TR 101 830-2 (draft) V0.0.0 (2004-xx)

Technical Report

Transmission and Multiplexing (TM); Spectral management on metallic access networks; Part 2: Technical methods for performance evaluations

Work Item Reference	DTS/TM-06030
Permanent Document	**TM6(01)20**
Filename	m01p20a8.pdf (rev 8)
Date	july 6, 2004

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Mark the new changes, valid since nov 24, 2003

Reference

DTS/TM-06030

Keywords

spectral management, unbundling, access network, local loop, transmission, modem, POTS, IDSN, ADSL, HDSL, SDSL, VDSL, xDSL

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Association à but non lucratif enregistrée à la
Sous-Préfecture de Grasse (06) N° 7803/88

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Intellectual Property Rights

Foreword

This Technical Report (TR) has been produced by ETSI Technical Committee Transmission and Multiplexing (TM).

The present document is part 2 of a multi-part deliverable covering Transmission and Multiplexing (TM); Access networks; Spectral management on metallic access networks, as identified below:

Part 1: "Definitions and signal library".

Part 2: "Technical methods for performance evaluations".

Part 3: "Construction methods for spectral management rules".

NOTE: Part 3 is under preparation.

1 Scope

The present document gives guidance on a common methodology for studying the impact on xDSL performance (maximum reach, noise margin, maximum bit rate) in noisy cables when changing parameters within various Spectral Management scenarios. These methods enable reproducible results and a consistent presentation of the assumed conditions (characteristics of cables and xDSL equipment) and configuration (chosen technology mixture and cable fill) of each scenario.

The technical methods include computer models for calculating:

- xDSL receiver capability of detecting signals under noisy conditions;
- xDSL transmitter characteristics;
- cable characteristics
- cross talk cumulation in cables, originating from a mix of xDSL disturbers;

The *objective* is to provide the technical means for evaluating the performance of xDSL equipment within a chosen scenario, such as calculations and measurements. This includes the description of *performance properties* of equipment. Another objective is to assist the reader with applying this methodology by providing examples on how to specify the *configuration* and the *conditions* of a scenario in an unambiguous way. The distinction is that a configuration of a scenario can be controlled by access rules while the conditions of a scenario cannot.

Possible applications of this document include:

- Studying access rules, for the purpose of bounding the cross talk in unbundled networks.
- Studying deployment rules, for the various systems present in the access network.
- Studying the impact of cross talk on various technologies within different scenarios.

The scope of this Spectral Management document is explicitly restricted to the methodology for defining scenarios and quantifying the performance of equipment within such a scenario. All judgement on what access rules are required, what performance is acceptable, or what combinations are spectral compatible, is explicitly beyond the scope of this document. The same applies for how realistic the example scenarios are.

The models in this document are not intended to set requirements for DSL equipment. These requirements are contained in the relevant transceiver specifications. The models in this document are intended to provide a reasonable estimate of real-world performance but may not include every aspect of modem behaviour in real networks. Therefore real-world performance may not accurately match performance numbers calculated with these models.

2 References

For the purposes of this Technical Report (TR) the following references apply:

SpM

[1] ETSI TR 101 830-1 (v1.3.1): " Transmission and Multiplexing (TM); Spectral Management on metallic access networks; Part 1: Definitions and signal library".

[2] ANSI T1E1.4, T1.417-2003: "Spectrum Management for loop transmission systems".

ISDN

[3] ETSI TS 102 080 (v1.4.1): "Transmission and Multiplexing (TM); Integrated Services Digital Network (ISDN) basic rate access; Digital transmission system on metallic local lines".

HDSL

[4] ETSI TS 101 135 (v1.5.3): "Transmission and Multiplexing (TM); High bit-rate Digital Subscriber Line (HDSL) transmission systems on metallic local lines; HDSL core specification and applications for combined ISDN-BA and 2 048 kbit/s transmission".

SDSL

[5] ETSI TS 101 524 (v1.2.1): "Transmission and Multiplexing (TM); Access transmission system on metallic access cables; Symmetrical single pair high bit rate Digital Subscriber Line (SDSL)".

[6] ITU-T Recommendation G.991.2 (12/03): "Single-Pair High-Speed Digital Subscriber Line (SHDSL) transceivers".

ADSL

[7] ETSI TS 101 388 (v1.3.1): "Transmission and Multiplexing (TM); Access transmission systems on metallic access cables; Asymmetric Digital Subscriber Line (ADSL) - European specific requirements".

[8] ITU-T Recommendation G.992.1: "Asymmetric digital subscriber line (ADSL) transceivers".

3 Definitions and abbreviations

3.1 Definitions

For the purposes of the present documents on spectral management, the following terms and definitions apply:

Local Loop Wiring: Part of a metallic access network, terminated by well-defined ports, for transporting signals over a distance of interest. This part includes mainly cables, but may also include a main distribution frame (MDF), street cabinets, and other distribution elements. The local loop wiring is usually passive only, but may include active splitter-filters as well.

Loop provider: Organization facilitating access to the local loop wiring. (NOTE: In several cases the loop provider is historically connected to the incumbent network operator, but other companies may serve as loop provider as well.)

Network operator: Organization that makes use of a local loop wiring for transporting telecommunication services. (NOTE: This definition covers incumbent as well as competitive network operators.)

Access Port: An Access Port is the physical location, appointed by the loop provider, where to inject signals (for transmission purposes) into the local loop wiring.

NT-access port (or NT-port for short): is an access port for injecting signals, labelled by the loop provider as "NT-port". (NOTE: Such a port is commonly located at the customer premises, and intended for injecting "upstream" signals.)

LT-access port (or LT-port for short): is an access port for injecting signals, labelled as labelled by the loop provider as "LT-port". (NOTE: Such a port is commonly located near the telecommunication exchange, and intended for injecting "downstream" signals.)

Transmission technique: electrical technique used for the transportation of information over electrical wiring.

Transmission equipment: equipment connected to the local loop wiring that uses a transmission technique to transport information.

Transmission system: A set of transmission equipment that enables information to be transmitted over some distance between two or more points.

Upstream transmission: transmission direction from a port, labelled as NT-port, to a port, labelled as LT-port. This direction is usually from the customer premises, via the local loop wiring, to the telecommunication exchange.

Downstream transmission: transmission direction from port, labelled as LT-port, to a port, labelled as NT-port. This direction is usually from the telecommunication exchange via the local loop wiring, to the customer premises.

Noise margin: the ratio (P_{n2}/P_{n1}) by which the received noise power P_{n1} may increase to power P_{n2} until the recovered signal does no longer meet the predefined quality criteria. This ratio is commonly expressed in dB.

Signal margin: the ratio (P_{s1}/P_{s2}) by which the received signal power P_{s1} may decrease to power P_{s2} until the recovered signal does no longer meet the predefined quality criteria. This ratio is commonly expressed in dB.

Max data rate: the maximum data rate that can be recovered according to predefined quality criteria, when the received noise is increased with a chosen noise margin (or the received signal is decreased with a chosen signal margin).

Performance: is a measure of how well a transmission system fulfils defined criteria under specified conditions. Such criteria include reach, bit rate and noise margin.

Access Rule: Mandatory rule for achieving access to the local loop wiring, equal for all network operators who are making use of the same network cable that bounds the cross talk in that network cable.

Deployment Rule: Voluntary rule, irrelevant for achieving access to the local loop wiring and proprietary for each individual network operator. (NOTE: A deployment rule reflects a network operator's own view about what the maximum length or maximum bit rate may be for offering a specific transmission service to ensure a chosen minimum quality of service.)

Spectral management rule: A generic term, incorporating (voluntary) deployment rules, (mandatory) access rules and all other (voluntary) measures to maximize the use of local loop wiring for transmission purposes.

Spectral management: The art of making optimal use of limited capacity in (metallic) access networks. This is for the purpose of achieving the highest reliable transmission performance and includes:

- Designing of deployment rules and their application.
- Designing of effective access rules.
- Optimised allocation of resources in the access network, e.g. access ports, diversity of systems between cable bundles, etc.
- Forecasting of noise levels for fine-tuning the deployment.
- Spectral policing to ensure network integrity.
- Making a balance between conservative and aggressive deployment (low or high failure risk).

Cable management plan (CMP): A list of selected access rules dedicated to a specific network. This list may include associated descriptions and explanations.

Spectral compatibility: A generic term for the capability of transmission systems to operate in the same cable. The precise definition is application dependent and has to be defined for each group of applications.

Cable fill: (or degree of penetration): number and mixture of connected transmission techniques to the ports of a binder or cable bundle that are injecting signals into the access ports.

Signal Category: is a class of signals meeting the minimum set of specifications identified in ETSI-TR-101-830-1. (NOTE: Some signal categories may distinct between different sub-classes, and may label them for instance as signals for "downstream" or for "upstream" purposes.)

PSD mask: The absolute upper bound of a PSD, measured within a specified resolution band. The purpose of PSD masks is usually to specify maximum PSD levels for stationary signals.

PSD template: The expected average values of the PSD of a stationary signal. The purpose of PSD templates is usually to perform simulations. The levels are usually below or equal to the associated PSD masks

Power back-off: is a generic mechanism to reduce the power. It has many purposes, including the reduction of power consumption, receiver dynamic range, cross talk, etc.

Power cut-back: is specific variant of power back-off, used to reduce the dynamic range of the receiver. It is characterized by a frequency independent reduction of the in-band PSD. It is used, for instance, in ADSL and SDSL.

EC: The abbreviation EC normally means Echo Cancelled. However, within the context of ADSL this abbreviation is used to designate ADSL systems with spectral overlap of downstream over upstream. In this context, the usage of the abbreviation "EC" was only kept for historical reasons. The usage of the echo cancelling technology is not only limited to spectrally overlapped systems, but can also be used by FDD systems.

NOTE: These definitions are intended to replace the definitions currently in part 1 (See [1], ETSI TR 101 830-1). Future revisions of part 1 will adopt these definitions as well.

3.2 Abbreviations

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

ADSL	Asymmetric Digital Subscriber Line
BER	Bit Error Ratio
CAP	Carrierless Amplitude/Phase modulation
CMP	Cable Management Plan (see clause 3.1 on definitions)
DMT	Discrete Multitone modulation
DFE	Decision Feedback Equalizer
EC	Echo Cancelled (see also under EC, in clause 3.1 on definitions)
FBL	Fractional Bit Loading (see clause 5.2.4 on DMT detection models)
FDD	Frequency Division Duplexing/Duplexed
GABL	Gain adjusted Bit Loading (see clause 5.2.4 on DMT detection models)
HDSL	High bit rate Digital Subscriber Line
ISDN	Integrated Services Digital Network
LT-port	Line Termination port (*commonly at central office side*)
LTU	Line Termination Unit
NT-port	Network Termination port (*commonly at customer side*)
NTU	Network Termination Unit
PAM	Pulse Amplitude modulation
PBO	Power Back-Off (see also clause 3.1 on definitions)
PCB	Power Cut-Back (see also clause 3.1 on definitions)
PSD	Power Spectral Density (single sided)
QAM	Quadrature Amplitude modulation
RBL	Rounded Bit Loading (see clause 5.2.4 on DMT detection models)
REC	Receiver
SDSL	Symmetrical (single pair high bit rate) Digital Subscriber Line
SNR	Signal to Noise Ration (*ratio of powers*)
TBD	To be defined / decided
TBL	Truncated Bit Loading (see clause 5.2.4 on DMT detection models)
TRA	Transmitter
VDSL	Very-high-speed Digital Subscriber Line
xDSL	(all systems) Digital Subscriber Line
2B1Q	2-Binary, 1-Quarternairy (*Special variant of a 4-level PAM line code*)

4 Transmitter signal models for xDSL

A transmitter model in this clause is mainly a PSD description of the transmitted signal under matched conditions, plus an output impedance description to cover mismatched conditions as well.

PSD *masks* of transmitted xDSL signals are specified in several documents for various purposes, for instance in Part 1 of Spectral Management [1]. These PSD masks, however, cannot be applied directly to the description of a transmitter model. One reason is that masks are specifying an upper limit, and not the expected (averaged) values. Another reason is that the definition of the true PSD of a time-limited signal requires no resolution bandwidth at all (it is defined by means of an autocorrelation, followed by a Fourier transform) while PSD *masks* do rely on some resolution bandwidth. They describe values that are (slightly) different from the true PSD; especially at steep edges (e.g. guard bands), and for modelling purposes this difference is sometimes very relevant.

To differentiate between several PSD descriptions, *masks* and *templates* of a PSD are given a different meaning. Masks are intended for proving compliance to standard requirements, while templates are intended for modelling purposes. This clause summarizes various xDSL transmitter models, by defining *template* spectra of output signals.

In some cases, models are marked as "default" and/or as "alternative". Both models are applicable, but in case a preference of either of them does not exist, the use of the "default" models is recommended. Other (alternative) models may apply as well, provided that they are specified.

4.1 Generic transmitter signal model

A generic model of an xDSL transmitter is essentially a linear signal source. The Thevenin equivalent of such a source equals an ideal voltage source U_s having a real resistor R_s in series. The output voltage of this source is random in nature (as a function of the time), is uncorrelated with any other transmitter signal, and occupies a relatively broad spectrum.

This generic model can be made specific by defining:

- The output impedance R_s of the transmitter.
- The template of the PSD, measured at the output port, when terminated with an external impedance equal to R_s. This is identified as the "matched condition", and under these conditions the output power equals the maximum power that is available from this source. Under all other (mis-matched) termination conditions the output power will be lower.

4.2 Cluster 2 transmitter signal models

4.2.1 Transmitter signal model for "ISDN.2B1Q"

The PSD template for modelling the "ISDN.2B1Q" transmit spectrum is defined by the theoretical sinc-shape of PAM encoded signals, with additional filtering and with a noise floor. The PSD is the maximum of both power density curves, as summarized in expression 1 and the associated table 1. The coefficient q_N scales the total signal power of $P_1(f)$ to a value that equals P_{ISDN}. This value is dedicated to the used filter characteristics, but equals q_N=1 when no filtering is applied ($f_L \to 0, f_H \to \infty$). The source impedance equals 135Ω.

$$P_1(f) = P_{ISDN} \times \frac{2 \times q_N}{f_X} \times \mathrm{sinc}^2\left(\frac{f}{f_X}\right) \times \frac{1}{1+\left(f/f_H\right)^{2\cdot N_H}} \times \frac{1}{1+\left(f_L/f\right)^2} \quad [W/Hz]$$

$$P_2(f) = \frac{10^{(P_{floor\_dBm}/10)}}{1000} \quad [W/Hz]$$

$$P(f) = \max\left(P_1(f), P_2(f)\right) \quad [W/Hz]$$

Where:

$P_{ISDN} = \left(10^{P_{ISDN\_dBm}/10}\right)/1000$ [W]

R_S = 135 [Ω]

$\mathrm{sinc}(x) = \sin(\pi \cdot x) / (\pi \cdot x)$

Default values for remaining parameters are summarized in table 1.

Expression 1: PSD template for modelling "ISDN.2B1Q" signals.

Different ISDN implementations, may use different filter characteristics, and noise floor values. Table 1 specifies *default* values for ISDN implementations, in case 2nd order Butterworth filtering has been applied. The default noise floor equals the maximum PSD level that meets the out-of-band specification of the ISDN standard [3].

Type	f_X [kHz]	f_H [kHz]	f_L [kHz]	N_H	q_N	P_{ISDN_dBm} [dBm]	P_{floor_dBm} [dBm/Hz]
ISDN.2B1Q	80	$1 \times f_X$	0	2	1.1257	13.5	–120

Table 1: Default parameter values for the ISDN.2B1Q templates, as defined in expression 1. These default values are based on 2nd order Butterworth filtering.

4.2.2 Transmitter signal model for "ISDN.MMS.43"

<for further study>

4.3 Cluster 3 transmitter signal models

4.3.1 Transmitter signal model for "HDSL.2B1Q"

The PSD templates for modelling the spectra of various "HDSL.2B1Q" transmitters are defined by the theoretical sinc-shape of PAM encoded signals, with additional filtering and a noise floor. The PSD template is the maximum of both power density curves, as summarized in expression 2 and associated table 2.

The coefficient q_N scales the total signal power of $P_1(f)$ to a value that equals P_0. This value is dedicated to the used filter characteristics, but equals q_N=1 when no filtering is applied ($f_L \rightarrow 0$, $f_H \rightarrow \infty$). The source impedance equals 135Ω.

$$P_1(f) = P_{HDSL} \times \frac{2 \times q_N}{f_X} \times \mathrm{sinc}^2\left(\frac{f}{f_X}\right) \times \frac{1}{1+\left(f_L/f\right)^2} \times \frac{1}{1+\left(f/f_{H1}\right)^{2 \cdot N_{H1}}} \times \frac{1}{1+\left(f/f_{H2}\right)^{2 \cdot N_{H2}}} \quad [W/Hz]$$

$$P_2(f) = \frac{10^{(P_{floor\_dBm}/10)}}{1000} \quad [W/Hz]$$

$$P(f) = \max\left(P_1(f), P_2(f)\right) \quad [W/Hz]$$

<u>Where:</u>

$P_{HDSL} = \left(10^{P_{HDSL\_dBm}/10}\right)/1000$ [W]

R_S = 135 [Ω]

sinc(x) = sin($\pi \cdot x$) / ($\pi \cdot x$)

Default values for remaining parameters are summarized in table 2.

Expression 2: PSD template for modelling "HDSL.2B1Q" signals.

Different HDSL implementations, may use different filter characteristics, and noise floor values. Table 2 summarizes *default* values for modelling HDSL transmitters, and *alternative* values in case higher order Butterworth filtering has been applied to dedicated implementations..

The default power level P_{HDSL} equals the maximum power allowed by the HDSL standard [4], since a nominal value does not exist in that standard. The default noise floor P_{floor} equals a value observed for various implementations.

Default

Name	Type	f_X kHz	f_L kHz	f_{H1}	N_{H1}	f_{H2}	N_{H2}	q_N	P_{HDSL_dBm} dBm	P_{floor_dBm} dBm/Hz
D1	HDSL.2B1Q/1	1160	3	$0.42\times f_x$	3	N/A	N/A	1.4662	14	–121.5
D2	HDSL.2B1Q/2	584	3	$0.50\times f_x$	3	N/A	N/A	1.3501	14	–133
D3	HDSL.2B1Q/3	392	3	$0.50\times f_x$	3	N/A	N/A	1.3642	14	–117

Alternatives

Name	Type	f_X kHz	f_L kHz	f_{H1}	N_{H1}	f_{H2}	N_{H2}	q_N	P_{HDSL_dBm} dBm	P_{floor_dBm} dBm/Hz
A1	HDSL.2B1Q/2	584	3	$0.68\times f_x$	4	N/A	N/A	1.1915	14	–133
A2	HDSL.2B1Q/2	584	3	$0.68\times f_x$	4	$1.50\times f_x$	2	1.1965	14	–133

Table 2: Parameter values for the HDSL.2B1Q templates, as defined in expression 2. The alternative values are based on higher order Butterworth filtering. Choose $f_{H2}=\infty$ and $N_{H2}=1$ when not applicable (N/A).

4.3.2 Transmitter signal model for "HDSL.CAP"

The PSD templates for modelling signals generated by HDSL.CAP transmitters are different for single-pair and two-pair HDSL systems. The PSD templates for modelling the "HDSL.CAP/2" and "HDSL.CAP/1" transmit spectra for two-pair and single-pair systems are defined in terms of break frequencies, as summarized in table 3. These template are taken from the nominal shape of the transmit signal spectra, as specified in the ETSI HDSL standard [4]
The associated values are constructed with straight lines between these break frequencies, when plotted against a <u>*logarithmic*</u> frequency scale and a <u>*linear*</u> dBm scale. The source impedance equals $R_s=135\Omega$.

HDSL.CAP/2	2-pair 135 Ω
[Hz]	[dBm/Hz]
1	–57
3,98 k	–57
21,5 k	–43
39,02 k	–40
237,58 k	–40
255,10 k	–43
272,62 k	–60
297,00 k	–70
1,188 M	–120
30 M	–120

HDSL.CAP/1	1- pair 135 Ω
[Hz]	[dBm/Hz]
<TBD>	<TBD>

Table 3. PSD template values at break frequencies for modelling "HDSL.CAP/2" and "HDSL.CAP/1".

NOTE: A PSD template for HDSL.CAP/1 is currently for further study.

4.3.3 Transmitter signal model for "SDSL"

The PSD templates for modelling the spectra of "SDSL" transmitters are defined by the theoretical sinc-shape of PAM encoded signals, plus additional filtering and a noise floor. The transmit spectrum is defined in three distinct frequency bands, as summarized in expression 3 and the associated table 4. (NOTE: These models are applicable to SDSL 16-UC-PAM at rates up to 2,312 Mb/s.)

The break frequency f_{int} is the frequency where the curves for P_1(f) and P_2(f) intersect. This PSD template is taken from the nominal shape of the transmit signal spectrum, as specified in the ETSI SDSL standard [5]. The source impedance equals R_s=135Ω.

$$f < f_{int}: \quad P_1(f) = \frac{K_{sdsl}}{R_s \times f_X} \times \text{sinc}^2\left(\frac{f}{f_X}\right) \times \frac{1}{1+\left(f/f_H\right)^{2\cdot N_H}} \times \frac{1}{1+\left(f_L/f\right)^2} \quad [W/Hz]$$

$$f_{int} \le f \le 1{,}5MHz: \quad P_2(f) = K_x \times \left(\frac{f}{f_0}\right)^{-1{,}5} \quad [W/Hz]$$

$$f > 1{,}5MHz: \quad P_3(f) = -110 \quad [dBm/Hz]$$

R_s = 135 Ω
sinc(x) = sin(π·x) / (π·x)
f_{int} = is the lowest frequency above f_H where the expressions for P_1(f) and P_2(f) intersect
Parameter values are defined in table 4

Expression 3. PSD template values for modelling both the symmetric and asymmetric modes of SDSL.

Mode	Data Rate *R* [kb/s]	TRA	Symbol Rate f_{sym} [kbaud]	f_X	f_H	f_L [kHz]	f_0 [Hz]	N_H	K_{SDSL} [V^2]	K_X [W/Hz]
Sym	< 2048	both	(*R*+ 8 kbit/s)/3	f_{sym}	f_X/2	5	1	6	7.86	$0.5683\cdot10^{-4}$
Sym	≥ 2048	both	(*R*+ 8 kbit/s)/3	f_{sym}	f_X/2	5	1	6	9.90	$0.5683\cdot10^{-4}$
Asym	2048	LTU	(*R*+ 8 kbit/s)/3	2×f_{sym}	f_X×2/5	5	1	7	16.86	$0.5683\cdot10^{-4}$
Asym	2048	NTU	(*R*+ 8 kbit/s)/3	f_{sym}	f_X×1/2	5	1	7	15.66	$0.5683\cdot10^{-4}$
Asym	2304	LTU	(*R*+ 8 kbit/s)/3	2×f_{sym}	f_X×3/8	5	1	7	12.48	$0.5683\cdot10^{-4}$
Asym	2304	NTU	(*R*+ 8 kbit/s)/3	f_{sym}	f_X×1/2	5	1	7	11.74	$0.5683\cdot10^{-4}$

Table 4. Parameter values for the SDSL templates, as defined in expression 3.

Power back-off (both directions)

The transmitter signal model includes a mechanism to cutback the power for short loops, and will be activated when the "Estimated Power Loss" (EPL) of the loop is below a threshold loss PL_{thres}. This EPL is defined as the ratio between the total transmitted power (in W), and the total received power (in W). This loss is usually expressed in dB as EPL_{dB}.

This power back-off *PBO* is equal for all in-band transmit frequencies, and is specified in expression 4. Mark that this model is based on a smooth cutback mechanism, although practical SDSL modems may cut back their power in discrete steps ("staircase"). This expression is simplified for simulation purposes. The SDSL power back-off is described in [5], (ETSI-SDSL, clause 9.2.6).

$$PBO_{dB} = \begin{cases} = 0dB & (if \ \ \Delta_{PL} < 0) \\ = \Delta_{PL} & (if \ \ 0 \le \Delta_{PL} \le 6dB) \\ = 6dB & (if \ \ \Delta_{PL} > 6dB) \end{cases} \qquad where \ \ \Delta_{PL} = \left(PL_{thres,dB} - EPL_{dB}\right)$$

Expression 4: Power back-off of the transmitted signal (in both directions), as a function of the estimated power loss (EPL) and a threshold loss of $PL_{thres,dB}$=6.5 db, and represents some average of the "staircase".

4.4 Cluster 4 transmitter signal models

4.4.1 Transmitter signal model for "ADSL over POTS"

The PSD template for modelling the "ADSL over POTS" transmit spectrum (EC variant) is defined in terms of break frequencies, as summarized in table 5. The associated values are constructed with straight lines between these break frequencies, when plotted against a _logarithmic_ frequency scale and a _linear_ dBm scale. The frequency Δf in this table refers to the spacing of the DMT sub carriers of ADSL. The source impedance equals 100Ω.

ADSL over POTS (EC) DMT carriers [k_1:k_2]	**Up [7:31]**
f **[Hz]**	***P*** **[dBm/Hz]**
0	–101
3.99k	–101
4 k	–96
6.5×Δf (≈ 28.03)	–38
31.5×Δf (≈ 135.84)	–38
53.0×Δf (≈ 228.56)	–90
686 k	–100
1.411M	–100
1.630M	–110
5.275M	–112
30M	–112
Δf = 4.3125 kHz	

ADSL over POTS (EC) DMT carriers [k_3:k_4]	**Down [7:255]**
f **[Hz]**	***P*** **[dBm/Hz]**
0	–101
3.99 k	–101
4 k	–96
6.5×Δf (≈ 28.03)	–40
255.5×Δf (≈ 1101.84)	–40
f_x = <TBD>	–90
3.093M	–90
4.545M	–112
30M	–112
Δf = 4.3125 kHz	

Table 5. PSD template values at break frequencies for modelling "ADSL over POTS".

NOTE The definition of a value f_x, representing the steepness of the downstream slope near 1.1 MHz, has been left for further study.
Values like f_x = 3093 kHz, based on the PSD mask specification in the standard, require a slope of at least –36 dB/octave. These values are seen as too pessimistic for a PSD template definition.
Values like f_x = 1201 kHz have been proposed as an alternative, and require a slope of at least –402 dB/octave. These values are seen as too optimistic and unrealistic.

Power cut back (downstream only)

The transmitter signal model includes a mechanism to cut-back the power for short loops, and will be activated when the band-limited power P_{rec}, received within a specified frequency band at the other side of the loop, exceeds a threshold value P_{thres}. This frequency band is from 6.5×Δf to 18.5×Δf, where Δf = 4.3125 kHz, and covers 12 consecutive sub carriers (7…18).

The cut back mechanism reduces the PSD template to a level PSD_{max}, as specified expression 5, for those frequencies where the downstream PSD template exceeds this level. For all other frequencies, the PSD template remains unchanged. Note that this model is based on a smooth cutback mechanism, although practical ADSL modems may cut back their power in discrete steps (“staircase”).

$$PSD_{\max,dBm} = \left\{ \begin{array}{l} = -40 dBm/Hz \\ = -40 dBm/Hz - 2\times\Delta_P \\ = -52 dBm/Hz \end{array} \right\} \quad \begin{array}{l} (if\ \ \Delta_P < 0dB) \\ (if\ \ 0 \le \Delta_P \le 6dB) \\ (if\ \ \Delta_P > 6dB) \end{array} \qquad where\ \Delta_P = \left(P_{rec,dBm} - P_{thres,dBm}\right)$$

Expression 5: Maximum PSD values of the transmitted downstream signal, as a function of the band-limited received power P_{rec} and a threshold level of $P_{thres,dBm}$ = 2.5 dBm, and represents some average of the “staircase”.

4.4.2 Transmitter signal model for "ADSL.FDD over POTS"

The PSD template for modelling "ADSL.FDD over POTS" transmit spectra is defined in terms of break frequencies, as summarized in table 6 and 7.

- Table 6 is to be used for modelling "guard band FDD modems", usually equipped with steep filtering for improving the separation between up and downstream signals. They leave 7 sub-carriers unused to enable this guard band.
- Table 7 is to be used for modelling "adjacent FDD modems", usually enhanced by echo cancellation for improving the separation between up and downstream signals. Because a guard band is not needed here, only 1 sub-carrier is left unused.

The associated values are constructed with straight lines between these break frequencies, when plotted against a <u>*logarithmic*</u> frequency scale and a <u>*linear*</u> dBm scale. The frequency Δf in this table refers to the spacing of the DMT sub-carriers of ADSL. The source impedance equals 100Ω.

Guard band FDD (using filters)

ADSL.FDD over POTS DMT carriers [k_1:k_2]	***Up [7:30]***
f [Hz]	***P [dBm/Hz]***
0	–101
3.99k	–101
4 k	–96
6.5×Δf (≈ 28.03)	–38
30.5×Δf (≈ 131.53)	–38
40.5×Δf (≈ 174.66)	–90
686 k	–100
1.411M	–100
1.630M	–110
5.275M	–112
30M	–112
Δf = 4.3125 kHz	

ADSL.FDD over POTS DMT carriers [k_3:k_4]	***Down [38:255]***
f [Hz]	***P [dBm/Hz]***
0	–101
3.99 k	–101
4 k	–96
27.5×Δf (≈ 118.59)	–96
37.0×Δf (≈ 159.56)	–47.7
37.5×Δf (≈ 161.72)	–40
255.5×Δf (≈ 1101.84)	–40
f_x = **<TBD>**	–90
3.093M	–90
4.545M	–112
30M	–112
Δf = 4.3125 kHz	

Table 6. PSD template values at break frequencies for modelling "ADSL.FDD over POTS", implemented as "guard band FDD" (with filtering). This PSD allocates 7 unused sub-carriers.

Adjacent FDD (using echo cancellation)

ADSL.FDD over POTS DMT carriers [k_1:k_2]	***Up [7:31]***
f [Hz]	***P [dBm/Hz]***
0	–101
3.99k	–101
4 k	–96
6.5×Δf (≈ 28.03)	–38
31.5×Δf (≈ 135.84)	–38
41.5×Δf (≈ 178.97)	–90
686 k	–100
1.411M	–100
1.630M	–110
5.275M	–112
30M	–112
Δf = 4.3125 kHz	

ADSL.FDD over POTS DMT carriers [k_3:k_4]	***Down [33:255]***
f [Hz]	***P [dBm/Hz]***
0	–101
3.99 k	–101
4 k	–96
22.5×Δf (≈ 97.03)	–96
32.0×Δf (≈ 138.00)	–47.7
32.5×Δf (≈ 140.16)	–40
255.5×Δf (≈ 1101.84)	–40
f_x = **<TBD>**	–90
3.093M	–90
4.545M	–112
30M	–112
Δf = 4.3125 kHz	

Table 7. PSD template values at break frequencies for modelling "ADSL.FDD over POTS", implemented as "adjacent FDD" (with echo cancelling). This PSD allocates 1 unused sub carrier, since a guard band is not required here.

NOTE. The definition of a value f_x, representing the steepness of the downstream slope near 1.1 MHz, has been left for further study. See the note in clause 4.4.1 for further details.

Power cutback (downstream only)

The transmitter signal model includes a mechanism to cutback the power for short loops, using the same mechanism as specified in expression 5, for modelling "ADSL over POTS" transmitters.

4.4.3 Transmitter signal model for "ADSL over ISDN"

The PSD template for modelling the "ADSL over ISDN" transmit spectrum (EC variant) is defined in terms of break frequencies, as summarized in table 8. The associated values are constructed with straight lines between these break frequencies, when plotted against a *logarithmic* frequency scale and a *linear* dBm scale. The frequency Δf in this table refers to the spacing of the DMT sub-carriers of ADSL. The source impedance equals 100Ω.

ADSL over ISDN (EC) DMT carriers [k_1:k_2]	***Up* [33:63]**
f [Hz]	***P [dBm/Hz]***
0	–90
50	–90
22.5×Δf (≈ 97.03)	–85.3
32.5×Δf (≈ 140.16)	–38
63.5×Δf (≈ 273,84)	–38
67.5×Δf (≈ 291.09)	–55
74.5×Δf (≈ 321.28)	–60
80.5×Δf (≈ 347.16)	–97.8
686	–100
1.411M	–100
1.630M	–110
5.275M	–112
30M	–112
Δf = 4.3125 kHz	

ADSL over ISDN (EC) DMT carriers [k_3:k_4]	***Down* [33:255]**
f [Hz]	***P [dBm/Hz]***
0	–90
50 k	–90
22.5×Δf (≈ 97.03)	–85.3
32.5×Δf (≈ 140.16)	–40
255.5×Δf (≈ 1101.84)	–40
f_x = **<TBD>**	–90
3.093M	–90
4.545M	–112
30M	–112
Δf = 4.3125 kHz	

Table 8. PSD template values at break frequencies for modelling "ADSL over ISDN (EC)".

NOTE. The definition of a value f_x, representing the steepness of the downstream slope near 1.1 MHz, has been left for further study. See the note in clause 4.4.1 for further details.

Power cutback (downstream only)

The transmitter signal model includes a mechanism to cut-back the power for short loops, and will be activated when the band-limited power P_{rec}, received within a specified frequency band at the other side of the loop, exceeds a threshold value P_{thres}. This frequency band is from 35.5×Δf to 47.5×Δf, where Δf = 4.3125 kHz, and covers 12 consecutive sub carriers (36…47).

The cut back mechanism reduces the PSD template to a level PSD_{max}, as specified expression 6, for those frequencies where the downstream PSD template exceeds this level. For all other frequencies, the PSD template remains unchanged. Note that this model is based on a smooth cutback mechanism, although practical ADSL modems may cut back their power in discrete steps (“staircase”).

$$PSD_{\max,dBm} = \begin{cases} = -40 dBm/Hz \\ = -40 dBm/Hz - \frac{4}{3} \times \Delta_P \\ = -52 dBm/Hz \end{cases} \quad \begin{matrix} (if \quad \Delta_P < 0dB) \\ (if \quad 0 \le \Delta_P \le 9dB) \\ (if \quad \Delta_P > 9dB) \end{matrix} \qquad where \ \Delta_P = (P_{rec,dBm} - P_{thres,dBm})$$

Expression 6: Maximum PSD values of the transmitted downstream signal, as a function of the band-limited received power P_{rec} and a threshold level of $P_{thres,dBm}$ = –0.75 dBm, and represents some average of the “staircase”.

4.4.4 Transmitter signal model for "ADSL.FDD over ISDN"

The PSD template for modelling "ADSL.FDD over ISDN" transmit spectra is defined in terms of break frequencies, as summarized in table 9 and 10.

- Table 9 is to be used for modelling "guard band FDD modems", usually enhanced by steep filtering for improving the separation between up and downstream signals. They leave 7 sub-carriers unused to enable this guard band.
- Table 10 is to be used for modelling "adjacent FDD modems", usually enhanced by echo cancellation for improving the separation between up and downstream signals. Because a guard band is not needed here, no sub-carrier is left unused.

The associated values are constructed with straight lines between these break frequencies, when plotted against a <u>*logarithmic*</u> frequency scale and a <u>*linear*</u> dBm scale. The frequency Δf in this table refers to the spacing of the DMT sub-carriers of ADSL. The source impedance equals 100Ω.

Guard band FDD (using filters)

ADSL.FDD over ISDN DMT carriers [k_1:k_2]	*Up [33:56]*
f [Hz]	***P [dBm/Hz]***
0	–90
50	–90
22.5×Δf = 97.03	–85.3
32.5×Δf = 140.16	–38
56.5×Δf = 243.66	–38
60.5×Δf = 260.91	–55
67.5×Δf = 291.09	–60
73.5×Δf = 316.97	–97.8
686	–100
1.411M	–100
1.630M	–110
5.275M	–112
30M	–112
Δf = 4.3125 kHz	

ADSL.FDD over ISDN DMT carriers [k_3:k_4]	*Down [64:255]*
f [Hz]	***P [dBm/Hz]***
0	–90
53.5×Δf = 230.72	–90
63.0×Δf = 271.79	–52
63.5×Δf = 273.84	–40
255.5×Δf = 1101.84	–40
f_x = **<TBD>**	–90
3.093M	–90
4.545M	–112
30M	–112
Δf = 4.3125 kHz	

Table 9. PSD template values at break frequencies for modelling "ADSL.FDD over ISDN", implemented as "guard band FDD" (with filtering). This PSD allocates 7 unused sub-carriers.

adjacent FDD (using echo cancellation)

ADSL.FDD over ISDN DMT carriers [k_1:k_2]	*Up [33:63]*
f [Hz]	***P [dBm/Hz]***
0	–90
50	–90
22.5×Δf = 97.03	–85.3
32.5×Δf = 140.16	–38
63.5×Δf = 273.84	–38
67.5×Δf = 291.09	–55
74.5×Δf = 321.28	–60
80.5×Δf = 347.16	–97.8
686	–100
1.411M	–100
1.630M	–110
5.275M	–112
30M	–112
Δf = 4.3125 kHz	

ADSL.FDD over ISDN DMT carriers [k_3:k_4]	*Down [64:255]*
f [Hz]	***P [dBm/Hz]***
0	–90
53.5×Δf = 230.72	–90
63.0×Δf = 271.79	–52
63.5×Δf = 273.84	–40
255.5×Δf = 1101.84	–40
f_x = **<TBD>**	–90
3.093M	–90
4.545M	–112
30M	–112
Δf = 4.3125 kHz	

Table 10. PSD template values at break frequencies for modelling "ADSL.FDD over ISDN", implemented as "adjacent FDD" (with echo cancelling). This PSD has no guard band.

NOTE. The definition of a value f_x, representing the steepness of the downstream slope near 1.1 MHz, has been left for further study. See the note in clause 4.4.1 for further details.

Power cutback (downstream only)

The transmitter signal model includes a mechanism to cutback the power for short loops, using the same mechanism as specified in expression 6, for modelling "ADSL over ISDN" transmitters.

4.5 Cluster 5 transmitter signal models

4.5.1 Transmitter signal model for "VDSL"

<for further study>

5 Generic receiver performance models for xDSL

A receiver performance model is capable of predicting up to what performance a data stream can be recovered from a noisy signal. In all cases it assumes that this recovery meets predefined quality criteria such as a maximum BER (Bit Error Ratio). Values like BER<10^{-7}, during a time interval of several minutes, are not uncommon.

The word *performance* refers within this context to a variety of quantities, including noise margin, signal margin and max data rate. When the receiver is ideal (zero internal receiver noise, infinite echo cancellation, etc), quantities like noise margin and signal margin become equal.

Performance models are implementation and line code specific. Performance modelling becomes more convenient when broken down into a combination of smaller sub models (see figure 1):

- A line code independent *input* (sub)model that evaluates the effective SNR from received signal, received noise, and various receiver imperfections. Details are described in clause 5.1.
- A line code dependent *detection* (sub)model that evaluates the performance (e.g. the noise margin at specified bit rate) from the effective SNR. Details are described in clause 5.2.

- An (optional) *echo-coupling* (sub)model that evaluates what portion of the transmitted signal flows into the receiver. Details are described in clause 5.3.

The flow diagram in figure 1 represents an xDSL transceiver that is connected via a common wire pair to another transceiver (not shown). This wire pair transports the transmitted signal, received signal and received noise simultaneously.

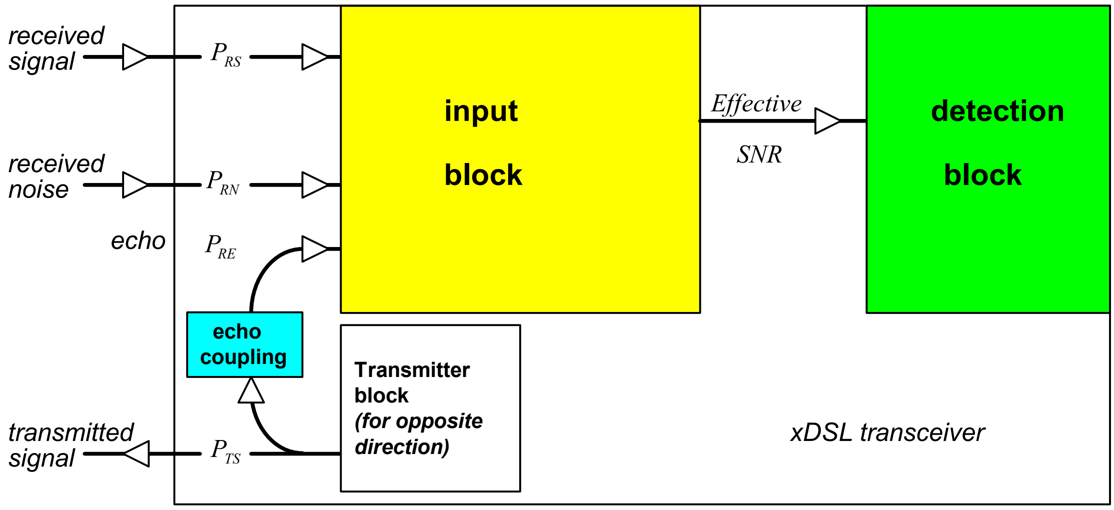


Figure 1: Flow diagram of a transceiver model, build up from individual sub models.

The input block of the flow diagram in figure 1 requires values for *signal*, *noise* and *echo*. The flow diagram illustrates this for an xDSL transceiver that is connected via a common wire pair to another transceiver (not shown), which transports the following three, flows simultaneously:

- The received *signal* power P_{RS} carries the data that is to be recovered. This signal originates from the transmitter at the other side of the wire pair, and its level is attenuated by cable loss.
- The received *noise* power P_{RN} is all that is received when the transmitters at both sides of the link under study are silent. The origin of this noise is mainly cross talk from internal disturbers connected to the same cable (cross talk noise), and partly from external disturbers (ingress noise).
- The received *echo* power P_{RE} is all that is received when the transmitter at the other end of the wire pair is silent, as well as all internal and external disturbers. It is a residue that will be received when a transmitter and a receiver are combined into a transceiver, and co-connected via a hybrid to the same wire pairs. When the hybrid of that transceiver is unbalanced due to mismatched termination impedances (of the cable), then a portion (P_{RE}) of the transmitted signal (P_{TS}) will leak into the receiver and is identified as echo.

The input block in figure 1 is to evaluates a quantity called ***effective SNR*** (Signal to noise Ratio) that indicates to what degree the received signal is deteriorated by noise, residual echo and all kinds of implementation imperfections. Due to signal processing in the receiver, the *input* SNR (the ratio between signal power, and the power-sum of noise and echo) will change into the *effective* SNR at some virtual internal point at the receiver. The effective SNR can be better or worse then the input SNR. Receivers with build-in echo cancellation can take advantage of a-priori knowledge on the echo, and can suppress most of this echo and thus improving the effective SNR. On the other hand, all analog receiver electronics produce shot noise and thermal noise, the A/D-converter produces quantization noise, and the equalization has its limitations as well. The combination of all these individual imperfections deteriorates the effective SNR.
In principle all parameters of the effective SNR can be assumed as frequency dependent, but this dependency has often been omitted here for reasons of simplicity. In addition, external change of signal and noise levels will modify the value of this effective SNR.

The detection block of the flow diagram in figure 1 requires this effective SNR to evaluate from that the performance as *margin* (such as noise margin, or signal margin). For many detection models, this margin is not provided by a closed expression, but by an equation from which this margin is to be solved. A simulation program may follow an iterative approach to solve this: controlling this margin in the input block so that the effective SNR changes and the equation in the detection block can be met.
In principle, the detection block is dedicated to line-code specific imperfections only, but may also include receiver imperfections that are not covered by the input block.

The echo-coupling block is optional, in case the input block does not cover the related imperfections. Simple (first order) models for the input block cannot distinct between receiver imperfection originated from echo and from other causes. When these simplified models are used, the echo-coupling block will not be required in the receiver performance model.

This chapter 5 details on (sub)models for above mentioned blocks in a receiver performance model, but is restricted to *generic* performance models only. Clause 6 is dedicated to implementation *specific* models by additionally assigning values to all parameters of a generic model.

5.1 Generic input models for effective SNR

An input (sub) model describes how to evaluate the effective SNR, as intermediate result (see figure 1), from various input quantities and imperfections. To simplify further analysis of performance quantities like *noise margin* and *signal margin*, the effective SNR is often expressed in its offset format, characterized by an additional parameter m. The associated expression is defined for each model individually.

With this parameter m the external noise level can be increased (for noise margin calculations) or the external signal level can be decreased (for signal margin calculations). The convention is that when m=1 (equals zero dB) the effective *offset* SNR equals the effective SNR itself. When the value of parameter m increases, the effective offset SNR decreases.

5.1.1 First order input model

This input model is quite a simplified model that assumes that the SNR of the input signal is internally modified by internal receiver noise (P_{RN0}). Most imperfections of the receiver (such as front-end noise, imperfect echo suppression, imperfect equalization and quantization noise) are assumed to be concentrated in a single virtual internal noise source (P_{RN0}). Figure 2 shows the flow diagram of an xDSL transceiver model showing the elements of a first order input model for effective SNR evaluation, and how to incorporate it in the receiver performance model.

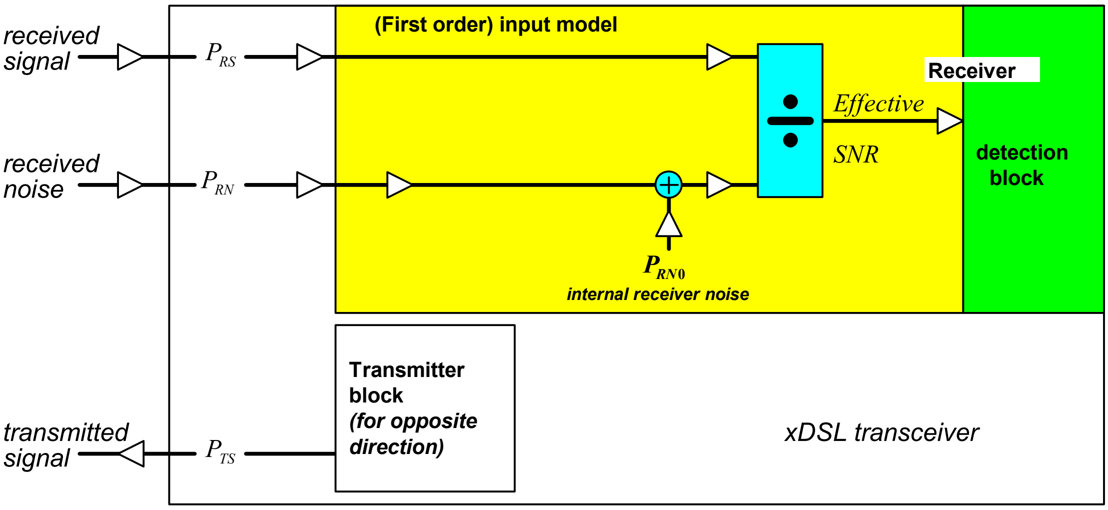


Figure 2: Flow diagram of a transceiver model that incorporates a linear first order input model for the determination of the effective SNR.

Expression 7 summarizes how to evaluate the effective SNR for this model, and it has been specified in plain and offset formats. Table 11 summarizes the involved parameters.

Plain format:	$SNR(f)$	$= \dfrac{P_{RS}}{P_{RN} + P_{RN0}}$
Noise offset format:	$SNR_{ofs,N}(m, f)$	$= \dfrac{P_{RS}}{P_{RN} \times m + P_{RN0}}$
Signal offset format:	$SNR_{ofs,S}(m, f)$	$= \dfrac{P_{RS} / m}{P_{RN} + P_{RN0}}$

Expression 7: Effective SNR, in various formats, when using the first order input model

Input quantities	linear	in dB	remarks
Received signal power	P_{RS}	$10\times\log_{10}(P_{RS})$	External signal
Received noise power (cross talk)	P_{RN}	$10\times\log_{10}(P_{RN})$	External noise
Model Parameters			
Internal receiver noise power	P_{RN0}	$10\times\log_{10}(P_{RN0})$	Internal noise
Output quantities			
Signal to noise ratio (effective)	SNR	$10\times\log_{10}(SNR)$	Frequency dependent

Table 11: Involved parameters and quantities for a first order input model. All PSD levels may be frequency dependent.

5.2 Generic detection models

This clause identifies several generic (sub) models for the detection block: one line code independent model derived from the Shannon capacity limit, and various line code dependent models dedicated to PAM, CAP/QAM or DMT line coding. Table 12 summarizes the naming convention for input and output quantities.

Input quantities	linear	in dB	remarks
Signal to Noise Ratio	*SNR*	$10 \times \log_{10}(SNR)$	Ratio of powers (frequency dependent)
Output quantities			
Noise margin	m_n	$10 \times \log_{10}(m_n)$	Ratio of noise powers
Signal margin	m_s	$10 \times \log_{10}(m_s)$	Ratio of signal powers

Table 12. Symbols used for input and output quantities of detection models.

On input, the detection block requires an effective SNR, as provided by the input block. This SNR is a function of the frequency f. When the offset format is used for describing the SNR, it will also be a function of the offset parameter m. This offset format is specified individually for each model in clause 5.1.

On output, the detection block evaluates a signal margin m_n (or a noise margin m_s when more appropriated). This margin parameter is a dominant measure for the transport quality that is achieved under noisy conditions.

- The *Noise Margin* m_n indicates how much the received noise power can increase before the transmission becomes unreliable.
- The *Signal Margin* m_s indicates how much the received signal power can decrease before the transmission becomes unreliable.

Unless explicitly specified otherwise, the word *margin* refers in this document to *noise margin*.

NOTE From an xDSL deployment point of view, the analysis of noise margin is preferred over signal margin, since the (cross talk) noise is the quantity that may increase when more systems are connected to the same cable. Many xDSL implementations, however, do report margin numbers that are not exactly equal to this noise margin, since the detection circuitry cannot make a distinction between external noise (due to cross talk) and internal noise (due to imperfect electronics). These margins are often an estimate closer in value to the signal margin then the noise margin.

5.2.1 Generic Shifted Shannon detection model

The calculation of the margin m using the generic Shifted Shannon detection model, is equivalent with solving the equation in expression 8. It has been derived from Shannon's capacity theorem, by reducing the effective SNR ("shifting" on a dB scale) by the SNR-gap Γ, to account for the imperfections of practical detectors. The associated parameters are summarized in table 13.

The effective SNR is to be evaluated by using one of the input models described in clause 5.1. Depending on what offset format is used for the SNR expression (see clause 5.1), the calculated margin m will represent the noise margin m_n or the signal margin m_s.

$$\boxed{f_b = \int_{f_c - B/2}^{f_c + B/2} \log_2\left(1 + \frac{SNR_{ofs}(m, f)}{\Gamma}\right) \cdot df}$$

Expression 8: Equation of the Shifted Shannon detection model, for solving the margin *m*.

Model Parameters	linear	in dB	remarks
SNR gap	Γ	$10\times\log_{10}(\Gamma)$	
Data rate	f_d		all payload bits that are transported in 1 sec
Line rate	f_b		= data rate + overhead bit rate
Center frequency	f_c		Center value of the most relevant spectrum
Bandwidth	B		Width of most relevant spectrum

Table 13. Parameters used for Shifted Shannon detection models.

The various parameters used within this generic detection model are summarized in table 13. The model can be made specific by assigning values to all these model parameters.

- The SNR-gap (Γ) is a performance parameter that indicates how close the detection approaches the Shannon capacity limit.
- The line rate is usually higher then the data rate (0…30%) to transport overhead bits for error correction, signalling and framing.
- The bandwidth is a parameter that indicates what frequency range of the received spectrum is relevant for data transport. The model assumes that only frequencies within this range can pass the receive filters.

5.2.2 Generic PAM detection model

The calculation of the margin m using the generic PAM detection model is equivalent with solving the equation in expression 9. This model assumes ideal decision feedback equalizer (DFE) margin calculations. The associated parameters are summarized in table 14.

The effective SNR is to be evaluated by using one of the input models described in clause 5.1. Depending on what offset format is used for the SNR expression (see clause 5.1), the calculated margin m will represent the noise margin m_n or the signal margin m_s.

$$\boxed{SNR_{req} = \Gamma\times\left(2^{2\cdot b}-1\right) = \exp\left(\frac{1}{f_s}\times\int_0^{f_s}\ln\left(1+\sum_{n=N_L}^{N_H} SNR_{ofs}\left(m, f+nf_s\right)\right)\cdot df\right)}$$

Expression 9: Equation of the PAM-detection model, for solving the margin *m*.

The SNR gap Γ, being used in the above expression 9, is a combination of various effects. This Γ parameter is often split-up into the following three parts:

- A theoretical modulation gap Γ_{PAM} (in the order of 9.75 dB, at BER=10^{-7})
- A theoretical coding gain Γ_{coding} (usually in the order of 3-5 dB), to indicate how much additional improvement is achieved by the chosen coding mechanism.
- An empirical implementation loss Γ_{impl} (usually 1.6 dB or more), indicating how much overall deterioration is caused by implementation dependent imperfections in echo cancellation, equalization, etc, without identifying its true cause.

When Γ is split-up into the above three parts, its value shall be evaluated as follows:

SNR gap (linear): $\Gamma = \Gamma_{PAM} \;/\; \Gamma_{coding} \times \Gamma_{impl}$

SNR gap (in dB): $\Gamma_{\_dB} = \Gamma_{PAM\_dB} - \Gamma_{coding\_dB} + \Gamma_{impl\_dB}$

Model Parameters	linear	in dB	remarks
SNR gap (effective)	Γ	$10\times\log_{10}(\Gamma)$	$= SNR_{req} / (2^{2\cdot b}-1)$
SNR gap in parts:	Γ_{PAM} Γ_{coding} Γ_{impl}	$10\times\log_{10}(\Gamma_{PAM})$ $10\times\log_{10}(\Gamma_{coding})$ $10\times\log_{10}(\Gamma_{impl})$	Modulation gap for PAM Coding gain Implementation loss
Required SNR	SNR_{req}	$10\times\log_{10}(SNR_{req})$	$= \Gamma\times(2^{2\cdot b}-1)$
Data rate	f_d		all payload bits that are transported in 1 sec
Line rate	f_b		= data rate + overhead bit rate
Symbol rate	f_s		$= f_b / b$
Bits per symbol	b		$= f_b / f_s$ (can be non-integer)
Summation range	N_L, N_H		On default: N_L=–2 and N_H=+1

Table 14. Parameters used for PAM detection models.

The various parameters in table 14 used within this generic detection model have the following meaning:

- The SNR-gap (Γ) and required SNR (SNR_{req}) are equivalent parameters and can be converted from one to the other. The advantage of using Γ over SNR_{req} is that Γ can be defined with similar meaning for all theoretical models in the frequency domain (Shifted Shannon, CAP, PAM, DMT). The advantage of using SNR_{req} over Γ is that this quantity is closer related to the SNR observed at the decision point of the detection circuitry.
- The line rate is usually higher then the data rate (0…30%) to transport overhead bits for error correction, signalling and framing. The symbol rate is the line rate divided by the number of bits packed together in a single symbol.
- The summation range for n is from N_L to N_H, and this range has to be defined to make this generic model specific. Commonly used values for PAM, using over sampling, are N_L=–2 and N_H=+1. This corresponds to T/3-spaced equalization. Wider ranges are not excluded.

5.2.3 Generic CAP/QAM detection model

The calculation of the margin m using the generic CAP/QAM detection model is equivalent with solving the equation in expression 10. This model assumes ideal decision feedback equalizer (DFE) margin calculations. The associated parameters are summarized in table 15.

The effective SNR is to be evaluated by using one of the input models described in clause 5.1. Depending on what offset format is used for the SNR expression (see clause 5.1), the calculated margin m will represent the noise margin m_n or the signal margin m_s.

$$\boxed{SNR_{req} \equiv \Gamma \times \left(2^b - 1\right) = \exp\left(\frac{1}{f_s} \times \int_0^{f_s} \ln\left(1 + \sum_{n=N_L}^{N_H} SNR_{ofs}\left(m, f + nf_s\right)\right) \cdot df\right)}$$

Expression 10: Equation of the CAP/QAM-detection model, for solving the margin *m*.

The (effective) SNR gap Γ, being used in the above expression 10, is a combination of various effects. This Γ parameter is often split-up into the following three parts:

- A theoretical modulation gap Γ_{CAP} (in the order of 9.8 dB for BER=10^{-7})
- A theoretical coding gain Γ_{coding} (usually in the order of 3-5 dB), to indicate how much additional improvement is achieved by the chosen coding mechanism.

- An empirical implementation loss Γ_{impl} (usually 1.6 dB or more), indicating how much overall deterioration is caused by implementation dependent imperfections in echo cancellation, equalization, etc, without identifying its true cause.

When Γ is split-up into the above three parts, its value shall be evaluated as follows:

SNR gap (linear): $\Gamma = \Gamma_{CAP} \;/\; \Gamma_{coding} \;\times\; \Gamma_{impl}$

SNR gap (in dB): $\Gamma_{\_dB} = \Gamma_{CAP\_dB} \;-\; \Gamma_{coding\_dB} \;+\; \Gamma_{impl\_dB}$

Model Parameters	linear	in dB	remarks
SNR gap (effective)	Γ	$10\times\log_{10}(\Gamma)$	$= SNR_{req} / (2^{b}-1)$
SNR gap in parts:	Γ_{CAP} Γ_{coding} Γ_{impl}	$10\times\log_{10}(\Gamma_{PAM})$ $10\times\log_{10}(\Gamma_{coding})$ $10\times\log_{10}(\Gamma_{impl})$	Modulation gap for CAP/QAM Coding gain Implementation loss
Required SNR	SNR_{req}	$10\times\log_{10}(SNR_{req})$	$= \Gamma\times(2^{-b}-1)$
Data rate	f_d		all payload bits that are transported in 1 sec
Line rate	f_b		= data rate + overhead bit rate
Symbol rate	f_s		$= f_b / b$
Bits per symbol	b		$= f_b / f_s$ (can be non-integer)
Summation range	N_L, N_H		On default: N_L=0 and N_H=+3

Table 15. Parameters used for CAP/QAM detection models.

The various parameters in table 15 used within this generic detection model have the following meaning:

- The SNR-gap (Γ) and required SNR (SNR_{req}) are equivalent parameters and can be converted from one to the other. The advantage of using Γ over SNR_{req} is that Γ can be defined with similar meaning for all theoretical models in the frequency domain (Shannon, CAP, PAM, DMT). The advantage of using SNR_{req} over Γ is that this quantity is closer related to the SNR observed at the decision point of the detection circuitry.

- The line rate is usually higher then the data rate (0..30%), to transport overhead bits for error correction, signalling and framing. The symbol rate is the line rate divided by the number of bits packed together in a single symbol.

- The summation range for n is from N_L to N_H, Commonly used values for CAP/QAM systems using over sampling are N_L=0 and N_H=+3. This holds when the carrier frequency positions the spectrum low in the frequency band (e.g. CAP-based HDSL). Other values may be more appropriated when the carrier frequency moves the spectrum to higher frequencies (e.g. CAP based VDSL).

5.2.4 Generic DMT detection model

The calculation of the margin m using the generic DMT detection model is equivalent with solving the equations in expression 11, for a given line rate f_b (or given *data* line rate f_{bd}). The associated parameters are summarized in table 16, and function *load* is specified by the chosen bit-loading algorithm. The effective SNR is to be evaluated by using one of the input models described in clause 5.1. Depending on what offset format $SNR_{ofs}(m, f)$ is used to express this effective SNR for margins other then m=1 (equals zero dB), the solved margin m will result in the noise margin m_n or the signal margin m_s.

$$
\begin{aligned}
b_k &= \log_2\left(1 + \frac{SNR_{ofs}(m, f_k)}{\Gamma}\right) && [bit/tone/symbol] \\
f_{bd} &= f_{sd} \times b = f_{sd} \times \sum_{k \in tones} load(b_k) && [bit/s] \\
f_b &= f_{bd} + f_{bs} && [bit/s]
\end{aligned}
$$

Expression 11: Equations of the DMT-detection model, for solving the margin *m* for a given *data* line rate f_{bd}, and a given *data* symbol rate f_{sd}. The latter excludes all DMT symbols dedicated to synchronisation.

Bit-loading algorithm

The DMT sub-carriers are all positioned (centred) at a multiple of the sub-carrier frequency spacing Δf, and each sub-carrier theoretically may carry any fragment of a symbol, while a symbol can carry many bits (typically a few hundred or more). The way this bit space (bits per tone per symbol) is used to load each sub-carrier with bits is implementation dependent.

Bit-loading algorithms do commonly use masking. Masking means skipping carriers for loading when their bit space b_k is below some predefined minimum value b_{min}, and limiting the bit-loading to some pre-defined maximum when the bit space b_k exceeds some predefined maximum b_{max}. This masking process is summarized in expression 12.

$b_k < b_{min}$	$\Rightarrow$	$load(b_k) \equiv 0$
$b_{min} \leq b_k \leq b_{max}$	$\Rightarrow$	$load(b_k) \equiv b_k$
$b_k > b_{max}$	$\Rightarrow$	$load(b_k) \equiv b_{max}$

Expression 12: The bit loading used in (fractional) bit-loading algorithms.

When the data transport is operating on its limits (margin m=1, or zero dB), the following bit-loading algorithms may apply, in addition to masking:

- *Fractional bit-loading* (FBL), sometimes referred to as *water-filling* - is a pure theoretical approach enabling loading of any real number of bits per symbol in any sub-carrier k (including non-integer fractions). This maximizes the use of the available capacity, but is unpractical to implement.
- *Truncated bit-loading* (TBL) - is a more feasible algorithm in practice, and loads on each sub-carrier k a number of bits equal to the largest non-negative integer *below* the bit space b_k.
- *Rounded bit-loading* (RBL) - is also feasible in practice, and loads each sub-carrier k a number of bits equal to the nearest non-negative integer of bit space b_k.
- *Gain adjusted bit-loading* (GABL) - is a sophisticated combination of rounded bit-loading and adjustment of powers to each of the sub-carriers, so that each individual bit space b_k approaches a rounded value (minimizes the loss of capacity), while the total transmit power is kept unchanged on average.

In various applications, it may be assumed that the capacity of well-designed *gain adjusted* bit-loading algorithms closely match those achieved by *fractional* bit-loading algorithm. For the sake of simplicity, and for making capacity calculations in this document less implementation dependent, the fractional bit-loading algorithm with constraint number of bits per sub-carrier and symbol, as in expression 12, is used as default for DMT calculations all over this document, unless specified explicitly otherwise.

NOTE When calculating the bit-loading, the used total power needs to be reduced by the amount of power spent on the cyclic extension. Text for detailed guidance to this note is currently under study.

SNR-Gap

The (effective) SNR gap Γ, being used in the above expression 11, is a combination of various effects. This Γ parameter is often split-up into the following three parts:

- A theoretical modulation gap Γ_{DMT} (in the order of 9.75 dB at BER = 10^{-7})
- A theoretical coding gain Γ_{coding} (usually in the order of 3 - 5 dB), to indicate how much additional improvement is achieved by using channel coding

- An empirical adjustment for all *unidentified* implementation losses Γ_{impl} (usually a few dB as well), indicating how much overall performance degradation is caused by implementation dependent imperfections (e.g. echo cancellation, analogue front end realization, equalization).

When Γ is split-up into the above three parts, its value shall be evaluated as follows:

SNR gap (linear): $\Gamma = \Gamma_{DMT} \;/\; \Gamma_{coding} \;\times\; \Gamma_{impl}$

SNR gap (in dB): $\Gamma_{\_dB} = \Gamma_{DMT\_dB} \;-\; \Gamma_{coding\_dB} \;+\; \Gamma_{impl\_dB}$

The margin value, which can be either noise margin or signal margin, is not included in the equations for SNR gap as it is contained in the offset SNR expression as described in clause 5.1.

Involved parameters

Input quantities	linear	in dB	remarks
Signal to Noise Ratio (effective value)	*SNR*	$10\times\log_{10}(SNR)$	Frequency dependent ratio of powers
Model Parameters	linear	in dB	remarks
SNR gap (effective)	Γ	$10\times\log_{10}(\Gamma)$	$= SNR_{req} / (2^{2\cdot b}-1)$
SNR gap in parts:	Γ_{DMT} Γ_{coding} Γ_{impl}	$10\times\log_{10}(\Gamma_{DMT})$ $10\times\log_{10}(\Gamma_{coding})$ $10\times\log_{10}(\Gamma_{impl})$	Modulation gap for DMT Coding gain Implementation loss
Symbol rate		f_s	Symbol rate, being the total number of *all* DMT symbols, transmitted in 1 second (Thus *data* symbols and *synch* symbols)
		f_{sd}	Symbol rate fragment, being the rate of *data* symbols only (without the overhead of synch symbols) that carry payloads bits
Line rate		f_b	Line rate, being the total number of *all* bits (for data, synch and other overhead) that is to be transported in 1 sec
		f_{bd}	Line rate fragment, caused by the bits in *data* symbols only
		f_{bs}	Line rate fragment, caused by the bits in *synch* symbols only
Data rate		f_d	All payload bits that are to be transported in 1 sec (also known as "net data bits")
Available set of sub-carriers		$\{k\}$	Can be a subset of all possible sub-carriers. (e.g. $k \in [7:255]$)
Center frequency location of tone k; k ∈ tones		f_k	$f_k = k\times\Delta f$ Δf = 4.3125 kHz in all current DMT systems
Bits per data symbol		$b = \Sigma\, b_k$	$b = f_{bd} / f_{sd}$ The bits of each data symbol are spread out over all used sub-carriers, in fragments of b_k
Bit-loading algorithm		 FBL TBL RBL GABL	Can be one of: • Fractional bit-loading (a.k.a. water filling) • Truncated bit-loading • Rounded bit-loading • Gain adjusted bit-loading
Minimum bit loading		b_{min}	Minimum number of bits per sub-carrier and per data symbol
Maximum bit loading		b_{max}	Maximum number of bits per sub-carrier and per data symbol
Output quantities	linear	in dB	remarks
Noise margin	m_n	$10\times\log_{10}(m_n)$	
Signal margin	m_s	$10\times\log_{10}(m_s)$	

Table 16: Parameters used for DMT detection models.

The various parameters in table 16, used within this generic detection model, have the following meaning:

- The SNR-gap (Γ) is a parameter that shows how far from the Shannon capacity limit a modem is performing at a certain bit error rate.

- The symbol rate f_s, in [baud] or [symbols/s], refers to *all* symbols being transmitted in one second. Most of them are so called *data* symbols, because they carrying bits for data transport, but after sending many data symbols, an additional *synch* symbol may be transmitted to keep the DMT transmission synchronized. The bits in each symbol are spread out over all involved DMT sub-carriers.
  The symbol rate is the sum of two fragments:
  - The *data* symbol rate f_{sd}, referring only to the rate of *data* symbols
  - The *synch* symbol rate f_{ss}, referring only to the rate of remaining *synch* symbols

  In ADSL, for example, one additional *synch* symbol is transmitted after sending 68 *data* symbols, and 4000 *data* symbols are transmitted in one second.
  In VDSL, for example, the *data* symbol rate and (total) symbol rate are equal as there is no extra synchronisation symbol as in ADSL.

- The line rate f_b [bits/s] refers to *all* bits being transmitted over the line in one second, including *all* overhead bits. Examples of overhead bits are bits for synchronization, all types of coding, the embedded operation channel, etc.
  Similar to the symbol rate, the line rate is the sum of two fragments:
  - The *data* line rate f_{bd}, refers to all bits in *data* symbols only, and covers payload bits as well as all overhead bits in a *data* symbol
  - The *synch* line rate f_{bs}, refers to all bits in the remaining *synch* symbols, and can be considered as 100% overhead for transporting payload bits.

  The bits in each symbol are spread out over the involved sub-carriers.

- The data rate f_d, in [bits/s], refers to the rate of *payload* bits only (also known as net data bits) that are to be transported by the DMT system. This rate does not include any transmission overhead, and is therefore lower then the line rate. Performance requirements are usually specified for these rates only, as for example the ETSI standard for ADSL [7].

- The available sub-carriers are specified by a list of integers, indicating what center frequencies are allocated to individual sub-carriers. For instance in ADSL it can contain any of the sub-carriers from tone 7 to tone 255.

- The center frequency of a sub-carrier k is $k \times \Delta f$, where Δf is the sub-carrier spacing.

- b_{min} and b_{max} are the minimum and maximum number of bits, respectively, used in the masking process of the bit loading.

5.3 Generic models for echo coupling

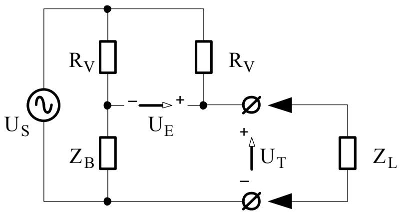
5.3.1 Linear echo coupling model

This model describes a property of linear hybrids in transceivers, and models what portion of the transmitted signal couples directly into the receiver. The hybrid is characterized by two parameters:

- R_V, representing the output impedance of the transceiver. Commonly used values are the design impedances of the modems under test, including as 100Ω for ADSL and 135Ω for SDSL.

- Z_B, representing the termination impedance that causes that the hybrid is perfectly balanced. This means that when the hybrid is terminated with this "balance impedance", no echo will flow into the receiver. For well-designed hybrids, this balance impedance is a "best guess" approximation of the "average" impedance of cables being used.

Figure 3 shows an equivalent circuit diagram of the above hybrid, represented as a Wheatstone bridge. The associated transfer function $H_E(j\omega)$ expresses what portion of the transmit signal will appear as echo.



$$H_E(j\omega) = \frac{U_E}{U_T} = 1 - \frac{Z_B}{R_V + Z_B} \times \frac{U_S}{U_T} = \frac{(Z_L - Z_B) \times R_V}{(R_V - Z_B) \times Z_L}$$

$$\frac{P_{RE}}{P_{TS}} = \left| H_E(j\omega) \right|^2$$

Figure 3: Flow diagram of the basic model for echo coupling. The identifiers P_{RE} and P_{TS} refer to power flow values used in figure 1.

When using this basic model for echo coupling in a full simulation, value R_V can be made equal to the design impedance of the modem under test, and value Z_B can be made equal to the complex and frequency dependent input impedance of the cable, terminated at the other cable end with a load impedance equal to R_V. Values for R_V and Z_B are implementation specific.

6 Specific receiver performance models for xDSL

This chapter defines parameter values for the generic performance models of the previous chapter 5, to provide implementation specific models for various xDSL modems.

6.1 Receiver performance model for "HDSL.2B1Q"

<left for further study>

6.2 Receiver performance model for "HDSL.CAP"

This calculation model is capable for predicting a performance that is benchmarked against the performance requirements of an ETSI compliant HDSL-CAP modem [4]. The reach predicted by this model, under the stress conditions (loss, noise) of the associated the ETSI HDSL specification [4], is close to the reach required by that ETSI specification.

The receiver performance model for ETSI compliant HDSL-CAP is build-up from the following building blocks:

- A first order (linear) input model for the input block, specified in clause 5.1.1, that combines all imperfections (front-end noise, residual echo, equalization errors), in one virtual noise source.
- The generic CAP/QAM detection model, specified in clause 5.2.3.
- The parameter values specified in table 17 of the succeeding clause.

The parameter values, used in the receiver performance model for ETSI compliant HDSL-CAP, are summarized in table 17. Parts of them are directly based on HDSL specifications. The remaining values are based on theory, followed by an iterative fit of the model to meet the ETSI reach requirements for HDSL-CAP under the associated stress conditions.

Various parameters are derived directly from the above-mentioned parameters. Their purpose is to simplify the required expression of the used CAP/QAM-detection model.

Model Parameter		HDSL.CAP/2	HDSL.CAP/1
SNR-Gap (effective)	$\Gamma_{\_dB}$	6.8 dB	6.8 dB
SNR-Gap in parts	Γ_{CAP_dB} Γ_{coding_dB} Γ_{impl_dB}	9.8 dB <TBD> <TBD>	9.8 dB <TBD> <TBD>
Receiver noise	P_{RN0_dB}	–105 dBm	–105 dBm
Data rate	f_d	2×1024 kb/s	1×2048 kb/s
Line rate	f_b	1168 kb/s	2330 kb/s
Carrier frequency	f_c	138.30 kHz	226.33 kHz
bits per symbol	b	5	6
Summation bounds in the CAP/QAM model	N_H N_L	+3 0	+3 0
Derived Parameter			
Required SNR	SNR_{req}	$\Gamma\times(2^b-1)$	$\Gamma\times(2^b-1)$
	SNR_{req_dB}	≈ 21.7 dB	≈ 24.8 dB
Symbol rate	f_s	f_b / b = 233.6 kbaud	f_b / b = 388.3 kbaud

Table 17. Values for the parameters of the performance model, obtained from ETSI requirements for HDSL-CAP [4].

6.3 Receiver performance model for "SDSL"

This calculation model is capable for predicting a performance that is benchmarked against the performance requirements of an ETSI compliant SDSL modem [5]. The reach predicted by this model, under the stress conditions (loss, noise) of the associated the ETSI SDSL specification [5] is close to the reach required by that ETSI specification. Deviations of predictions and requirements are less then 4.5% in reach, and less then 125m. The validity of the predicted performance holds for a wider range of stress conditions. (NOTE: These models are applicable to SDSL 16-UC-PAM at rates up to 2,312 Mb/s.)

The receiver performance model for ETSI compliant SDSL is build-up from the following building blocks:

- A first order (linear) input model for the input block, specified in clause 5.1.1, that combines all imperfections (front-end noise, residual echo, equalization errors), in one virtual noise source (P_{RN0}).
- The generic PAM detection model, specified in clause 5.2.2.
- The parameter values specified in table 18 of the succeeding clause.

The parameter values, used in the receiver performance model for ETSI compliant SDSL, are summarized in table 18. Part of them is directly based on SDSL specifications. The remaining values are based on theory.

Various parameters are derived from the above-mentioned parameters. Their purpose is to simplify the required expression of the used PAM-detection model.

Model parameter		SDSL model	
		≤ 256 kb/s	> 256 kb/s
SNR-Gap (effective)	$\Gamma_{\_dB}$	6.95 dB	6.25 dB
SNR-Gap in parts	Γ_{PAM_dB} Γ_{coding_dB} Γ_{impl_dB}	9.75 dB 4.4 dB 1.6 dB	9.75 dB 5.1 dB 1.6 dB
Receiver noise	P_{RN0_dB}	–140 dBm	
Data rate	f_d	192 … 2304 kb/s	
Line rate	f_b	f_d + 8 kb/s	
bits per symbol	b	3	
Summation bounds in PAM model	N_H N_L	+1 –2	
Derived Parameter			
Required SNR	SNR_{req}	$\Gamma \times (2^{2b}-1)$	
	SNR_{req_dB}	≈18 dB	
Symbol rate	f_s	f_b / 3	

Table 18. Values for the parameters of the performance model, obtained from ETSI requirements for SDSL [5]. The echo suppression is captured in the overall implementation loss ($\Delta\Gamma_{impl}$).

6.4 Receiver performance model for "ADSL over POTS" (EC)

This calculation model is capable of predicting a performance that is benchmarked against the performance requirements of an ETSI compliant “ADSL over POTS” modem. The reach predicted by this model, under the stress conditions of the associated ETSI ADSL specification [7], is close to the minimum reach required by that ETSI specification. Deviations between the predicted reach and this "benchmarked" reach are less then 100m. The validity of the predicted performance holds for a wider range of stress conditions.

The receiver performance model for ETSI compliant “ADSL over POTS” is build-up from the following building blocks:

- A first order (linear) input model for the input block specified in clause 5.1.1, that combines all kinds of imperfections (front-end noise, residual echo and equalization errors), in one virtual noise source (P_{RN0}).
- The generic DMT detection model, specified in clause 5.2.4.
- The parameter values specified in table 19 of the succeeding clause.

The parameter values, used in the receiver performance model for ETSI compliant “ADSL over POTS” modems, are summarized in table 19. Parts of them are directly based on ADSL specifications. The remaining values are based on theory.

Model parameter		DMT	model	
		Upstream	Downstream	Remarks
SNR-Gap (effective)	Γ_{dB}	7.5 dB	7.5 dB	
SNR-Gap in parts	Γ_{DMT_dB} Γ_{coding_dB} Γ_{impl_dB}	9.75 dB 4.25 dB 2.0 dB	9.75 dB 4.25 dB 2.0 dB	
Receiver noise	P_{RN0_dB}	–120 dBm	–135 dBm	
Symbol rate	f_s	69/68 × 4000baud	69/68 × 4000 baud	*See clause 5.2.4*
	f_{sd}	4000 baud	4000 baud	
Data rate	f_d	32 … 640 kb/s	32 … 6144 kb/s	These are minimum ranges only; wider ranges are usually supported
Line rate	f_{bd}	$f_{bl} = f_d + 16 \times f_{sd}$ $f_{bh} = (f_d + 8 \times f_{sd}) \times 1.13$ $f_{bd} = max(f_{bl} , f_{bh})$	$f_{bl} = f_d + 16 \times f_{sd}$ $f_{bh} = (f_d + 8 \times f_{sd}) \times 1.13$ $f_{bd} = max(f_{bl} , f_{bh})$	*See clause 5.2.4*
	f_b	$f_b = 69/68 \times f_{db}$	$f_b = 69/68 \times f_{db}$	
Bits per symbol	b	f_{bd} / f_{sd}	f_{bd} / f_{sd}	
Available set of sub-carriers	$\{k\}$	$k \in [7:31]$	$k \in [7:63 , 65:255]$	*DMT tone k = 64 does not convey any bits because it is reserved as pilot tone.*
Center frequency location of tone k; k ∈ tones	f_k	$f_k = k \times \Delta f$ Δf = 4.3125 kHz	$f_k = k \times \Delta f$ Δf = 4.3125 kHz	
Bit-loading algorithm		FBL	FBL	*See (clause 5.2.4)*
Minimum bit-loading	b_{min}	2	2	*Bits per tone per symbol*
Maximum bit-loading	b_{max}	15	15	*Bits per tone per symbol*

Table 19: Values for the performance parameters extracted from the ETSI performance requirements under ETSI stress conditions.

6.5 Receiver performance model for "ADSL.FDD over POTS"

The receiver performance models for ETSI compliant “ADSL.FDD over POTS” are build-up from the following building blocks:

- A first order (linear) input model for the input block specified in clause 5.1.1, that combines all kinds of imperfections (front-end noise, residual echo and equalization errors), in one virtual noise source (P_{RN0}).
- The generic DMT detection model, specified in clause 5.2.4.
- The parameter values specified in table 20.

The model assumes a guard band of 7 tones between up and downstream, and this guard band makes additional modelling of imperfections in echo suppression irrelevant.

The parameter values, used in the receiver performance model for ETSI compliant “ADSL.FDD over POTS” modems, are summarized in table 20. Parts of them are directly based on ADSL specifications. The remaining values are extracted from the ADSL performance requirements [7] or based on theory.

Model parameter		DMT	model	
		Upstream	Downstream	Remarks
SNR-Gap (effective)	Γ_{dB}	9.3 dB	8.9 dB	
SNR-Gap in parts	Γ_{DMT_dB} Γ_{coding_dB} Γ_{impl_dB}	9.75 dB 4.25 dB 4.3 dB	9.75 dB 4.25 dB 3.9 dB	
Receiver noise	P_{RN0_dB}	–120 dBm	–140 dBm	
Symbol rate	f_s f_{sd}	69/68 × 4000baud 4000 baud	69/68 × 4000 baud 4000 baud	See clause 5.2.4
Data rate	f_d	32 … 640 kb/s	32 … 6144 kb/s	These are minimum ranges only; wider ranges are usually supported
Line rate	 f_{bd} f_b	$f_{bl} = f_d + 16 \times f_{sd}$ $f_{bh} = (f_d + 8 \times f_{sd}) \times 1.13$ $f_{bd} = \max(f_{bl}, f_{bh})$ $f_b = 69/68 \times f_{db}$	$f_{bl} = f_d + 16 \times f_{sd}$ $f_{bh} = (f_d + 8 \times f_{sd}) \times 1.13$ $f_{bd} = \max(f_{bl}, f_{bh})$ $f_b = 69/68 \times f_{db}$	See clause 5.2.4
Bits per symbol	b	f_{bd} / f_{sd}	f_{bd} / f_{sd}	
Available set of tones	$\{k\}$	see table [21]	see table [21]	
Center frequency location of tone k; $k \in$ tones	f_k	$f_k = k \times \Delta f$ Δf = 4.3125 kHz	$f_k = k \times \Delta f$ Δf = 4.3125 kHz	
Bit-loading algorithm		FBL	FBL	See clause 5.3.4
Minimum bit-loading	b_{min}	<TBD> (see note)	<TBD> (see note)	Bits per tone per symbol
Maximum bit-loading	b_{max}	<TBD> (see note)	<TBD> (see note)	Bits per tone per symbol

Table 20: Values for the performance parameters extracted from the minimum ETSI performance requirements under ETSI stress conditions.

ADSL.FDD over POTS versions	Available set of tones {k}, for upstream	Available set of tones {k}, for downstream
Guard band FDD	$k \in [7:30]$	$k \in [38:63, 65:255]$
Adjacent FDD	$k \in [7:31]$	$k \in [33:63, 65:255]$

Table 21: Set of sub carriers, available for different versions of "ADSL.,FDD over POTS" (with or without any guard band between up and downstream). DMT tone k = 64 does not convey any bits because it is reserved as pilot tone.

NOTE The ADSL standard specifies the bit-loading as integer values between 2 and 15, however the use of a model with "Fractional" bit-loading enables the use of non-integer values to account for other receiver properties as well. This enables the modelling of other receiver characteristics, as if the bit-loading caused them.

Using values for minimum bit-loading between 1.5 and 2 may account for the power adjustment of individual levels that minimizes the loss of capacity. A value of 1.5 may be too optimistic and a value of 2 may be too pessimistic, and therefore this level has been left for further study,

Using values for maximum bit-loading lower than 15 may account for imperfections in the equalizer causing an upper limit of the effective SNR at the detector. Practical implementations of ADSL that facilitate effective SNR values above 55 dB can take advantage of the full bit-loading range (up to 15). The ETSI reach requirements, however, are based on bit rates for short loops (high SNR) that are significantly lower then expected from effective SNR values better then 55 dB. Therefore the value for this maximum bit-loading has been left for further study.

When this model is used for simulation purposes, the chosen values for minimum and maximum bit-loading shall be specified.

6.6 Receiver performance model for "ADSL over ISDN" (EC)

This calculation model is capable of predicting a performance that is benchmarked against the performance requirements of an ETSI compliant “ADSL over ISDN” modem. The reach predicted by this model, under the stress conditions of the associated ETSI ADSL specification [7], is close to the minimum reach required by that ETSI specification. Deviations between the predicted reach and this "benchmark" reach are in most cases less then 80m. The validity of the predicted performance holds for a wider range of stress conditions.

The receiver performance model for ETSI compliant “ADSL over ISDN” is build-up from the following building blocks:

- A first order (linear) input model for the input block specified in clause 5.1.1, that combines all kinds of imperfections (front-end noise, residual echo and equalization errors), in one virtual noise source (P_{RN0}).
- The generic DMT detection model, specified in clause 5.2.4.
- The parameter values specified in table 22 of the succeeding clause.

The parameter values, used in the receiver performance model for ETSI compliant “ADSL over ISDN” modems, are summarized in table 22. Parts of them are directly based on ADSL specifications. The remaining values are based on theory.

Model parameter		DMT	model	
		Upstream	Downstream	Remarks
SNR-Gap (effective)	Γ_{dB}	7.8 dB	7.5 dB	
SNR-Gap in parts	Γ_{DMT_dB} Γ_{coding_dB} Γ_{impl_dB}	9.75 dB 4.25 dB 2.3 dB	9.75 dB 4.25 dB 2.0 dB	
Receiver noise	P_{RN0_dB}	–120 dBm	–135 dBm	
Symbol rate	f_s	69/68 × 4000 baud	69/68 × 4000 baud	*See clause 5.2.4*
	f_{sd}	4000 baud	4000 baud	
Data rate	f_d	32 … 640 kb/s	32 … 6144 kb/s	These are minimum ranges only; wider ranges are usually supported
Line rate	f_{bd}	$f_{bl} = f_d + 16 \times f_{sd}$ $f_{bh} = (f_d + 8 \times f_{sd}) \times 1.13$ $f_{bd} = \max(f_{bl}, f_{bh})$	$f_{bl} = f_d + 16 \times f_{sd}$ $f_{bh} = (f_d + 8 \times f_{sd}) \times 1.13$ $f_{bd} = \max(f_{bl}, f_{bh})$	*See clause 5.2.4*
	f_b	$f_b = 69/68 \times f_{db}$	$f_b = 69/68 \times f_{db}$	
Bits per symbol	b	f_{bd} / f_{sd}	f_{bd} / f_{sd}	
Available set of sub-carriers	$\{k\}$	$k \in [33:63]$	$k \in [33:95, 97:255]$ *Tone 96 = pilot tone*	*DMT tone k = 96 does not convey any bits because it is reserved as pilot tone.*
Center frequency location of tone k; $k \in$ tones	f_k	$f_k = k \times \Delta f$ Δf = 4.3125 kHz	$f_k = k \times \Delta f$ Δf = 4.3125 kHz	
Bit-loading algorithm		FBL	FBL	*See (clause 5.2.4)*
Minimum bit-loading	b_{min}	2	2	*Bits per tone per symbol*
Maximum bit-loading	b_{max}	15	15	*Bits per tone per symbol*

Table 22: Values for the performance parameters extracted from the ETSI performance requirements under ETSI stress conditions.

6.7 Receiver performance model for "ADSL.FDD over ISDN

The downstream receiver performance model for ETSI compliant “ADSL.FDD over ISDN” is build-up from the following building blocks:

- A first order (linear) input model for the input block specified in clause 5.1.1, that combines all kinds of imperfections (front-end noise, residual echo and equalization errors), in one virtual noise source (P_{RN0}).
- The generic DMT detection model, specified in clause 5.2.4.
- The parameter values specified in table 23.

The model assumes a guard band of 7 tones between up and downstream, and this guard band makes additional modelling of imperfections in echo suppression irrelevant.

The parameter values, used in the receiver performance model for ETSI compliant “ADSL.FDD over ISDN” modems, are summarized in table 23. Parts of them are directly based on ADSL specifications. The remaining values are extracted from the ADSL performance requirements [7] or based on theory.

Model parameter		DMT	model	
		Upstream	Downstream	Remarks
SNR-Gap (effective)	Γ_{dB}	9.6 dB	9.0 dB	
SNR-Gap in parts	Γ_{DMT_dB} Γ_{coding_dB} Γ_{impl_dB}	9.75 dB 4.25 dB 4.6 dB	9.75 dB 4.25 dB 4.0 dB	
Receiver noise	P_{RN0_dB}	–120 dBm	–140 dBm	
Symbol rate	f_s f_{sd}	69/68 × 4000 baud 4000 baud	69/68 × 4000 baud 4000 baud	*See clause 5.2.4*
Data rate	f_d	32 … 640 kb/s	32 … 6144 kb/s	*These are minimum ranges only; wider ranges are usually supported*
Line rate	f_{bd}	$f_{bl} = f_d + 16 \times f_{sd}$ $f_{bh} = (f_d + 8 \times f_{sd}) \times 1.13$ $f_{bd} = \max(f_{bl}, f_{bh})$	$f_{bl} = f_d + 16 \times f_{sd}$ $f_{bh} = (f_d + 8 \times f_{sd}) \times 1.13$ $f_{bd} = \max(f_{bl}, f_{bh})$	*See clause 5.2.4*
	f_b	$f_b = 69/68 \times f_{db}$	$f_b = 69/68 \times f_{db}$	
Bits per symbol	b	f_{bd} / f_{sd}	f_{bd} / f_{sd}	
Available set of tones	$\{k\}$	See table 24	See table 24	
Center frequency location of tone k; k ∈ tones	f_k	$f_k = k \times \Delta f$ Δf = 4.3125 kHz	$f_k = k \times \Delta f$ Δf = 4.3125 kHz	
Bit-loading algorithm		FBL	FBL	*See clause 5.2.4*
Minimum bit-loading	b_{min}	<**TBD**> (*see note*)	<**TBD**> (see note)	*Bits per tone per symbol*
Maximum bit-loading	b_{max}	<**TBD**> (see note)	<**TBD**> (see note)	*Bits per tone per symbol*

Table 23: Values for the performance parameters extracted from the ETSI performance requirements under ETSI stress conditions.

ADSL.FDD over ISDN versions	Available set of tones {k}, for upstream	Available set of tones {k}, for downstream
Guard band FDD	$k \in [33:56]$	$k \in [64:95 , 97:255]$
Adjacent FDD	$k \in [33:63]$	$k \in [64:95 , 97:255]$

Table 24: Set of sub carriers, available for different versions of "ADSL.FDD over ISDN" (with or without any guard band between up and downstream). DMT tone k = 96 does not convey any bits because it is reserved as pilot tone.

NOTE The ADSL standard specifies the bit-loading as integer values between 2 and 15, however the use of a model with "Fractional" bit-loading enables the use of non-integer values to account for other receiver properties as well. This enables the modelling of other receiver characteristics, as if the bit-loading caused them.

Using values for minimum bit-loading between 1.5 and 2 may account for the power adjustment of individual levels that minimizes the loss of capacity. A value of 1.5 may be too optimistic and a value of 2 may be too pessimistic, and therefore this level has been left for further study,

Using values for maximum bit-loading lower than 15 may account for imperfections in the equalizer causing an upper limit of the effective SNR at the detector. Practical implementations of ADSL that facilitate effective SNR values above 55 dB can take advantage of the full bit-loading range (up to 15). The ETSI reach requirements, however, are based on bit rates for short loops (high SNR) that are significantly lower then expected from effective SNR values better then 55 dB. Therefore the value for this maximum bit-loading has been left for further study.

When this model is used for simulation purposes, the chosen values for minimum and maximum bit-loading shall be specified.

6.8 Receiver performance model for "VDSL"

<left for further study>

7 Transmission and reflection models

7.1 Summary of test loop models

This section is for further study, and is intended to refer to various cable models, being published in several documents

8 Cross talk models

Cross talk models account for the fact that the transmission is impaired by cross talk originated from discrete disturbers distributed over the local loop wiring. In practice this is not restricted to a linear cable topology, since wires may fan out into different directions to connect for instance different customers to a central office

The most simple topology models assume that all disturbers are co-located at only two locations; one at each end of a cable. These approximations may be adequate for situations above for instance 1 km in which the fan out of the wires can be ignored.

More advanced topology models require a multi-node co-location approach. An example is the insertion of repeaters that introduces co-located disturbers in-between. Another example is deploying VDSL from the cabinet for the situation that all customers are distributed along the cable.

This clause summarizes different cross talk models for different topologies, sorted by complexity, and provides several cross talk models to predict how much noise is coupled into a victim wire pair.

8.1 Overview of different network topologies

Simulation result is highly dependent on the chosen network topology, which is very country and location specific. A summary of those topologies is for further study.

8.2 Validity limitations of cross talk modelling

<for further study>

8.3 Generic cross talk models for two-node co-location

The cross talk models in this sub clause apply to scenarios in which it can be assumed that all customers are virtually co-located, and that they are all served from the central office. The result is that such a cross talk model requires only two nodes (one on the LT side, and another one on the "common" NT side). These nodes are interconnected by means of a multi wire pair cable.

It is recommended to use the multi-node approach in clause 8.4, because this makes the simulation more realistic. A two-node approach implies that all NT disturbers are virtually collocated, and this is too pessimistic. This may cause pessimistic downstream results. Cross talk models are built up from several building blocks, and the way these blocks are interconnected is defined by means of a topology diagram.

8.3.1 Basic diagram for two-node topologies

The basic flow diagram for describing a topology in which xDSL equipment is assumed to be co-located at two nodes (the two ends of a cable) is shown in figure 4 and 5. Up and downstream performance are evaluated separately. The approach of this diagram can be described in three distinct steps.

- The diagram combines for each node the output disturbance of individual disturbers (P_{d1}, P_{d2}, …) by modelling *cross talk cumulation* as an isolated building block. This is because the cumulation from different disturbers cannot be modelled by a simple *linear* power sum of all individual disturbers. Since each wire pair couples at a different ratio to the victim wire pair, the cumulation requires some *weighted* power sum that accounts for the statistical distribution of all involved cross talk coupling ratios.
  By modelling cross talk cumulation as an isolated building block, the cumulated disturbance can be thought as if it was virtually generated by a single equivalent disturber ($P_{d.eq}$). This has been indicated in figure 4 and 5 by a box drawn around the involved building blocks. Using the equivalent disturber concept as intermediate yields an elegant concept to break down the complexity of a full noise scenario into smaller pieces.

- Next, the diagram evaluates what noise level (P_{XN}) is coupled into the victim wire pair. Figure 4 and 5 illustrate what portion of the equivalent disturbance is coupled into the victim wire pair by using models for *NEXT* and *FEXT*. On top of this, background noise (P_{bn}) can be added to represent all remaining unidentified noise sources. Since it is a generic diagram, the power level of this background noise level is left undefined here, but commonly used values are zero, or levels as low as P_{bn}=–140 dBm/Hz.

- When all building blocks are modelled for the same impedance as implemented in the modem under study, the noise level (P_{RN}) received by the modem under test equals the level derived so far (P_{XN}). In practice, these models are normalized at some chosen reference impedance R_n, and this R_n may be different from the impedance implemented in the modem under study (targeted at its design impedance R_V). This "mismatch" will cause a change in the level of the disturbance, and this effect is modelled by the noise injection building block.

The succeeding clauses summarizes some generic models for the individual building blocks of figure 4 and 5.

The transfer functions H_{next} and H_{fext} of the building blocks for NEXT and FEXT are linear and frequency dependent. The model for the topology assumes that all disturbers are uncorrelated, which causes that the cross talk power P_{XN} behind the summation block is the sum of all individual powers. This transfer functions are specified in expression 13.

$$\begin{aligned} P_{XN,NT} &= P_{d.eq,NT} \times \left|H_{next}\right|^2 + P_{d.eq,LT} \times \left|H_{fext}\right|^2 + P_{bn,NT} \\ P_{XN,LT} &= P_{d.eq,LT} \times \left|H_{next}\right|^2 + P_{d.eq,NT} \times \left|H_{fext}\right|^2 + P_{bn,LT} \end{aligned}$$

Expression 13: Evaluation of the cross talk power levels, that flow into the noise injection blocks of the two-node topology models in figure 4 and 5.

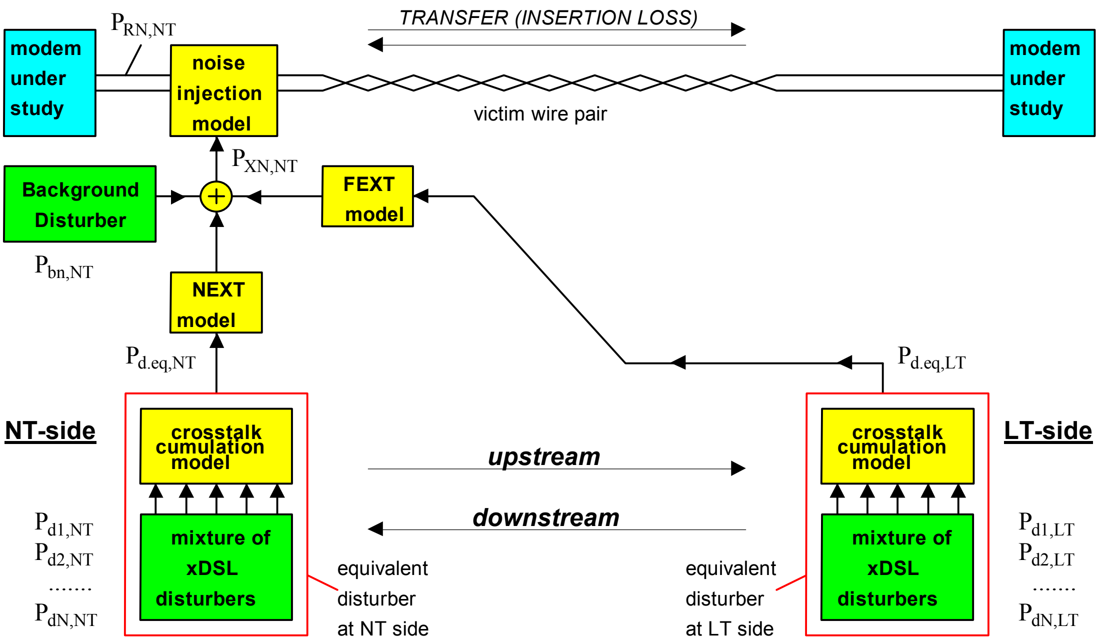


Figure 4: Flow diagram of the basic model for two-node topologies, for evaluating downstream performance

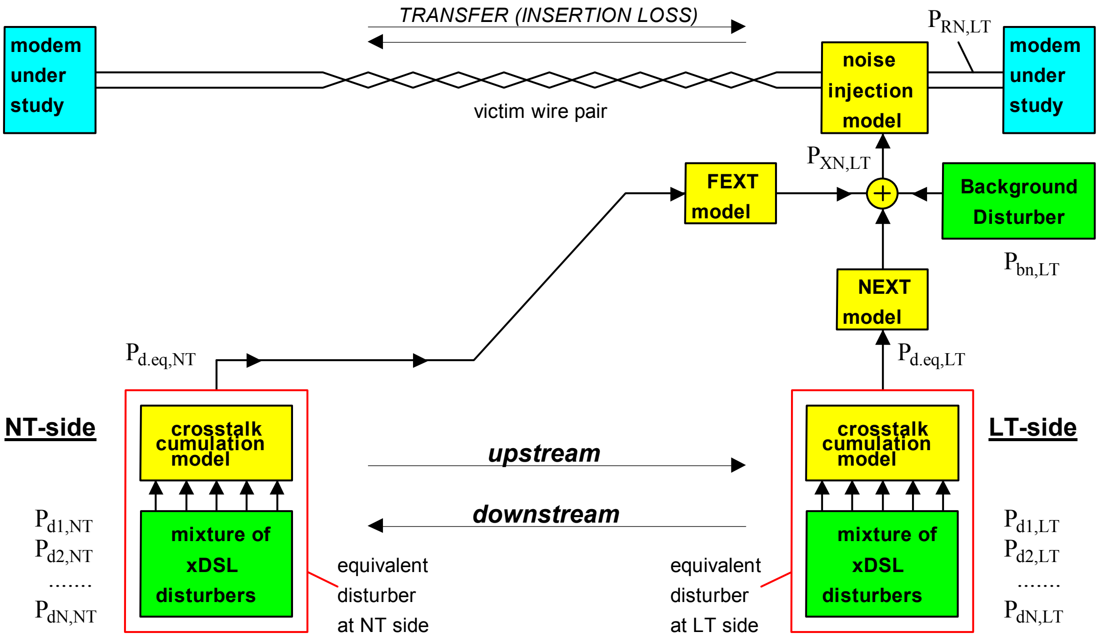


Figure 5: Flow diagram of the basic model for two-node topologies, for evaluating upstream performance

8.3.2 Models for cross talk cumulation

The noise that couples into a victim wire pair, and originates from several co-located disturbers connected to different wire pairs, cumulate in level. This cumulation cannot be modelled by a simple *linear* power sum of all individual disturbers, because each wire pair couples at different ratio to the victim wire pair. Therefore the cumulation requires some *weighted* power sum that accounts for the statistical distribution of all involved cross talk coupling ratios.
On input, the cumulation building block requires the levels ($P_{d1}\ldots P_{dN}$) of all involved individual disturbers that are co-located. On output, the cumulation building block evaluates the level of the equivalent disturbance ($P_{d.eq}$). This sub clause provides expressions to model building blocks for cross talk cumulation.

8.3.2.1 FSAN sum for cross talk cumulation

The FSAN sum is one of the possible expressions to model cross talk cumulation, and is specified in expression 14. The (frequency dependent) power level of the equivalent disturbance, which cumulates from M individual disturbers, is expressed below.

The factor K_n weights this sum when $K_n \neq 1$. For $K_n > 1$ the FSAN sum results in a power level that's is always equal or less then the linear sum (K_n) of these powers. This factor is cable dependent, and assumed to be frequency independent. Values ranging between K_n=1/0,6 and K_n = 1/0,8 have been observed in practice. On default, K_n=1/0,6 is commonly used, but this parameter must be explicitly specified when using this model for cross talk cumulation in a performance evaluation.

$$P_{d.eq} = \left(P_{d1}{}^{K_n} + P_{d2}{}^{K_n} + P_{d3}{}^{K_n} + \cdots + P_{dM}{}^{K_n}\right)^{1/K_n}$$

Expression 14: FSAN sum for cumulating the power levels of M individual disturbers into the power level of an equivalent disturber

In the special case that all M disturbers generates equal power levels (P_d) at all frequencies of interest, the FSAN sum simplifies into $P_{d.eq} = P_d \times M^{1/Kn}$.

The FSAN sum ignores differences in source impedances of different disturber types. For cumulating disturbance from sources with different impedances, their *available* power levels are to be combined according to the FSAN sum. This available power of a source is the power dissipated in a load resistance, equal to the source impedance.

8.3.3 Models for cross talk coupling

The spread in cross talk coupling between wire pairs in a real twisted pair cable is significant, and the coupling fluctuates rapidly when the frequency increases. The cross talk from a single disturber is therefore random in nature.

When the number of co-located disturbers increases, the fluctuations reduce significantly. Models for cross talk coupling take advantage of this effect and their simplicity increases when the number of co-located disturbers increases.

Equivalent cross talk coupling of a cable is the ratio between the level of the cross talk in the victim wire pair and the level of an equivalent disturber evaluated by some cross talk cumulation model, while connecting as much individual disturbers as possible to the cable under study.

This cross talk sum will be different for each wire pair, due to the random nature of the coupling. Commonly accepted models for equivalent cross talk coupling represent 99% of the victim wire pairs. This is to approximate 100% of the cases, without being pessimistic for the very last extreme 1% case.

This sub clause provides expressions to model the building blocks for *equivalent* cross talk coupling.

8.3.3.1 Basic models for equivalent NEXT and FEXT

Expression set 15 specifies how to model the transfer functions of the equivalent NEXT and FEXT building blocks. The specification is based on the following constants, parameters and functions:

- Variable $\boldsymbol{f}$ identifies the frequency.
- Constant $\boldsymbol{f_0}$ identifies a chosen reference frequency, commonly set to f_0 = 1 MHz.
- Variable $\boldsymbol{L}$ identifies the physical length of the cable between the two nodes in meters. Constant $\boldsymbol{L_0}$ identifies a chosen reference length, commonly set to L_0 = 1 km.
- Function $\boldsymbol{s_T(f, L)}$ represents the frequency and length dependent amplitude of the transmission function of the actual test loop, normalized to a reference impedance R_n. This value equals $s_T=|s_{21}|$, where s_{21} is the transmission s-parameter of the loop normalized to R_n This R_n is commonly set to 135Ω.
- Constant $\boldsymbol{K_{xn}}$ identifies an empirically obtained number that scales the NEXT transfer function $H_{next}(f, L)$.
- Constant $\boldsymbol{K_{xf}}$ identifies an empirically obtained number that scales the FEXT transfer function $H_{fext}(f, L)$.

$$
\begin{aligned}
H_{next}(f,L) &= K_{xn} \times \left(f/f_0 \right)^{0.75} \times \sqrt{1-|s_T(f,L)|^4} \\
H_{fext}(f,L) &= K_{xf} \times \left(f/f_0 \right) \times \sqrt{L/L_0} \times |s_T(f,L)|
\end{aligned}
$$

Expression 15: Transfer functions of the basic models for NEXT and FEXT

8.3.4 Models for cross talk injection

Several sub models for various building blocks within the cross talk model ignore the fact that when the modem and cable impedance will change, the noise (and signal) observed by the receiver will change as well. For instance, when the input impedance (Z_{xdsl}) of the receiver under test decreases, the received noise level will decreases as well. To account for this effect, a cross talk injection block is included in the topology models in figure 4 and 5.

The transfer function of the cross talk injection block identified as H_{xi}, and is frequency and impedance dependent. Expression 16 illustrates how to use this transfer function for evaluating the power level P_{RN} from power level P_{XN}.

$$P_{RN} = P_{XN} \times |H_{xi}|^2$$

Expression 16: Evaluation of the receive noise level from the cross talk noise level under matched conditions, by a transfer function of the noise injector.

A transfer function that models the impact of impedance mismatch can be very complex, and therefore several simplified transfer functions are commonly used to approximate this effect. This sub clause summarizes a few of these approximations.

8.3.4.1 Forced noise injection

When cross talk is modelled by means of *forced* noise injection, then all impedance and frequency dependency of noise injection is ignored. The associated transfer function is shown in expression 17.

$$H_{xi}(f) = 1$$

Expression 17: Transfer function for forced noise injection.

8.3.4.2 Current noise injection

When cross talk is modelled by means of *current* noise injection, then it is assumed that the impedance dependency can be represented by the equivalent circuit diagram shown in figure 6. The associated transfer function is shown in expression 18.

- The *injection condition* holds when the performance is evaluated. Impedance Z_{LX} represents the cable with its terminating impedance at the other ends of the line. Z_{LX} is usually a frequency dependent and complex impedance.
- The *calibration condition* holds for the situation that noise has been evaluated. Z_{cal} should be a well-defined impedance. This can be a complex artificial impedance approximating Z_{LX}, or simply a fixed real impedance. In the special case that $Z_{cal} \equiv Z_{LX}$, the concept of "current injection" simplifies into "forced injection" as described in the previous clause.

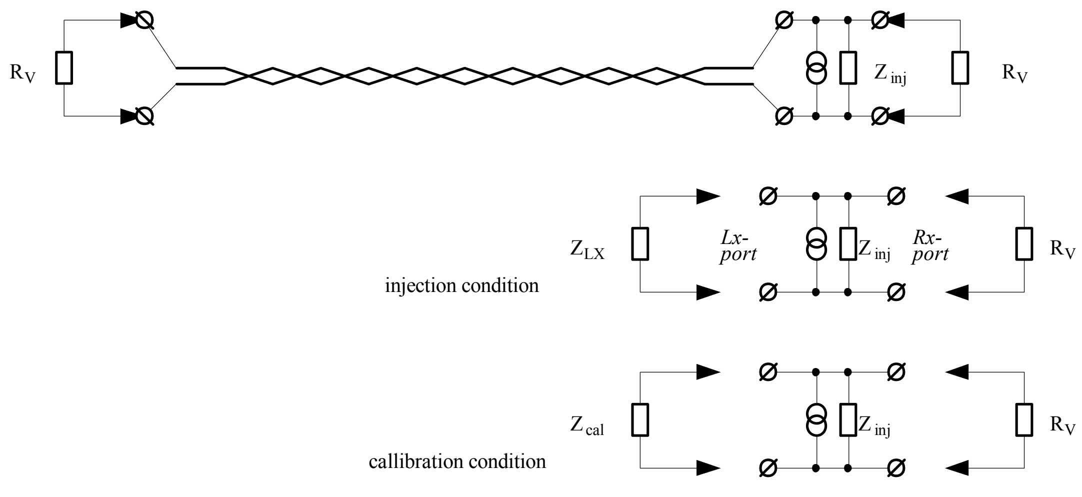


Figure 6: Current injection enables modelling of the impedance dependent behaviour of cross talk noise levels.

The transfer function $H_{xi}(f)=(U_i/U_c)$ between (a) the signal voltage U_i over impedance R_V during injection condition, and (b) U_c during calibration condition, equals:

$$H_{xi}(f, Z_{LX}) = \left(\frac{1/Z_{cal} + 1/Z_{inj} + 1/R_V}{1/Z_{LX} + 1/Z_{inj} + 1/R_V} \right)$$

Expression 18: Transfer function to model impedance dependency according to the current injection method.

8.4 Generic cross talk models for multi-node co-location

<for further study>

9 Examples of evaluating various scenarios

This chapter has left for further study, and is intended to show how the models in this document can be used to perform spectral management studies.

Annex A: Bibliography

- ETSI-TM6(97)02: "Cable reference models for simulating metallic access networks", R.F.M. van den Brink, ETSI-TM6, Permanent document TM6(97)02, revision 3, Luleå, Sweden, June 1998 (970p02r3).

History

Document history		
V0.0.0	28 January 2002	Creation of TOC and first draft.
Rev 1	15 march 2002	Textual refinement of Scope, TOC and introductory text.
Rev 2	6 June 2002	Minor refinement on definitions.
Rev 3	6 Dec 2002	Insertion of models for receiver input block (+echo loss), a 2-node cross talk model, and PSD templates for transmitter signal models HDSL.CAP/2 and SDSL.
Rev 4	28 may 2003	Insertion of generic detection models (Shifted Shannon; PAM, CAP/QAM). Correction of SDSL transmitter PSD. Rephrasing of few words and corrections of typo errors.
Rev 5	29 Aug 2003	Insertion of transmitter/disturber models for ISDN.2B1Q, and receiver performance models for HDSL.CAP and SDSL.
Rev 6	4 Feb 2004	Insertion of transmitter/disturber models for HDSL.2B1Q, and receiver performance models for the EC versions of "ADSL over POTS" and "ADSL over ISDN". Restructuring clause 5: Rephrasing the generic text of clause 5.1, by leaving out all details on out how to model echo. Move of the text of clause 7.2 to clause 5.3, because that was more appropriated, and a slightly rephrased for clarity. Removal of empty placeholders for all kinds of template PSDs for proprietary systems, since no contribution is received nor expected. These placeholders can be reinserted as soon as a proposal for the associated template PSDs is contributed to ETSI-TM6.
Rev 7	26 may 2004	Addition to the scope on the application of these computer models. Addition of definitions/terminology, identified in "part 3" to achieve consistent terminology among all 3 part. Refinement of the associated text. Correction of typo in HDSL.CAP/2 template (P=70 dBm/Hz @ f=297kHz). Insertion of PSD templates for four flavours of ADSL (EC and FDD, over POTS and over ISDN). Insertion of the generic DMT detection model.
Rev 8	6 July 2004	Insertion of text for power back-off for SDSL and ADSL transmitter models, and for the definitions. Insertion of receiver models for the FDD variants of ADSL, while leaving the values for min and maximum bit-loading explicitly for further study. Refinement of words in "definitions" and update of all "references" and "abbreviations". This version will be forwarded for AbC (Approval by Correspondence).