E.16 Overall signal delay

Void.

Annex F: Specification of test signals of DVB-T modulator

F.1 Introduction

In order to compare simulated data within a DVB-T modem it is necessary to specify test points, signal formats and a subset of modes. The present document contains the specifications of how to do this. The present document should be accurate enough to enable comparison of simulated data at different points within the modulator.

F.2 Input signal

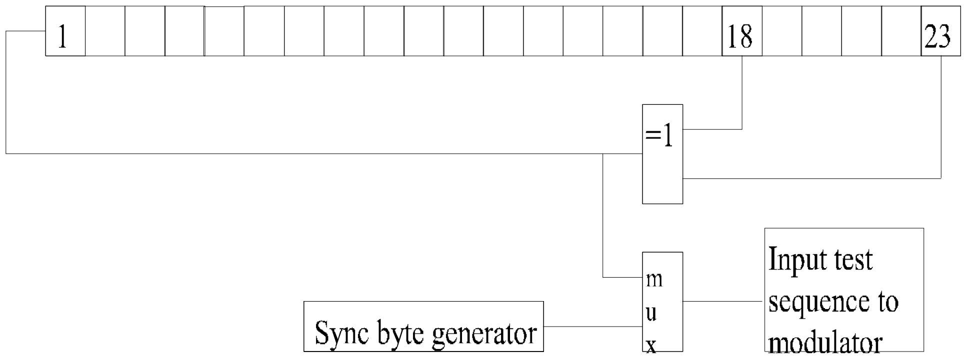


Figure F.1: Input test sequence generator for DVB-T modulator

The number of bits in a super-frame is depending on the actual DVB-T mode. The maximum number of Reed-Solomon/MPEG-2 packets in a super-frame is 5 292. This corresponds to 7 959 168 input bits that is shorter than a maximum length sequence of length 2^{23}-1 = 8 388 607. The input test sequence to the modulator can therefore be generated by a shift register of length 23 with suitable feedback. The generator polynomial should be $1 + x^{18} + x^{23}$. The PRBS data on every 188 byte is replaced by the sync byte content, 47 HEX. This means that during the sync bytes the PRBS generator should continue, but the source for the output is the sync byte generator instead of the PRBS generator. The input test sequence starts with a sync byte as the first eight bits, and the initialization word in the PRBS generator is "all ones". The PRBS generator is reset at the beginning of each super-frame. The test sequence at the beginning of each super-frame starts with:

0100 0111 0000 0000 0011 1110 0000 0000 0000 1111 1111 1100 (first byte is sync byte 47 HEX).

The corresponding HEX numbers are: 47 00 3E 00 0F FC.

There are up to eight possible phases of the energy dispersal with respect to the start of the super-frame. The first sync byte in the sequence, i.e. the first 8 bits should be inverted by the energy dispersal block. The length of the input signal can in principle be arbitrary. However, it is not meaningful to have a sequence shorter than one OFDM symbol. The maximum length will in practice be limited by the amount of data. Very large data files may be difficult to handle and interchange. One super-frame is therefore regarded as the longest sequence of interest. The outer interleaver will spread data across the super-frame boundaries**. The ambiguity in the output sequence caused by this is circumvented by using the second super-frame in the simulated sequence as the output signal.** This means that the simulator should produce one super-frame before useful data starts to appear at the output.

The file format for storing data allows for variable lengths of simulated data since the length indicator is contained in the header of the file. Simulations with different lengths can therefore be compared over the length of the shortest sequence.

F.3 Test modes

The file header in the test file contains information about the specific DVB-T mode used for the simulation. By reading this information a complete description of the set-up is obtained. In order to ease comparison of data and to reduce the amount of simulations necessary a set of "preferred modes" are defined. The preferred test mode for non-hierarchical transmission is:

Inner code rate: 2/3;

Modulation method: 64 QAM;

FFT size: 8 k;

Guard interval: 1/32.

For hierarchical transmission the preferred mode is:

Inner code rate HP: 2/3;

Inner code rate LP: 3/4;

Modulation method: QPSK in 64 QAM, $\alpha = 2$;

FFT size: 8 k;

Guard interval: 1/32.

F.4 Test points

The simulated data can be probed at different points within the modulator. Eight test points are defined, which are related to the interfaces described in figure 9.1:

1) at input (A);
2) after mux adaptation, energy dispersal (B);
3) after outer encoder (C);
4) after outer interleaver (D);
5) after inner encoder (E);
6) after inner interleaver (F);
7) after frame adaptation (H);
8) after guard interval insertion (J).

F.5 File format for interchange of simulated data

F.5.0 General

The file header as well as simulated data from the modem are stored as ASCII characters on files **with carriage return and line feed at the end of each line**. In order to interchange data it is important that the same file format be used by everyone. A file containing such data should have a header which has the following information:

- text string with a maximum of 80 characters (affiliation, time, place etc.);
- "printf" string used to store the data in the data section of the file;

- test point description;
- length of data buffer;
- constellation;
- hierarchy;
- code rate (code rate for HP);
- code rate LP (not applicable for non-hierarchical transmission modes);
- guard interval;
- transmission mode;
- simulated data (HEX or floating point).

The specification for each of these entries are given in tables F.1 to F.8.

F.5.1 Test point number

Table F.1: Test point number

Test point	Interface	Text contained in file header
1	A	at input
2	B	after MUX adaptation and energy dispersal
3	C	after outer coder
4	D	after outer interleaver
5	E	after inner coder
6	F	after inner interleaver
7	H	after frame adaptation
8	J	after guard interval insertion

F.5.2 Length of data buffer

The length indicator specifies the number of lines contained in the data section of the file which has two floating points or one two digit HEX on each line.

F.5.3 Bit ordering after inner interleaver

The signal at test point 4 after inner interleaver should contain data from one carrier on each line. The bit ordering should be according to table F.2.

Table F.2: Bit ordering in the signal representation at test point 4, after the inner interleaver

Modulation method	Bit ordering	Representation
QPSK	y_{0q} y_{1q}	2-digit HEX (00 to 03)
16 QAM	y_{0q} y_{1q} y_{2q} y_{3q}	2-digit HEX (00 to 0F)
64 QAM	y_{0q} y_{1q} y_{2q} y_{3q} y_{4q} y_{5q}	2-digit HEX (00 to 3F)

F.5.4 Carrier allocation

The signal contains 1 705 or 6 817 active carriers for the 2 k and 8 k modes respectively. In order to ease comparison of different data sets the allocation of these into the FFT bins should be specified. The signal is arranged such that it is centred around half the sampling frequency.

Table F.3: Carrier allocation

	FFT bins containing zeros	FFT bins containing active	FFT bins containing zeros
2 k mode	0 to 171	172 (K_{min}) to 1 876 (K_{max})	1 877 to 2 047
8 k mode	0 to 687	688 (K_{min}) to 7 504 (K_{max})	7 505 to 8 191

F.5.5 Scaling

At test point 7 (after frame adaptation) the data should be scaled such that: "Vector length of a boosted pilot" is equal to unity.

The gain factor through the IFFT should be equal to unity. This gain factor is defined as:

$$\eta = \sqrt{\frac{\sum_N (z\,z^*)}{\sum_N (x\,x^*)}}$$

where x are the complex numbers representing one complete OFDM symbol at the input of the IFFT including data carriers, pilots and null-carriers. And z is the complex signal for the corresponding OFDM symbol at the IFFT output before guard interval insertion. The number N is equal to the IFFT size (2 k or 8 k). The asterisk denotes complex conjugate. This ensures correct scaling of data at test point 8 (after guard interval insertion).

F.5.6 Constellation

The possible constellations are listed in table F.4. The file header should contain one of them.

Table F.4: Constellations

QPSK
16-QAM
64-QAM

F.5.7 Hierarchy

The hierarchical identifier specifies if hierarchical mode is on or off and also the alpha value in case hierarchical mode is on. For non-hierarchical transmission alpha is set to one. Table F.5 contains the possible choices and the file header should contain one of them.

Table F.5 Hierarchical identifier

Non-hierarchical, alpha = 1
Hierarchical, alpha = 1
Hierarchical, alpha = 2
Hierarchical, alpha = 4

F.5.8 Code rate LP and HP

The code rate identifier specifies the code rate for the LP and HP streams. Table F.6 contains the possible choices and the file header should contain one of them.

Table F.6: Code rate identifier

Code rate identifier
1/2
2/3
3/4
5/6
7/8

F.5.9 Guard interval

Table F.7 contains the possible choices for the guard interval and the file header should contain one of them.

Table F.7: Guard interval identifier

Guard interval identifier
1/32
1/16
1/8
1/4

F.5.10 Transmission mode

The transmission mode can be either 2 k or 8 k. Table F.8 contains the possible choices and the file header should contain one of them.

Table F.8: Transmission mode identifier

Transmission mode identifier
2 048
8 192

F.5.11 Data format

The data at test point 1 to 6 are written to file using 2-digit HEX numbers with "printf" string % X\n.

At test point 7 and 8 each line in the file contains real and imaginary parts with at least 6 significant decimal digits each. The real and imaginary parts and separated by at least 2 spaces. The data is written to file using "printf" with % e\n.

F.5.12 Example

This is an example of a print-out of a file containing the data sequence at the input for the preferred mode for non-hierarchical transmission. The text in parenthesis is just for explanation and should not be contained in the file.

Stockholm, May 22, 1996, example of input data. Preferred non-hierarchical mode:

%X\n (Data stored in HEX format);

at input (Data at test point 1 at modulator input);

758 016 (One super-frame of data);

64-QAM (Constellation 64 QAM);

non-hierarchical, alpha = 1 (Non-hierarchical transmission);

2/3 (2/3 inner code rate);

0 (Don't care. Code rate LP);

1/32 (Guard interval = 1/32);

8 192 (8 k IFFT size);

47 (First data byte is sync byte 47 HEX);

00 (Rest of data).

Annex G: Theoretical background information on measurement techniques

G.0 Introduction

This informative annex presents a review of the theoretical background to the measurement techniques recommended in the present document. It is an attempt to gather the most relevant background information into one location, particularly for the benefit of engineers and technicians who are new to digital modulation techniques. It is hoped that it will provide a working knowledge of the theoretical and practical issues, particularly the potential sources of ambiguity and error, to help users of the present document make valid, accurate and repeatable measurements.

G.1 Overview

The basic purpose of a digital transmission system is to transfer data from A to B with as few errors as possible. It follows that the fundamental measure of system quality is the transmission error rate.

The transmission error rate is usually measured as the Bit Error Rate (BER), however it can also be informative to consider the error rate of other transmission elements such as bytes, MPEG packets, or m-bit modulation symbols. In practice, although a certain guaranteed minimum BER performance may be a system implementation goal, the system BER alone is not a particularly informative measurement.

The most important figure of merit for any digital transmission system is the BER expressed as a function of the ratio of wanted information power to unwanted interference power (C/N). This is underlined by the fact that most of the measurements in the present document are built around this central theme of BER vs. C/N (or, equivalently, BER vs. E_b/N_0).

There are measurements of the individual elements (power and BER measurements). There are measurements of the difference between theoretical and ideal performance (margin and degradation measurements). There are measurements intended to help identify the sources of transmission errors (interference, spectrum, jitter and I/Q measurements). There are measurements for monitoring the consequences of transmission errors at the system level (availability, error event logging).

G.2 RF/IF power ("carrier")

When describing the Quadrature Amplitude Modulated (QAM) signals employed by DVB-C or the Quadrature Phase Shift Keying (QPSK) signals employed by DVB-S, it is common to refer to the modulated RF/IF signal as "carrier" (C), mainly to distinguish it from "signal" (S) which is generally used to refer to the baseband demodulated signal.

Strictly, it is incorrect to describe this signal as "carrier" because QAM and QPSK (which is equivalent to 4-state QAM) are suppressed carrier modulation schemes. For OFDM, with thousands of suppressed carriers and assorted pilot tones, the label "carrier" is even more inappropriate. This is why deliberately the expression "wanted information power" is used in the clause above, and why the parameter is referred to as "RF/IF power" in the present document.

However, it is clear that engineers will continue to use "carrier" as a convenient shorthand for this parameter, particularly when talking about the "carrier"-to-noise ratio. It seems futile to attempt to change this, so instead it is clearly defined what is meant by "carrier" in this context. Carrier, more accurately called RF/IF power, is the total power of the modulated RF/IF signal as would be measured by a thermal power sensor in the absence of any other signals (including noise).

For DVB compliant systems the QAM/QPSK passband spectrum is shaped by root raised cosine filtering with a roll-off factor alpha (α) of 0,15 for DVB-C systems, or 0,35 for DVB-S systems. For an ideal QAM/QPSK system this means that all the RF/IF power will lie in the frequency band:

$$BW_{OCC\,(QAM)} = f_C \pm (1+\alpha)\times\frac{f_S}{2} \tag{G.1}$$

Equation G.1 defines the **occupied bandwidth** of the signal, where f_C is the carrier frequency, f_S is the symbol rate of the modulation, and α is the filter roll-off factor. RF/IF power (or "carrier") is the total power in this "rectangular" bandwidth, that is, with no further filtering applied.

For OFDM systems the definition of occupied bandwidth is expressed differently because of the radically different modulation technique, however the principle is very similar. The OFDM "shoulders" are not considered to be wanted information power, and are not included in the RF/IF power calculation, even though the power does actually come out of the transmitter:

$$BW_{OCC\,(OFDM)} = n\times f_{SPACING} \tag{G.2}$$

where n = 6 817 (8 k mode) or 1 705 (2 k mode) and $f_{SPACING}$ = 1 116 Hz (8 k mode) or 4 464 Hz (2 k mode).

In a real multi-signal system (e.g. a live CATV network) measurement of the RF/IF power in a single channel requires a frequency selective technique. This could employ a thermal power metre preceded by a suitably calibrated channel filter, a spectrum analyser with band power measurement capability, or a measuring receiver. Depending on the measurement technique a filter may be required to exclude the "shoulders" of a single OFDM signal.

G.3 Noise level

The noise level is the unwanted interference power present in the system when the wanted information power is removed. This is a less bounded quantity than the RF/IF power because there is no definitively "correct" bandwidth over which to measure the noise. The choice is to some extent arbitrary, but the "top three" choices are probably:

1) ***Channel bandwidth:*** In a channel based system such as a CATV network you could choose the channel bandwidth, for example 8 MHz, as the system noise bandwidth. This is considered by the DVB-MG to be inappropriate for C/N measurements in digital TV systems. It will result in misleadingly poor C/N ratios when the modulation symbol rate is low relative to the available channel bandwidth. It unnecessarily complicates conversion between C/N measurements made "*in the channel* " and "*in the receiver* " by introducing symbol rate dependent correction factors.

2) ***Symbol rate:*** For digital modulation employing Nyquist filtering split equally between the transmitter and receiver, the noise bandwidth of the receiver equals the symbol rate. This is considered by the DVB-MG to be appropriate for "*in the receiver* " C/N measurements of digital TV systems since this reflects the amount of noise entering the receiver independent of symbol rate.

3) ***The occupied bandwidth:*** For digital modulation employing Nyquist filtering the occupied bandwidth of the modulated signal is $(1 + \alpha) \times f_S$. This is considered by the DVB-MG to be appropriate for "*in the channel* " C/N measurements of digital TV systems since it exactly covers the transmitted spectrum, independent of symbol rate.

The DVB-MG have chosen **occupied bandwidth,** as defined by equation G.1, as the standard definition of noise bandwidth in DVB-C and DVB-S systems. This is primarily because "*in the channel* " C/N is considered to be the fundamental measurement, but also because a simple correction factor can be applied to determine the equivalent "*in the receiver* " C/N value.

The other possibility that should be mentioned is to assume that the noise power is evenly distributed across the frequency spectrum of interest and so can be described by a single noise power density value (N_0) which is the noise power present in a 1 Hz bandwidth. From this, the noise power present in any given system noise power bandwidth (BW_{SYS}) can be obtained by simple multiplication:

$$N = N_0 \times BW_{SYS} \tag{G.3}$$

By talking in terms of N_0 there is no need to define a noise bandwidth, but an assumption is made that the noise power spectrum is flat across the bandwidth of interest.

G.4 Energy-per-bit (E_b)

Trying to commission a DVB system against tight deadlines, Energy-per-bit (E_b) seems to be a rather academic concept, particularly since the directly measurable quantity is RF power.

However, it is useful to understand E_b, even if only to avoid confusion when it appears in technical specifications or discussions. Historically, use of E_b arises from information theory and as part of an academic desire to normalize the performance of different modulation formats and coding schemes for comparative purposes.

The Energy-per-bit is the energy expended in transmitting each single bit of information. E_b is of little practical use on its own, it is most useful in the context of a graph of BER vs. the E_b/N_0 ratio - the well-known "waterfall curve" (see figures G.1 and G.2).

By normalizing to an E_b/N_0 ratio on the X axis, the relative performance of various complexities of digital modulation and channel coding can be compared because the scaling effects of actual signal and noise powers, number of bits-per-symbol and symbol rate are removed. It is then simply a case of comparing the bit error probability for a given ratio.

Energy-per-bit can be easily translated to carrier power. Power is energy-per-second. Which can be expanded to energy-per-bit, times bits-per-symbol, times symbols-per-second. Expressed algebraically it gives:

$$C = E_b \times \log_2(M) \times f_S \tag{G.4}$$

G.5 C/N ratio and E_b/N_0 ratio

The parameters that can be directly measured are RF/IF or "carrier" power (C) and noise power in a certain bandwidth (N). From these measurements the C/N ratio can be computed.

With the equations above, knowledge of the other parameters (e.g. f_S) and a little algebra it is also possible to arrive at an equivalent E_b/N_0 ratio.

G.6 Practical application of the measurements

At this point it seems that C/N (or E_b/N_0) is defined, and indeed it is from an algebraic perspective.

However, there is scope for endless confusion in applying these simple formulae unless the user is very clear about where the C/N or E_b/N_0 ratio is being measured, and what values are being used for the subordinate parameters, most particularly the system noise bandwidth.

C/N (or E_b/N_0) can be measured "*in the channel*" or "*in the receiver*". The meaning of "*in the channel*" is fairly self-evident, "*in the receiver*" may need further explanation.

There are typically three filtering processes present in a receiver. The first (which is optional) is a relatively broadband tuneable pre-selection simply to reduce the power presented to the receiver RF front-end. The second, usually applied at an IF, is a high-order bandpass channel selection filter to extract the desired signal with (ideally) no modification of the signal spectrum. The third is the root-raised cosine Nyquist filtering, commonly implemented in the low pass filters following the I/Q demodulation.

For theoretical simplicity it is assumed that the receiver's bandwidth and band shape are defined totally by the low-pass root-raised cosine filters because the intended purpose of the other RF/IF filters is only signal pre-selection. So the receiver can be modelled as a broadband receiver with a root-raised cosine passband filter followed by I/Q demodulation.

With this in mind, "*in the receiver*" can be seen to mean "*after the bandwidth and band shape modifying effects of the receiver Nyquist filters has been taken into account*".

Whether artificially generating a specific C/N ratio or just measuring the existing C/N ratio it is important to understand the difference between the "*in the channel*" and "*in the receiver*" nodes.

On a more practical note, graphing the BER performance of a receiver versus E_b/N_0 removes the ambiguity introduced by varying noise bandwidth. If the "*in the channel*" E_b value is used then a certain BER curve is derived, if the slightly lower "*in the receiver*" E_b value is used then the E_b/N_0 ratio is slightly poorer for the same BER, the curve moves to the left (closer to the theoretical curve) and the implementation loss decreases because the loss due to the receive filters is not included. An example may help to explain this.

G.7 Example

Creation of a signal with a specific C/N ratio in order to test the performance of an Integrated Receiver Decoder (IRD), or perhaps to degrade an incoming RF/IF signal to a specific C/N ratio in order to establish the noise margin.

To do this, add broadband white Gaussian noise "*in the channel* " to the relatively noise free RF/IF signal. Measure (or compute) the carrier power and then adjust the noise power density to give the required noise power in the selected noise power bandwidth.

Taking the following QAM system parameters as an example:

Symbol rate:	f_S = 6,875 MHz;
Filter roll-off:	α = 0,15;
System noise bandwidth:	BW_{NOISE} = 8 MHz;
Constellation size:	M = 64;
Carrier power (in dB):	C = -25 dBm.

then:

$$C = -25 \text{ dBm}$$

$$E_b = C - 10 \times \log_{10}(\log_2(M) \times f_S) = -101{,}15 \text{ dBm}$$

If a C/N ratio of 23 dB is wanted, then:

$$N = C - \left(\frac{C}{N}\right)_{dB} = -48{,}00 \text{ dBm}$$

$$N_0 = N - 10 \times \log_{10}(BW_{NOISE}) = -117{,}03 \text{ dBm}$$

So the ratio of Carrier-to-Noise applied in an 8 MHz system bandwidth at RF/IF can be described as:

$$\frac{C}{N} = 23{,}00 \text{ dB}$$

$$\frac{E_b}{N_0} = 15{,}88 \text{ dB}$$

This signal is then passed through the receiver root-raised cosine filters. The equivalent noise bandwidth of a bandpass root-raised cosine filter is equal to the symbol rate f_S. The noise power originally defined in an 8 MHz system bandwidth is reduced accordingly:

$$N_{REC} = N + 10 \times \log_{10}\left(\frac{f_S}{BW_{NOISE}}\right) = -48{,}66 \text{ dB} \tag{G.5}$$

The noise power density N_0 is unchanged by the receive filter:

$$N_{0(REC)} = N_0 = -117{,}03 \text{ dBm.}$$

The signal power is already root-raised cosine shaped by the transmitter and so its power is only modified by the factor (1-α/4):

$$C_{REC} = C + 10 \times \log_{10}\left(1 - \frac{\alpha}{4}\right) = -25{,}17 \text{ dB} \tag{G.6}$$

The Energy-per-bit E_b is subject to this same reduction factor: $E_{b(REC)}$ = -101,32 dBm.

So the ratio of Carrier-to-Noise inside the receiver can be described as:

$$\frac{C_{REC}}{N_{REC}} = 23{,}49 \text{ dB}$$

$$\frac{E_{b(REC)}}{N_{0(REC)}} = 15{,}71 \text{ dB}$$

It is this received C/N (or E_b/N_0) ratio that, when demodulated translates directly to a Signal-to-Noise Ratio (SNR) in the I/Q domain. In the idealized case that white Gaussian noise is the only impairment present then this also determines the Modulation Error Ratio (MER).

A general formula for the C/N modification due to the receive filters can be derived:

$$\frac{C_{REC}}{N_{REC}} = \frac{C}{N} + 10 \times \log_{10}\left[\frac{\left(1 - \frac{\alpha}{4}\right)}{\left(\frac{f_S}{BW_{NOISE}}\right)}\right] \text{ dB} \tag{G.7}$$

and another for E_b/N_0:

$$\frac{E_{b(REC)}}{N_{0(REC)}} = \frac{E_b}{N_0} + 10 \times \log_{10}\left[1 - \frac{\alpha}{4}\right] \text{ dB} \tag{G.8}$$

For the C/N case the correction factor is dependent on filter roll-off, symbol rate and the system noise bandwidth used to define the noise power. However, **if the occupied bandwidth is used as the system noise bandwidth,** then equation G.7 simplifies to:

$$\frac{C_{REC}}{N_{REC}} = \frac{C}{N} + 10 \times \log_{10}\left[\frac{\left(1 - \frac{\alpha}{4}\right)}{\left(\frac{1}{1+\alpha}\right)}\right] \text{ dB} \tag{G.9}$$

and the correction factor becomes a constant dependent on the filter α only.

For DVB-C with filter α = 0,15 $\frac{C_{REC}}{N_{REC}} = \frac{C}{N} + 0{,}441$ dB;

For DVB-S with filter α = 0,35 $\frac{C_{REC}}{N_{REC}} = \frac{C}{N} + 0{,}906$ dB.

For comparison, if one were to always use the channel bandwidth (e.g. 8 MHz) as the system noise bandwidth then one should use equation G.7, the correction factor becomes symbol rate dependent, and ranges from +0,441 dB for a theoretical maximum occupancy symbol rate of 6,957 MBaud, through +0,492 dB for the example symbol rate of 6,875 MBaud, to +1,285 dB for a typical lower rate of 5,728 MBaud.

For the E_b/N_0 case the correction for the DVB-C standard filter roll-off of α = 0,15 the correction factor is -0,166 dB, and for the DVB-S standard filter roll-off of α = 0,35 it is -0,398 dB.

It is perhaps worth mentioning that using the C/N correction formula (equation G.7) gives correction factors which suggest that the C/N is actually improved by the receive filter, but this is only because the system noise bandwidth is larger than the receiver noise bandwidth.

The E_b/N_0 formula (equation G.8) more accurately reflects reality, the information-to-noise ratio is actually degraded by a small amount by the receive filter, because for the filter to pass the RF signal spectrum properly at the band edges it should also pass proportionately more noise power than signal power.

G.8 Signal-to-Noise Ratio (SNR) and Modulation Error Ratio (MER)

When a randomly modulated QAM or QPSK carrier and the associated passband noise are demodulated, approximately half the signal power and half the noise power will be delivered into each baseband component channel (I and Q). The demodulation process will have a certain gain, but this gain factor will apply equally to the signal and to the noise so the resulting SNR in each channel will be approximately the same as the C_{REC}/N_{REC} ratio computed above.

The vector sum of the mean I and Q signal powers ratioed to the vector sum of the mean I and Q noise powers will, at least theoretically, be exactly the same as the C_{REC}/N_{REC} ratio computed above.

This ratio of I/Q signal power to I/Q noise power expressed in dB is the definition given in the present document for both SNR and for MER. The difference between these two measurements lines in what perturbations of the received signal are included in the computation.

When the only significant impairment is noise then SNR and MER are equivalent, and are numerically equal to C_{REC}/N_{REC}. The relationship between C_{REC}/N_{REC} and C/N depends on the choice of system noise bandwidth. If the symbol rate is chosen as the system noise bandwidth (as defined in the present document clause 6.7) then the relationship is a fixed offset of a fraction of 1 dB as described above.

This would appear to suggest that C/N measured in the passband can be equated directly to SNR in baseband. Unfortunately other factors should also be considered in a real system. The SNR of the source modulator, the signal amplitude dependence of the noise floor of system components, and the fact that the receiver equalizer will have the effect of translating some linear impairments into noise. The exact interrelation of these parameters is the subject of further study.

G.9 BER vs. C/N

As was stated in the introduction, the Bit Error Rate (BER) as a function of Carrier-to-Noise ratio (C/N) is the most important figure of merit for any digital transmission system.

To evaluate the performance of modulator and demodulator realizations, measured BER values are compared against the theoretical limits of the Bit Error Probability (BEP) P_B. Regarding DVB satellite and cable transmission schemes the BEP is usually determined based on the following assumptions:

- the only noise present is additive white Gaussian noise;
- the channel itself does not introduce any linear or non-linear distortions;
- modulator and demodulator are perfect devices (no timing errors, ideal band-limiting filters).

Based on these assumptions it is possible to calculate fairly accurate upper limits for BEP vs. C/N.

Since C/N depends on noise bandwidth it is common practice to normalize C/N by using E_b/N_0 instead, where E_b is the Energy-per-bit and N_0 is the noise density. The transition from one value to the other is given by:

$$\frac{E_b}{N_0} = \frac{C}{N} \times \frac{BW_{NOISE}}{f_S \times m} \tag{G.10}$$

where BW_{NOISE} is the equivalent noise bandwidth, f_S is the symbol rate, and m is the number of bits-per-symbol, $m = \log_2(M)$, where M is the number of constellation points. When applying this formula it is important to be consistent in using either the "in the channel" C/N or the "in the receiver" C/N values.

If Forward Error Correction (FEC) is employed, the information rate R_I is increased up to the transmission rate R_T by adding the FEC information. The relation:

$$R_C = \frac{R_I}{R_T} \tag{G.11}$$

is called the FEC rate. The transmission rate of an FEC rate 1/2 system for example will be 2 times the information rate. Therefore the "Transmission Rate" E_b/N_0 will be 3 dB less than the "Information Rate" E_b/N_0, provided C/N stays constant. This results from the fact that half of the available signal power is spent on FEC information. To compensate for this effect E_b/N_0 should be increased by 3 dB in case of "Information Rate" BEP. In general, if the BEP should be calculated based on the information rate, E_b/N_0 should be increased by $10 \times \log_{10}(1/R_C)$ dB.

If the performance of different FEC schemes is to be compared for power limited channels like satellite transmission, the information rate should be used because it explicitly takes into account the signal power which is used for redundancy only, and which is therefore lost for the information itself. In case of bandwidth limited channels like cable results based on the transmission rate may be more appropriate.

G.10 Error probability of Quadrature Amplitude Modulation (QAM)

Each state in an M state QAM constellation represents a $\log_2(M) = m$ bit symbol. For example, each state in a 64 QAM constellation represents a 6-bit symbol.

When the received signal is perturbed by Additive White Gaussian Noise (AWGN) there is a probability that any particular symbol will be wrongly decoded into one of the adjacent symbols. The Symbol Error Probability P_S of QAM with M constellation points, arranged in a rectangular set, for m even, is given by (see "Digital Communication" [i.38]):

$$P_S\left(\frac{E_b}{N_0}\right) = 2\times\left(1-\frac{1}{\sqrt{M}}\right)\times \operatorname{erfc}\left[\sqrt{\frac{3\times\log_2(M)}{2\times(M-1)}\times\frac{E_b}{N_0}}\right]\times\left\{1-\frac{1}{2}\times\left(1-\frac{1}{\sqrt{M}}\right)\times\operatorname{erfc}\left[\sqrt{\frac{3\times\log_2(M)}{2\times(M-1)}\times\frac{E_b}{N_0}}\right]\right\} \tag{G.12}$$

where erfc(x) is the complimentary error function given by:

$$\mathrm{erfc}(x) = \frac{2}{\sqrt{\pi}} \int_x^{\infty} e^{-t^2} dt$$

For practical purposes equation G.12 can be simplified by omitting the, generally insignificant, joint probability term to give the approximation;

$$P_S\left(\frac{E_b}{N_0}\right) = 2 \times \left(1 - \frac{1}{\sqrt{M}}\right) \times erfc\left[\sqrt{\frac{3 \times \log_2(M)}{2 \times (M-1)} \times \frac{E_b}{N_0}}\right] \tag{G.13}$$

This approximation introduces an error which increases with degrading E_b/N_0, but is still less than 0,1 dB for 64 QAM at E_b/N_0 = 10 dB.

When M is not an even number (for example M = 5 (32 QAM) or M = 7 (128 QAM), then equation G.14 provides a good approximation to the upper bound on P_S ("Digital Communication" [i.38]):

$$P_S\left(\frac{E_b}{N_0}\right) \leq 1 - \left[1 - \mathrm{erfc}\left(\sqrt{\frac{3 \times \log_2(M)}{2 \times (M-1)} \times \frac{E_b}{N_0}}\right)\right]^2 \tag{G.14}$$

As already stated, the above equations for Symbol Error Probability are based certain simplifying assumptions which can be summarized as "the system is perfect except for the presence of additive white Gaussian noise", but within this rather generous constraint the equations for P_S are exact.

The corresponding Bit Error Probability (BEP) is less easily determined. It is directly related to the Symbol Error Probability (SEP) but the exact relationship depends on how many bit errors are caused by each symbol error, and that in turn depends on the constellation mapping and the use of differential encoding.

Two different approaches can be found in the literature. The first one makes no assumption about the constellation mapping and is based on the probability that any particular bit in a symbol of p bits is in error, given that the symbol itself is in error (see "Digital Communication" [i.38] and see also "Satellite Communications" [i.42]). This approach leads to:

$$P_B = \frac{2^{(p-1)}}{2^p - 1} \times P_S \tag{G.15}$$

The other approach assumes that an erroneous symbol contains just one bit in error. This assumption is valid as long as a Gray coded mapping is used and the BER is not too high. Under these assumptions:

$$P_B = \frac{1}{p} \times P_S \tag{G.16}$$

These approaches give different results for symbols of two or more bits. The second approach is generally adopted because DVB systems employ Gray code mapping. The results tabulated in annex D are based on equations G.12 and G.16.

It should be mentioned that for QAM systems DVB only employs Gray coding within each quadrant, the quadrant boundaries are not Gray coded, and the mapping is partially differentially coded. Further work is required to establish the exact P_B to P_S relationship for this combination of mapping and coding.

G.11 Error probability of QPSK

QPSK can be analysed as 4 QAM. Evaluation of the general QAM equation (G.12) for M = 4 gives:

$$P_S\left(\frac{E_b}{N_0}\right) = \operatorname{erfc}\left(\sqrt{\frac{E_b}{N_0}}\right) \times \left[1 - \frac{1}{4} \times \operatorname{erfc}\left(\sqrt{\frac{E_b}{N_0}}\right)\right] \tag{G.17}$$

Again this can be simplified by dropping the joint probability term to give:

$$P_S\left(\frac{E_b}{N_0}\right) = \operatorname{erfc}\left(\sqrt{\frac{E_b}{N_0}}\right)$$

Using the P_S to P_B relationship defined in equation G.16, the expression for P_B for QPSK modulation becomes:

$$P_B\left(\frac{E_b}{N_0}\right) = \frac{1}{2} \times \operatorname{erfc}\left(\sqrt{\frac{E_b}{N_0}}\right) \tag{G.18}$$

G.12 Error probability after Viterbi decoding

Since it is not possible to derive exact theoretical expressions for the performance of convolutional codes, only upper bounds can be presented in this annex. The upper bound:

$$P_B\left(\frac{E_b}{N_0}\right) \le \frac{1}{k} \times \frac{1}{2} \times \sum_{d=d_f}^{\infty} w(d) \times \operatorname{erfc}\left(\sqrt{R_c \times d \times \frac{E_b}{N_0}}\right) \tag{G.19}$$

provides a good approximation for infinite precision, soft decision Viterbi decoding and infinite path history, as long as E_b/N_0 is not too low (see "High-Rate Punctured Convolutional Codes for Viterbi and Sequential Decoding", IEEE Trans. Commun., vol 37 [i.39] and also see "Further Results on High-Rate Punctured Convolutional Codes for Viterbi and Sequential Decoding", IEEE Trans. Commun., vol 38 [i.40]).

In equation G.19, d_f specifies the free distance of the used code, *w(d)* can be derived from the transfer function of the convolutional code or determined directly by exhaustive search in the trellis diagram of the code, $R_c = k/n$ is the rate of the convolutional code, and E_b/N_0 is given for the transmission rate. Since erfc(x) converges to zero quite quickly for increasing x only very few terms of the sum should be taken into account. Values for d_f and *w(d)* can be found in table G.1 regarding convolutional codes used in DVB satellite transmissions. The performance of convolutional codes for low E_b/N_0 values can only be evaluated by simulations.

Table G.1: Free distance and weights *w(d)* for DVB convolutional codes

Code Rate Rc	1/2	2/3	3/4	5/6	7/8
free distance df	10	6	5	4	3
w(df)	36	3	42	92	9
w(df+1)	0	70	201	528	500
w(df+2)	211	285	1 492	8 694	7 437
w(df+3)	0	1 276	10 469	79 453	105 707
w(df+4)	1 404	6 160	62 935	791 795	1 402 089
w(df+5)	0	27 128	379 546	7 369 828	17 888 043
w(df+6)	11 633	117 019	2 252 394	67 809 347	221 889 258
w(df+7)	0	498 835	13 064 540	609 896 348	2 699 950 506
w(df+8)		2 103 480	75 080 308	5 416 272 113	32 328 278 848
w(df+9)		8 781 268	427 474 864	47 544 404 956	382 413 392 069

G.13 Error probability after RS decoding

A Reed-Solomon (RS) code is specified by the number of transmitted symbols (note) in a block N and the number of information symbols K (see "Error Control Coding Handbook" [i.41]).

Such a code will be able to correct up to t = (N-K)/2 symbol errors. As for DVB transmission N = 204 and K = 188 are used. Therefore up to t = 8 erroneous symbols can be corrected.

NOTE: Whereas the symbols mentioned in context with QAM and QPSK are related to the modulation the symbols mentioned here are just a group of bits.

The probability P_{BLOCK} of an undetected error for a block of N symbols as a function of the error probability of the incoming symbols P_{SIN} is given by:

$$P_{BLOCK} = \sum_{i=t+1}^{N} \binom{N}{i} \times P_{SIN}^{i} \times (1 - P_{SIN})^{N-i} \tag{G.20}$$

From this expression the probability:

$$P_S = \frac{1}{N} \times \sum_{i=t+1}^{N} \beta_i \times \binom{N}{i} \times P_{SIN}^{i} \times (1 - P_{SIN})^{N-i} \tag{G.21}$$

of a symbol error can be derived, where β_i is the average number of symbol errors remaining in the received block given that the channel caused i symbol errors. Of course $\beta_i = 0$ for $i \le t$. When $i > t$, β_i can be bounded by considering that if more than "t" errors occur, a decoder which can correct a maximum of "t" errors will at best correct "t" of the errors and at worst add "t" errors. So:

$$i - t \le \beta_i \le i + t \tag{G.22}$$

is the possible range for β_i. A good approximation is $\beta_i = i$ but also $\beta_i = t + i$ is used, which can be regarded as an upper limit. From G.21 the BEP can be calculated by using G.15 or G.16.

G.14 BEP vs. C/N for DVB cable transmission

For DVB transmission in cable networks, QAM-M systems with M = 16, 32 and 64 are specified. To evaluate the BEP after RS decoding, the following steps should be done:

a) calculate the SEP after QAM demodulation by using (G.12) or (G.14);

b) transform the SEP into a BEP by applying (G.15) or (G.16) to the SEP with p = m;

c) transform the resulting BEP into a SEP with p = 8 by using (G.15) or (G.16);

d) use (G.21) to calculate the SEP PS after RS decoding;

e) apply (G.15) or (G.16) to P_S with p = 8 to determine the final BEP;

f) if the BEP should be based on the information rate, shift the curve by:

- $10 \times \log_{10}(204/188) = 0{,}35$ dB to the right.

If just the BEP before Reed-Solomon is needed, only the first two steps are necessary. In this case there is no difference between information rate and transmission rate. All bits are regarded as information bits.

The limits before and after Reed-Solomon decoding for M = 64, $\beta_i = i$ and E_b, based on the transmission rate, are presented in figure G.1.

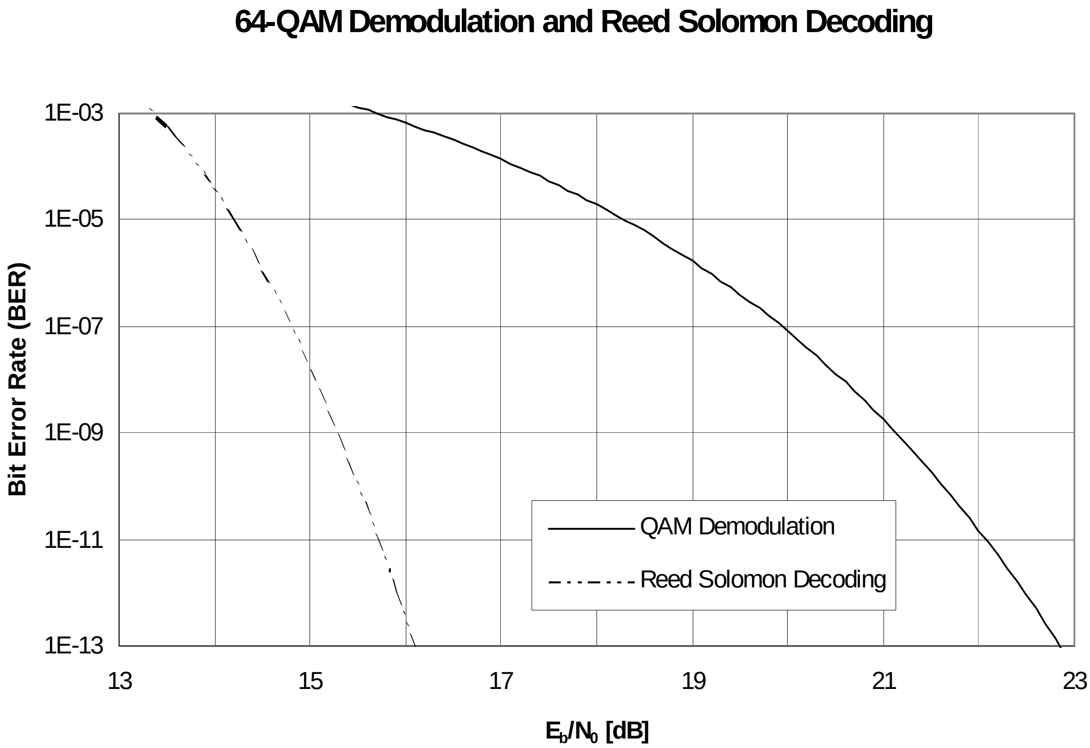


Figure G.1: BER for QAM-64 DVB cable transmission

G.15 BER vs. C/N for DVB satellite transmission

For satellite transmission three different BEPs are possible:

- BEP after QPSK demodulation;
- BEP after Viterbi decoding;

- BEP after Reed-Solomon decoding.

The BEP after QPSK can be derived from equation (G.17). There is no difference to be made between information bit rate and transmission bit rate.

The BEP after Viterbi decoding is expressed by equation (G.18). The result is based on the information rate, because RC is taken explicitly into account in equation (G.18).

BEP after Reed-Solomon decoding can be derived from the above result by applying the following steps to the outcome of equation (G.18):

a) transform the BEP after Viterbi decoding into a SEP by using equation (G.15) or (G.16) with p = 8;

b) use equation (G.17) to determine the SEP after Reed-Solomon decoding;

c) apply equation (G.15) or (G.16) to P_S with p = 8 to determine the final BEP;

d) if the BEP should be based on the information rate, shift the curve by:

$$10 \times \log_{10}(204/188) = 0{,}35 \text{ dB to the right.}$$

The results for the three different BEPs and for all the different code rates R_c are presented in figure G.2.

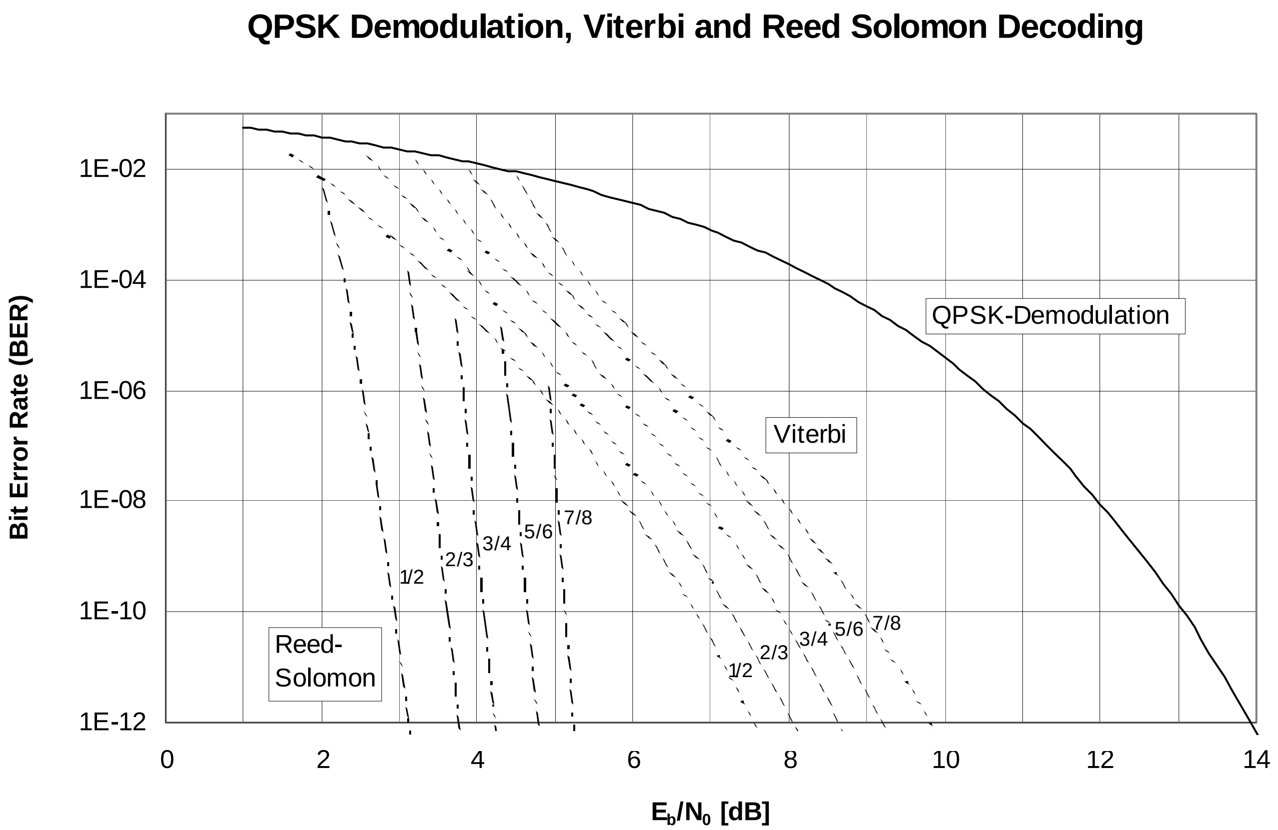


Figure G.2: BER for DVB satellite transmission

Since it is common practice in satellite transmission to refer the results to the information rates the curves for BEP after Reed-Solomon decoding have been shifted accordingly. The equation (G.19) is only valid for low error rates. Despite the fact that for decreasing E_b/N_0 the BER should converge to 1/2 the results according to (G.19) will possess a singularity for $E_b/N_0 = 0$. This behaviour is especially pronounced for $R_c = 7/8$, where the assumption of a low error rate is not fulfilled above a BEP of 10^{-4}.

G.16 Adding noise to a noisy signal

In a practical situation where noise is deliberately added to real signal in order to create a specific C/N ratio for measurement purposes, it is important to realize that there are two fundamental assumptions implicit in this technique.

The first assumption is that the input signal has a high C/N ratio and can, for practical purposes, be regarded as carrier only. The second assumption is that the input signal has a considerably better C/N ratio than the target C/N ratio. In practice noise may be added to an already noisy signal, and in this case there are accuracy issues related to the above assumptions that should be considered.

First consider how noise is typically added to a signal. Figure G.3 gives a simplified block diagram.

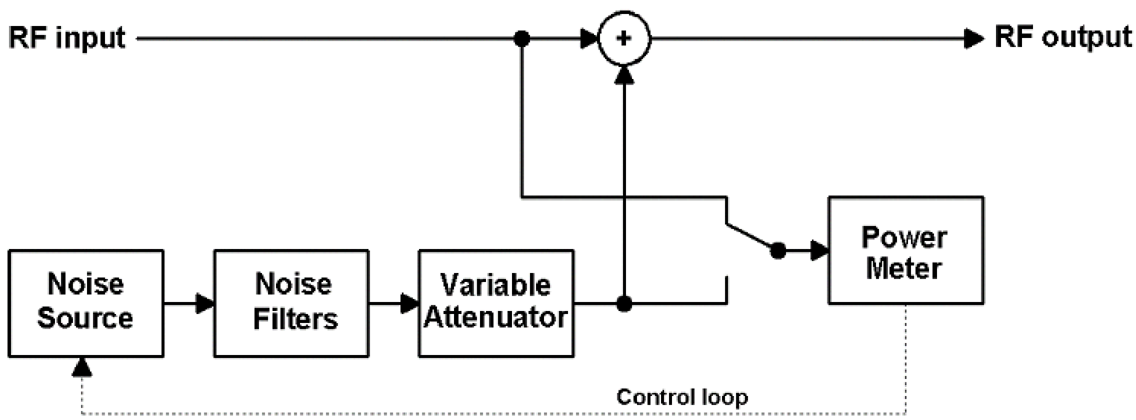


Figure G.3: Simplified block diagram of C/N test set

The input is the carrier signal to be impaired. The carrier power is measured using the power metre. A broadband Gaussian noise source is then filtered and attenuated appropriately to deliver the required noise density (N_0) across the frequency band of interest. The same power metre is used to set the noise power which helps ensure good C/N_0 ratio accuracy, The generated noise is added to the input signal to achieve the required C/N_0 ratio in the output signal. Finally, the carrier power is monitored and the power of the noise source is adjusted accordingly to maintain the required C/N_0.

In automated versions of this process, the user simply selects the desired C/N_0 ratio. This can be entered as C/N_0, but it is more typically entered as C/N which requires that the user also enters the receiver or system noise bandwidth, or it can be input as E_b/N_0 which requires that the user also enters the system bit rate.

From this description it is evident that it is assumed that all the measured input power is carrier and the noise power to achieve the required C/N ratio is computed accordingly. If the input already contains some noise or other carriers then this will:

a) appear at the output in addition to the generated noise;

b) cause the generated noise power to be too large because it is based on the C + N power at the input, not just the C power. This error is exacerbated if the input is not band limited.

A formula can be derived for the actual output C/N ratio as a sum of the theoretical C/N ratio and an error term:

$$CN_{actual} = \underbrace{10\times\log_{10}\left[\frac{C}{N_c}\right]}_{\substack{theoretical \\ C/N ratio}} - \underbrace{10\times\log_{10}\left[\frac{N_c}{N_i+N_c+N_n}\right]}_{error term} \quad \text{dB} \qquad (G.23)$$

Where N_c is the noise power added due to the carrier power, N_i is the noise power already present in the input, N_n is the noise power added due to the input noise. If further manipulation of the error term are performed then an expression can be derived in terms of the fractional input and output C/N ratios.

$$CN_{error} = 10 \times \log_{10} \left[\frac{1}{\frac{1}{CN_{in}} + \frac{CN_{out}}{CN_{in}} + 1} \right] \text{dB} \tag{G.24}$$

The error becomes significant if either the $1/CN_{in}$ or the CN_{out}/CN_{in} term in the denominator moves away from zero which will happen if either the C/N_{in} ratio or the C/N_{out} to C/N_{in} margin is reduced.

The present document gives a minimum value of 15 dB for the C/N_{in} ratio and for the C/N_{out} to C/N_{in} margin as a guideline figure. To meet this condition in satellite systems it is necessary to use a sufficiently large dish to get the required C/N ratio. A received C/N ratio of 20 dB or more is desirable.

Alternatively, it is possible to work with higher noise signals if it is possible to measure the carrier and noise power accurately, for example by measuring carrier plus noise then switching off the carrier and measuring noise only. Equation G.23 can then be used to compensate for the errors due to the input noise.

Annex H: Void

Annex I: PCR related measurements

Void.

Annex J: Bitrate related measurements

J.1 Introduction

J.1.1 Purpose of bitrate measurement

This annex is intended to clarify a bitrate measurement technique which will allow different vendors of equipment to display the same bitrate value on their equipment when they analyse the same transport stream.

The measurement technique in the present document should be applicable to the whole transport stream as well as its individual components. This should allow displays of transport stream information such as the traditional "bouncing bars" statistical multiplex display to be shown consistently on different equipment. This display is intended to dynamically show the different allocation of bitrate between different services. The intention is that the measurement should be stand-alone and non-intrusive.

The measurement technique should also be easy to implement so that cost-effective designs can be introduced to large MPTS systems. It should also be scalable so that as extra precision is required, a more expensive device can be built using the same principles.

The technique is also appropriate for non Transport Stream system, but the use in such systems is outside the scope of the present document.

J.1.2 User Rate versus Multiplex Rate

MPEG-2 transport streams are comprised of many different elements including but not limited to multiple compressed video and audio streams, teletext, table data, conditional access streams, IP data, and other private data. Each of these individual elements and the overall transport stream have data rates associated with them. The data rates can be time varying for the individual elements and the overall stream.

It is of importance to define the measurement of these rates and have a common definition for these measurements. Before the measurements can be defined, the multiplexing of all the elements into a transport stream needs to be understood with regards to rate calculations.

Figure J.1 depicts a general representation of the multiplexing process.

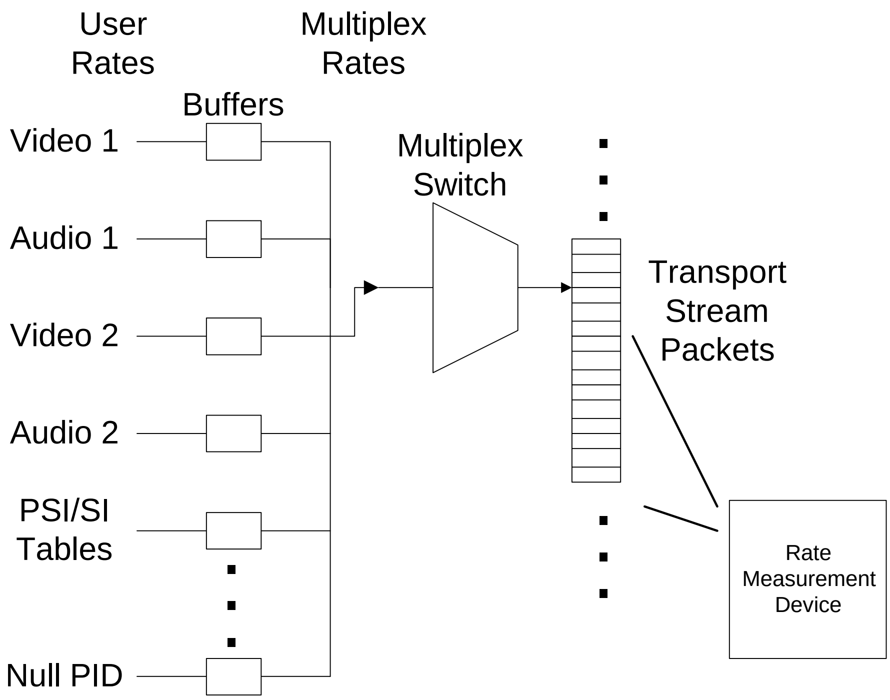


Figure J.1: General representation of the multiplexing process

This diagram represents a number of different elements being multiplexed into a single transport stream. Before all the streams are multiplexed together they can be considered to have User rates which are established by the user (e.g. 4 Mbits/s for Video 1). It can be modelled that each element has a User data rate entering the buffer and a Multiplex rate leaving the buffer since the data is extracted directly from the buffer and placed as a complete packet in the transport stream. Over the long term average, the User and Multiplex rates should be the same, but the creation of the transport stream through the multiplex process can either increase or decrease the User rate in the actual transport stream over a specific Time Gate. For example, the video might have a 4,1 Mbits of data over a one-second Time Gate in the transport stream, but in the next one second interval it could have 3,9 Mbits. But with respect to the PTS/DTS values in the stream, the video rate as set by the user could still be 4,0 Mbits/s.

The Multiplex rates will also depend upon what is actually being multiplexed together, and the measurement of the multiplex rate in the output stream will vary if different elements are combined. If only one video is being transmitted at one time and another video is being transmitted at another time, the output Multiplex rate will be different at those two times even if the User rate has not changed.

The User rate for video also needs to be better understood since a single number is often given for this rate (e.g. 4 Mbits/s). This number typically means the total number of bits in a GOP multiplied by the number of GOPs per second. The actual rate of video varies with each frame. An I frame typically receives a much higher percentage of the bits compared to the B and P frames. What generally happens is that even though the I frame has significantly more data than a B frame, it will take longer to transmit this frame and the Multiplex rate can approach the User rate. This definition of User rate for video applies to both the CBR and VBR approaches. In the CBR case, the user provides one value for the rate, while in the VBR case the user provides a minimum and maximum and typically lets compression equipment vary the rate between these parameters in order to maximize video quality based on some constraints. The rate as calculated by the compression equipment is still considered a User rate since it is before the video data is multiplexed into the transport stream.

Since the rates of the elements are less than or equal to the rate of the output transport stream, the positioning of these elements in the output stream is important to consider in calculating the User rate. For example, an element that generates 10 packets per second may have these packets placed at the beginning of the second, in the middle, dispersed throughout, etc. Buffer models in general restrict the packet placement but as an extreme example, it could be assumed that the packets are placed at the beginning of a second and the transport rate is 1,5040 Mbits/s. If the Time Gate of a rate measurement of this element is 0,1 s and this Time Gate started with the transmission of these packets, the first rate measurement would be 0,1504 Mbits/s. If the next measurement also uses 0,1 s of duration and starts just after the packet is transmitted, the rate would be 0,0 Mbits/s. Neither of these numbers matches the expected User rate of 0,01504 Mbits/s.

A real world example for a 256 kbit/s audio stream can easily indicate differences of 2 % in the User rate versus the Multiplex rate. This audio stream has approximately 200 packets per second with each audio frame containing about 5 packets. In a measurement interval of one second that begins in the second half of an audio frame, all 5 of the first packets can be transmitted in the second half of an audio frame, and all 5 of the last five packets can be transmitted in the first half of the last audio frame. These results in a Multiplex rate of 205 packets per second that is 2,5 % higher than the User rate of 200 packets per second. This error difference can increase with smaller measurement intervals since for a 100 ms interval the number of packets for the User rate would be 20 while the Multiplex rate could be 25 resulting in a 25 % difference.

J.1.3 User rate applications

The rate measurements for transport streams are computed for a variety of purposes. These include but are not limited to:

- Verification/conformance/troubleshooting - the overall transport stream rate or rates of individual elements are expected to be certain values as set by a user or compression/multiplex system. The user needs to validate that the rates in the stream meet the "expected" rates. This validation can be done over time or just once and can include statistics (e.g. minimum and maximum) as well as history of any rate calculation values. The validation would include all elements including video, audio, conditional access data, PSI/SI tables, etc.

- Video and audio quality - there is a strong correlation between video and audio quality and the rate at which these items are transmitted in the transport stream. There is especially a need to monitor the rate of the video since this rate often varies over time and if an video quality issue is determined by visual inspection, there would be a need to determine the rate of the video at that time. A service provider may also guarantee a minimum bit rate for video and audio for a particular program and with a contract, this provider will need to prove that those rates have been met.

- Sale of bandwidth - there is a need to monitor the rate of individual elements in a stream over a longer period so that a service provider can charge a user for the bandwidth that has been used in one hour or one day or one week, etc.

- Monitoring - there is a need to generate an alarm if the rate of a particular element or the whole stream goes outside some user-specified minimum and maximum range. This error could mean that an element is no longer being included in the transport stream due to a multiplexer malfunction. The accuracy of these rate measurements is not critical to the overall application.

J.2 Principles of Bit rate measurement

J.2.0 General

This is a difficult subject as a measured bitrate depends on the time over which the bitrate is averaged. Bit rate is usually expressed in terms of bits per second, but the actual value that is measured will depend on the way the bits are counted.

A bitrate measurement will depend on where in the system the bitrate is measured. For example, in a system, slightly different bitrates may be seen depending on whether the bitrate is measured before or after a large buffer.

J.2.1 Gate or Window function

On the assumption that Transport Stream packet based systems are being dealt with in the DVB world, there are 3 main choices when counting bytes:

- packet based - count only the synchronization bytes;

- byte based - count every byte when it arrives;

- bit based - count every bit as it arrives.

There are also 2 options for applying the window function:

- "continuously" rolling window;
- a jumping window (the end of each window is the start of the next window).

A jumping window is very undesirable as the bitrate measured will vary depending on when the window is first applied. This rules it out very early. A rolling window is therefore more desirable, but some caution is needed in the use of the term "continuous".

The most precise bitrates would be given with a bit based counting scheme. Here, each time a new bit is received, or sent, the total number of bits in the last time window (e.g. 1 s) could be counted and a value displayed. This would always give the most accurate value, but there are a number of serious technical difficulties in implementing this, particularly in offline and semi-offline systems. These difficulties include processing bandwidth and timing accuracy. A byte based system also requires large bandwidth, but both bit and byte based may be required in some special circumstances. Although the present document does not to define byte or bit based profiles, they can easily be added by counting the bytes or bits and adjusting the nomenclature appropriately.

A packet based approach may be favourable in situations where cheap implementations with reasonable accuracy are required. It is likely that most DVB Tx and Rx systems would have the capability of deriving some timing information on a packet basis.

J.2.2 "Continuous window"

If all transport streams were of a constant bitrate, not bursty, continuously clocked and could be easily analysed as a signal with fixed and uniform temporal sampling, then bitrate measurement would be easy.

In real systems (bursty ASI, Transport streams over IP, 1394b hubs, cascaded networks, etc.) the bytes and packets do not necessarily arrive on a uniform sampling grid and pragmatic measures need to be taken in defining the window function. To simplify implementation, systems have been considered where the window function is moved across the data in different ways: by byte, by packet, by fixed time interval.

There are several points to note about the algorithm in the present document:

1) Strictly speaking, this measure is not continuous.
2) It is a discrete measure whose bitrate values are only valid on time slice boundaries.
3) It is easy to implement and gives a new TS bitrate value every τ (11,1 μs to 1 s).
4) It is applicable to partial transport streams where only a subset of PIDs are being inspected.
5) It can be extended to measure the bitrate of the payload of TS packets.
6) It is repeatable between equipment vendors because the time slice can be made sufficiently small to ensure aliasing is not a problem e.g. when $\tau = 1 / 90$ kHz

J.2.3 Time Gate values:

20 ms: gives the peak bitrate of a stream based on variable bitrate elements within it.

1 s: gives a longer term "smooth" average.

user: could be used for elements such as subtitles which may only be present from time to time and may require windows of 1 minute or more.

J.2.4 Rate measurements in a transport stream

Only the Multiplex rates are available to be measured in the transport stream and not the original User rates. In general, it is the User rates that are of interest as outputs of a measurement device with some exception regarding issues of burstiness and buffer models.

Depending on the customer application, the parameters that should be used in the MG bitrate equation in clause 5.3.3. will be different if the user wants to measure User rates or Multiplex rates as finding the best accuracy for the User rates is different than finding the best accuracy for the Multiplex rates. The parameters also need to take into account tracking the changes in the rate versus time. The parameters should in general be different for elements that differ either in type or in rate in order to maintain accuracy.

Here are some general considerations for the parameters:

- For elements that have CBR, increasing T will push the measured Multiplex rate towards the User rate.

- For reasonable accuracy of the User rate, T should be large enough to include multiple elements of what is being measured. For example, if the rate of a SDT is being measured, it should include at least 10 different arrivals of the SDT.

Decreasing τ will cause the Multiplex rate to be more accurately tracked but will not increase the accuracy of calculating User rates for CBR streams. For VBR streams, a smaller τ to within some limits will allow the changes to be better averaged over time.

J.3 Use of the MG profiles

J.3.0 General

The profiles in clause 5.3.3.2 have been designed to have the properties described below.

J.3.1 MGB1 Profile - the backwards compatible profile

This is a backwards compatible profile where a 1 second jumping window is used to measure bitrate. In a rigidly CBR system, this will give a good indication of the bitrate, but will give aliasing and inaccuracy if the bitrate being measured is changing faster than every 1s. This makes it impractical for looking at VBR systems, or for looking at the bitrates of VBR components (e.g. stat-mux video) in a CBR transport stream.

This profile is included for backwards compatibility with existing equipment.

J.3.2 MGB2 Profile - the Basic bitrate profile

This profile is recommended for new designs. It is intended to give a good idea of the average bitrate of a system, yet have enough resolution (due to a small τ value) to show whether the bitrate is truly static or is varying with time. The values have been chosen to allow simple implementation.

J.3.3 MGB3 Profile - the precise Peak bitrate profile

This profile has a time gate which is small enough to show the variable bitrate characteristics of a statistical multiplex environment. The timeSlice is small enough to ensure that only a single packet header will occur in each timeSlice for most distribution systems. The time gate is short enough so that frame by frame averaging does not take place. The timebase chosen can be locked to, or derived from the PCR in a decoder or encoder environment for ease of implementation.

J.3.4 MGB4 Profile - the precise profile

This profile is intended to give a "true" smoothed bitrate. The timeSlice is small enough to ensure that only a single packet header will occur in each timeSlice for most distribution systems. The time gate is a little over 1 second to give a long time constant averaging to the data. The timebase chosen can be locked to, or derived from the PCR in a decoder or encoder environment for ease of implementation.

J.3.5 MGB5 Profile - the user profile

This profile is intended to give extensibility to the bitrate measurement algorithm. It allows different time gates and timeSlice values to be defined. These can be applied to the whole transport stream, or to individual components of the stream. It is important when using this profile that the results are carefully documented using the nomenclature in these guidelines. This will ensure that results can be repeated at a later date.

J.4 Error values in the measurements

J.4.0 General

It is worth noting the areas where errors can be introduced into the measurement:

- clock instability in the time gate and time slice functions;
- quantization due to counting elements which are too big e.g. too many or too few packet headers may fall within the time gate;
- aliasing due to having a timeSlice or a time Gate which is too large for the parameter being measured.

In real systems, the errors due to clock instability and quantization tend to be rather small. The biggest problem is inappropriate use of timeSlice and time gate values. This can be best demonstrated by an example.

Imagine a DVB-S statistical multiplex system (38.1 Mbit/s) where a particular video PID has a bitrate limit of 3 -5 Mbit/s and the hypothetical video encoder is able to change its bitrate every 80ms. Bit rate is measured by counting packet headers of a certain PID. The average video rate is 4 Mbit/s.

If the MGB4 profile is used,

DVB-S ≈ 38,1 Mbit/s   packet duration ≈ 40 µs   packets per τ ≈ 0,25

The clock frequency error uncertainty may be as high as 500 ppm. This would lead to an error in the duration of the time gate of 500 ppm (0,05 %). This could increase the 1 second window by 500 µs which at 5 Mbit/s could allow an extra 2 packets into the gate. This would give an error of:

$$= 2 \times 188 \times 8 \text{ bits/s}$$

$$= 0{,}06\ \%\text{ of }5\text{ Mbit/s}$$

The uncertainty due to quantization is equal to the element size which is counted which is 1 packet per time gate in this case:

$$= 188 \times 8 \text{ bit/s} = 1\,504 \text{ bit/s}$$

$$= 0{,}03\ \%\text{ of }5\text{ Mbit/s}$$

It can be seen that these values are all quite small. In case of the slightly contrived example of a sequence which requires the bitrate shown below:

Difficult 5 Mbit/s ←1 s→	Easy 3 Mbit/s ←1 s →	Difficult 5 Mbit/s ←1 s →	Easy 3 Mbit/s ←1 s →	Difficult 5 Mbit/s ←1 s →	Easy 3 Mbit/s ←1 s →

- The MGB4 profile will show a smoothed version of the above bitrate with peak values of 5 Mbit/s and 3 Mbit/s.
- The MGB3 profile will show much sharper edges to the bitrate changes and will report the peak values of 5 Mbit/s and 3 Mbit/s.

- The MGB1 profile, however will show different values depending on the moment when the 1 second window jumps to its next starting point. If it is synchronized with the start of the 1 second sequences, then it will report the correct values of 5 Mbit/s and 3 Mbit/s. If, however it starts its measurements 50 % of the way through a 1 second sequence, it will report that the bitrate is constant at 4 Mbit/s. **This is an error of 33 % at 3 Mbit/s or 20 % at 5 Mbit/s.**

Real errors are less than in this contrived example, but this source of error is the most significant in real systems. Note that in some monitoring applications errors of a few percent may be tolerable, whereas in other applications a precision of 1ppm or better may be required.

J.4.1 Very Precise measurements

In very accurate measurements, it may be necessary to count individual bytes, or individual bits to obtain the required precision. The same algorithm, nomenclature and synchronization as described in clause 5.3.3 may still be used and the results will be repeatable.

Annex K: DVB-T channel characteristics

K.0 Introduction

This annex provides some information on terrestrial channel profiles which can be used for off-line computer simulations and realtime simulations based on dedicated equipment. The properties of these profiles reflect realistic reception conditions and/or worst-case scenarios and were used to verify specific features of the DVB-T standard.

K.1 Theoretical channel profiles for simulations without Doppler shift

(quoted from ETSI EN 300 744 [i.9])

The performance of the DVB-T system has been simulated during the development of the standard ETSI EN 300 744 [i.9] with two channel models for fixed reception - F_1 and portable reception - P_1, respectively.

The channel models have been generated from the following equations where x(t) and y(t) are input and output signals respectively:

a) Fixed reception F_1:

$$y(t) = \frac{\rho_0 \cdot x(t) + \sum_{i=1}^{N} \rho_i \cdot e^{-j \cdot 2\pi \cdot \theta_i} \cdot x(t - \tau_i)}{\sqrt{\sum_{i=0}^{N} {\rho_i}^2}}$$

where:

- the first term before the sum represents the line of sight ray;
- N is the number of echoes equals to 20;
- θ_i is the phase shift from scattering of the i'th path - listed in table K.1;
- ρ_i is the attenuation of the i'th path - listed in table K.1;
- τ_i is the relative delay of the i'th path - listed in table K.1.

The Ricean factor K (the ratio of the power of the direct path (the line of sight ray) to the reflected paths) is given as:

$$\mathrm{K} = \frac{{\mathrm{r}_0}^2}{\sum_{\mathrm{i}=1}^{\mathrm{N}} {\mathrm{r}_\mathrm{i}}^2}$$

In the simulations a Ricean factor K = 10 dB has been used. In this case:

$$\rho_o = \sqrt{10 \cdot \sum_{i=1}^{N} {\rho_i}^2}$$

b) Portable reception, Rayleigh fading (P_1):

$$y(t) = k \cdot \sum_{i=1}^{N} \rho_i \cdot e^{-j \cdot 2\pi \cdot \theta_i} \cdot x(t - \tau_i) \text{ where } k = \frac{1}{\sqrt{\sum_{i=1}^{N} \rho_i^2}}$$

θ_i, ρ_i and τ_i are given in table K.1.

Table K.1: Attenuation, phase and delay values for F_1 and P_1

i	ρ_i	τ_i [µs]	θ_i [rad]
1	*0,057 662*	*1,003 019*	*4,855 121*
2	*0,176 809*	*5,422 091*	*3,419 109*
3	*0,407 163*	*0,518 650*	*5,864 470*
4	*0,303 585*	*2,751 772*	*2,215 894*
5	*0,258 782*	*0,602 895*	*3,758 058*
6	*0,061 831*	*1,016 585*	*5,430 202*
7	*0,150 340*	*0,143 556*	*3,952 093*
8	*0,051 534*	*0,153 832*	*1,093 586*
9	*0,185 074*	*3,324 866*	*5,775 198*
10	*0,400 967*	*1,935 570*	*0,154 459*
11	*0,295 723*	*0,429 948*	*5,928 383*
12	*0,350 825*	*3,228 872*	*3,053 023*
13	*0,262 909*	*0,848 831*	*0,628 578*
14	*0,225 894*	*0,073 883*	*2,128 544*
15	*0,170 996*	*0,203 952*	*1,099 463*
16	*0,149 723*	*0,194 207*	*3,462 951*
17	*0,240 140*	*0,924 450*	*3,664 773*
18	*0,116 587*	*1,381 320*	*2,833 799*
19	*0,221 155*	*0,640 512*	*3,334 290*
20	*0,259 730*	*1,368 671*	*0,393 889*
NOTE: Figures in italics are approximate values.			

NOTE: For practical implementations profiles with reduced complexity have been used successfully. In many cases it seems sufficient to use e. g. only the six paths with the highest amplitude.

K.2 Profiles for realtime simulations without Doppler shift

The following profiles were used in laboratory tests in a research project with satisfactory results.

NOTE: AC106 Validate (1995-1998).

Table K.2: Echo Profiles

Path	fixed delay [µs] C/I [dB]	Portable delay [µs] C/I [dB]	dense SFN delay [µs] C/I [dB]
#1 (main)	0 0	- -	0 0
#2	0,5 17,8	0,5 7,8	7,8 9,3
#3	1,95 17,9	1,95 7,9	11,6 5,5
#4	3,25 19,1	3,25 9,1	17,5 16,1
#5	2,75 20,4	2,75 10,4	20,0 14,5
#6	0,45 20,6	0,45 10,6	23,4 23,4
#7	- -	0,85 11,6	- -

K.3 Profiles for realtime simulation with Doppler shift (mobile channel simulation)

In the course of a research project (see note), three channel profiles were selected to reproduce the DVB-T service delivery situation in a mobile environment. Two of them reproduce the characteristics of the terrestrial channel propagation with a single transmitter, the third one reproduces the situation coming from an SFN operation of the DVB-T network.

NOTE: AC318 Motivate (1998-2000).

The following tables describe the composition of the chosen profiles.

- Typical Urban reception (TU6)
  This profile reproduces the terrestrial propagation in an urban area. It was originally defined by COST 207 [i.44] as a Typical Urban (TU6) profile and is made of 6 paths having wide dispersion in delay and relatively strong power. This channel profile has also been used for GSM and DAB tests.

Table K.3: Echo profile for Typical Urban reception (TU6)

Tap number	Delay (us)	Power (dB)	Doppler spectrum
1	0,0	-3	Rayleigh
2	0,2	0	Rayleigh
3	0,5	-2	Rayleigh
4	1,6	-6	Rayleigh
5	2,3	-8	Rayleigh
6	5,0	-10	Rayleigh

- Typical Rural Area reception (RA6)
  This profile reproduces the terrestrial propagation in an rural area. It has been defined by COST 207 [i.44] as a Typical Rural Area (RA6) profile and is made of 6 paths having relatively short delay and small power. This channel profile has been used for GSM and DAB tests.

Table K.4: Echo profile for Typical Rural Area reception (RA6)

Tap number	Delay (us)	Power (dB)	Doppler spectrum
1	0,0	0	Rice
2	0,1	-4	Rayleigh
3	0,2	-8	Rayleigh
4	0,3	-12	Rayleigh
5	0,4	-16	Rayleigh
6	0,5	-20	Rayleigh

- 0 dB Echo profile
  This profile has been defined by Motivate partners. Its composition has been largely influenced by the specific nature of the DVB-T signal, especially its spread spectrum technique (introducing an Inter Carrier Interference sensitivity to Doppler spread) and its use of a Guard Interval (introducing an Inter Symbol sensitivity to the echoes delays). Moreover, its definition has been driven by the analysis of the profiles encountered during the various field trials performed during the Motivate project.
  This profile is made of two rays having the same power, delayed by half the Guard Interval value and presenting a pure Doppler characteristic.

Table K.5: Profile for 0 dB echo reception

Tap number	Delay (us)	Power (dB)	Doppler spectrum	Frequency ratio
1	0	0	Pure Doppler	-1
2	1/2 Tg	0	Pure Doppler	+1

Annex L: The measurement of MER under ACE

When ACE method means Active Constellation Extension, it uses outer point constellation points of the useful carriers to reduce the Peak to Average value of the DVB-T2 signals. It is an option of the DVB-T2 standard and it is used for example in the L1 signalling parameters since ETSI EN 302 755 (V1.3.1) [i.27].

When ACE is applied on a COFDM signal, the professional receiver is not capable of referring to the theoretical reference point in order to measure the modulation error distance as in the formula.

The problem is to find a solution that allows MER measurement on carriers when Active Constellation Extension is used on a DVB-T2 signal:

1) Enable Active constellation point.

2) Disable ACE on some carriers based on a circular mask pattern. The number of useful carriers is less than 1% in order not to impact the Peak to Average Ratio of the final signal.

3) This method is a test mode located in the modulation part implementing ACE algorithm. The receiver can recover the MER based on the mask pattern definition.

The ACE method is defined in clause 9.6.1 of ETSI EN 302 755 [i.27].

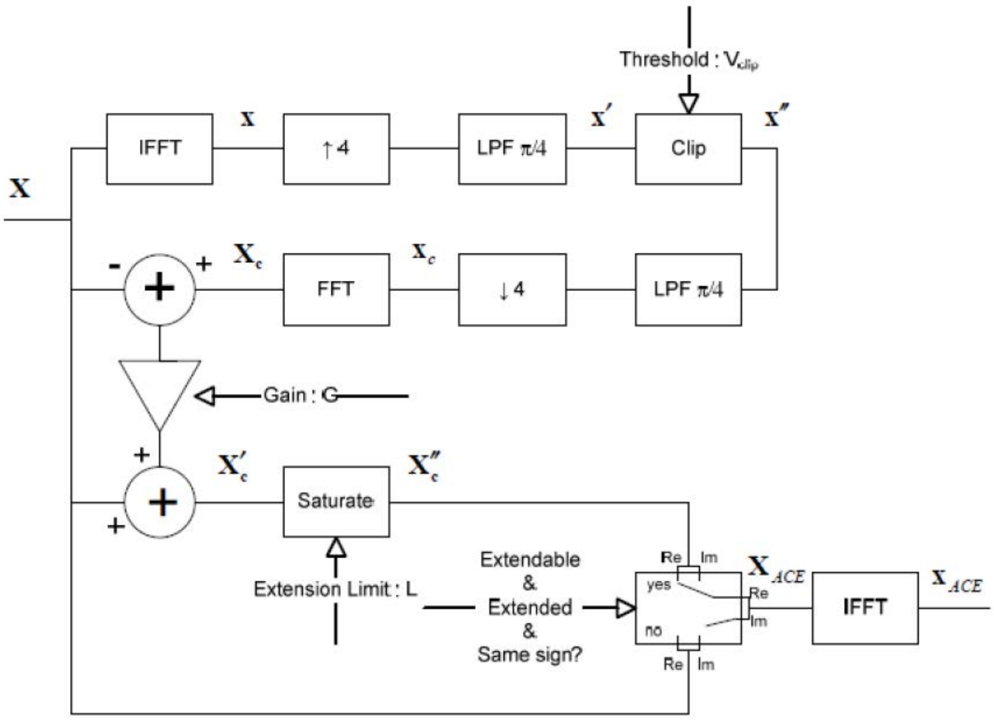


Figure L.1: Implementation of the ACE method (extracted from ETSI EN 302 755 [i.27])

In the figure L.1, "extendable" signification is modified from the DVB-T2 specification ETSI EN 302 755 [i.27] clause 9.6.1, in order to add a circular mask pattern exclusion.

Standard definition of "extendable":

> A component is defined as extendable if it is an active cell (i.e. an OFDM cell carrying a constellation point for L1 signalling or a PLP), and if its absolute amplitude is greater than or equal to the maximal component value associated to the modulation constellation used for that cell; a component is also defined as extendable if it is a dummy cell, a bias balancing cell or an unmodulated cell in the Frame Closing Symbol.

For the test mode, the text is adjusted:

> A component is defined as extendable if it is an active cell (i.e. an OFDM cell carrying a constellation point for L1 signalling or a PLP), and if its absolute amplitude is greater than or equal to the maximal component value associated to the modulation constellation used for that cell; a component is also defined as extendable if it is a dummy cell, a bias balancing cell or an unmodulated cell in the Frame Closing Symbol. **In test mode, active cells corresponding to the mask pattern index position list are not extended.**

The circular mask pattern is changed for each new T2 frame symbol in order to measure MER for all the carriers.

The mask pattern M[] for ACE is defined as follows:

```
Idx=Frame Idx mod 100;
i=0;
While (idx<Cuseful)
{
Idx=mod(Idx+div(Cuseful /100), Cuseful)
M[i]=idx;
i++;
}
```

For a given frame index, where Idx provides the carrier position and M[] is the mask pattern containing index position is ascendant form defined for a Frame number FrameIdx.

Where Cuseful

Cuseful = Cdata for data symbols

Cuseful= CP2 for P2 symbols

Cuseful = CFC for Frame closing symbols

Annex M: Bibliography

- ETSI TS 102 154: "Digital Video Broadcasting (DVB); Implementation guidelines for the use of Video and Audio Coding in Contribution and Primary Distribution Applications based on the MPEG-2 Transport Stream".
- Recommendation ITU-T G.826: "Error performance parameters and objectives for international, constant bit rate digital paths at or above the primary rate".
- ETSI TR 101 290 (V1.2.1) 2001-05: "Digital Video Broadcasting (DVB); Measurement guidelines for DVB systems".

History

Document history		
Edition 1	May 1997	Publication as ETSI ETR 290
V1.2.1	May 2001	Publication
V1.3.1	July 2014	Publication
V1.4.1	June 2020	Publication