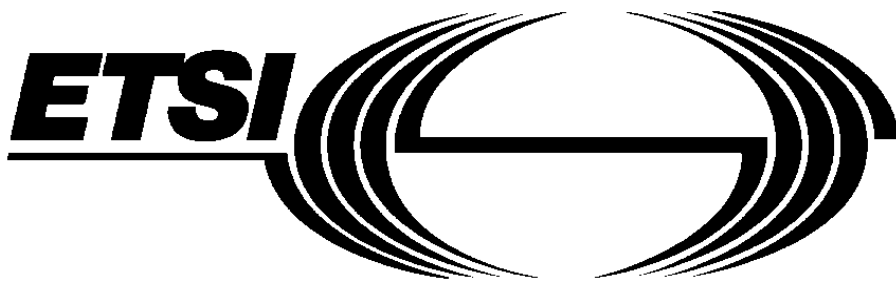# TR 101 830-2 (draft) V0.0.0 (2004-xx)

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

Rapporteur/Editor (*on behalf of KPN*)	**Rob F.M. van den Brink** TNO Telecom ~~PO Box 421~~ ~~2260 AK Leidschendam~~ PO-Box 5050 2600 GB Delft The Netherlands	tel: ~~+31 70 4462389~~ +31.15.2857059 fax: ~~+31 70 4463166~~ +31.15.2857349 email: R.F.M.vandenBrink@telecom.tno.nl **Mark the new changes, valid since nov 24, 2003**

ETSI

650 Route des Lucioles
F-06921 Sophia Antipolis Cedex - FRANCE

Tel.: +33 4 92 94 42 00 Fax: +33 4 93 65 47 16

Siret N° 348 623 562 00017 - NAF 742 C
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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); Acces 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 bitrate) 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 (choosen 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.

2   References

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

SpM

[1]   ETSI TR 101 830-1 " Transmission and Multiplexing (TM); Spectral Management on metallic access networks; Part 1: Definitions and signal library" V1.3.1 (2002-12), dec 2002.

[2]   ANSI T1E1.4/2000-002R6 "Spectrum Management for loop transmission systems" draft; revision 6, November 2000 (or a more recent version)

ISDN

[3]   ETSI TS 102 080 (V1.3.2): "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, v1.1.3: "Transmission and Multiplexing (TM); Access transmission system on metallic access cables; Symmetrical single pair high bitrate Digital Subscriber Line (SDSL)". Nov 2001.

[6]   ITU-T Recommendation G.991.2 (2001): "Single-Pair High-Speed Digital Subscriber Line (SHDSL) transceivers".

ADSL

[7]   ETSI TS 101 388, v1.3.1, (2002-05): "Transmission and Multiplexing (TM); Access transmission systems on metallic access cables; Asymmetric Digital Subscriber Line (ADSL) - European specific requirements", may 2002.

[8]   ITU-T Recommendation G.992.1 (1999): "Asymmetric digital subscriber line (ADSL) transceivers".

3   Definitions and abbreviations

3.1   Definitions

ED NOTE   When the definitions added to SpM part 3 have been agreed by ETSI-TM6, and moved from the living list to the draft of SpM-3, then these definitions will be included here as well.

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

upstream transmission: transmission direction from an NT-port to an LT-port, usually from the customer premises, via the access network, to the telecommunication exchange

downstream transmission: transmission direction from an LT-port to an NT-port, usually from the telecommunication exchange via the access network, to the customer premises

Noise margin: the ratio by which the received noise may increase until the recovered signal does not meet the predefined quality criteria. This ratio is commonly expressed in dB.

Signal margin: the ratio by which the received signal may decrease until the recovered signal does not meet the predefined quality criteria. This ratio is commonly expressed in dB.

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

Loop provider: company facilitating access to the local loop wiring. (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: company that makes use of a local loop wiring for transporting telecommunication services. This definition covers *incumbent* as well as *competitive* network operators.

Access Rule (*or metallic access rule*): Mandatory rule for achieving access to the local loop wiring, equal for all *network operators* that make 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*. A deployment rule reflects a network operators own view about what the maximum length or maximum bitrate may be for offering a specific transmission service to ensure a chosen minimum quality of service.

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
DMT	Discrete Multitone modulation
FDD	Frequency Division Duplexing/Duplexed
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
PSD	Power Spectral Density (single sided)
QAM	Quadrature Amplitude modulation
REC	Receiver
SDSL	Symmetrical (single pair high bitrate) Digital Subscriber Line
SNR	Signal to Noise Ration (*ratio of powers*)
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 linecode*)

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 mis-matched 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 (a slightly) different from the true PSD, especially at steep edges (e.g. guard bands), and for modeling 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 modeling purposes. This clause summarizes various xDSL transmitter models, by defining *template* spectra of output signals.

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 modeling 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 $q_N=1$ when no filtering is applied ($f_L \rightarrow 0$, $f_H \rightarrow \infty$).

The source impedance equals 135Ω.

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

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

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

Where:

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

$R_S = 135\ [\Omega]$

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

Default values for remaining parameters are summarized in table 1.

Expression 1: PSD template for modeling "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].

In case these default values are not appropriated for specific performance studies, other values may apply as well (provided that they are specified for these studies).

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 models for "HDSL.2B1Q"

The PSD templates for modeling the spectra of various "HDSL.2B1Q" transmitters is 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 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.

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

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

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

Where:
$P_{HDSL} = \left(10^{P_{HDSL\_dBm}/10}\right)/1000$ [W]
$R_S = 135$ [Ω]
$\text{sinc}(x) = \sin(\pi \cdot x)/(\pi \cdot x)$
Default values for remaining parameters are summarized in table 2.

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

Different HDSL implementations, may use different filter characteristics, and noise floor values. Table 2 summarizes *default* values for modeling HDSL transmitters, and *alternative* values in case higher order Butterworth filtering has been applied to dedicated implementations. It is recommended to use the default values for spectral management studies, unless motivated why alternative values are more appropriated.

The default power level P_{HDSL} equals the maximum power allowed by the HDSL standard [4], since a nominal specification does not exist. The default noise floor P_{floor} equals a value observed for various implementations. When these measurements were not available, the maximum PSD level was chosen here that meets the out-of-band specification of the HDSL standard [4].

Default

	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
	HDSL.2B1Q/1	1160	3	0.42×f_x	3	N/A	N/A	1.4662	14	-121.5
	HDSL.2B1Q/2	584	3	0.50×f_x	3	N/A	N/A	1.3501	14	-133
	HDSL.2B1Q/3	392	3	0.50×f_x	3	N/A	N/A	1.3642	14	-117

Alternatives

	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
H2.1	HDSL.2B1Q/2	584	3	0.68×f_x	4	N/A	N/A	1.1915	14	-133
H2.2	HDSL.2B1Q/2	584	3	0.68×f_x	4	1.50×f_x	2	1.1965	14	-133

Table 2: Default 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 models for "HDSL.CAP"

The PSD templates for modeling signals generated by HDSL.CAP transmitters are different for single-pair and two-pair HDSL systems. The PSD templates for modeling 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 *logarithmic* frequency scale and a *linear* dBm scale. The source impedance equals $R_s=135\Omega$.

HDSL.CAP/2	2 pair 135 Ω	HDSL.CAP/1	1- pair 135 Ω
[Hz]	[dBm/Hz]	[Hz]	[dBm/Hz]
1	-57	<TBD>	<TBD>
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	-90		
1,188 M	-120		
30 M	-120		

Table 3. PSD template values at break frequencies for modeling "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 modeling the spectra of "SDSL" transmitters is 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.

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 $R_s=135\Omega$.

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

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

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

R_s = 135 Ω
sinc(x) = sin($\pi \cdot$x) / ($\pi \cdot$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 Tempate values for modeling 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	0.5683·10^{-4}
Sym	$\geq$ 2048	both	(R+ 8 kbit/s)/3	f_{sym}	f_X/2	5	1	6	9.90	0.5683·10^{-4}
Asym	2048	LTU	(R+ 8 kbit/s)/3	2×f_{sym}	f_x×2/5	5	1	7	16.86	0.5683·10^{-4}
Asym	2048	NTU	(R+ 8 kbit/s)/3	f_{sym}	f_x×1/2	5	1	7	15.66	0.5683·10^{-4}
Asym	2304	LTU	(R+ 8 kbit/s)/3	2×f_{sym}	f_x×3/8	5	1	7	12.48	0.5683·10^{-4}
Asym	2304	NTU	(R+ 8 kbit/s)/3	f_{sym}	f_x×1/2	5	1	7	11.74	0.5683·10^{-4}

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

4.4 Cluster 4 transmitter signal models

4.4.1 Transmitter signal model for "ADSL over POTS"

<for further study>

4.4.2 Transmitter signal model for "ADSL over ISDN"

<for further study>

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

<for further study>

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

4.5 Cluster 5 transmitter signal models

4.5.1 Transmitter signal model for "VDSL"

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 modeling becomes more convenient when broken down into a cascade of smaller sub models:

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

This clause details all the above mentioned sub models, being used for evaluating the performance of receivers under noise conditions. This clause 5 is dedicated to *generic* performance models only. Clause 6 is dedicated to *specific* models by 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.
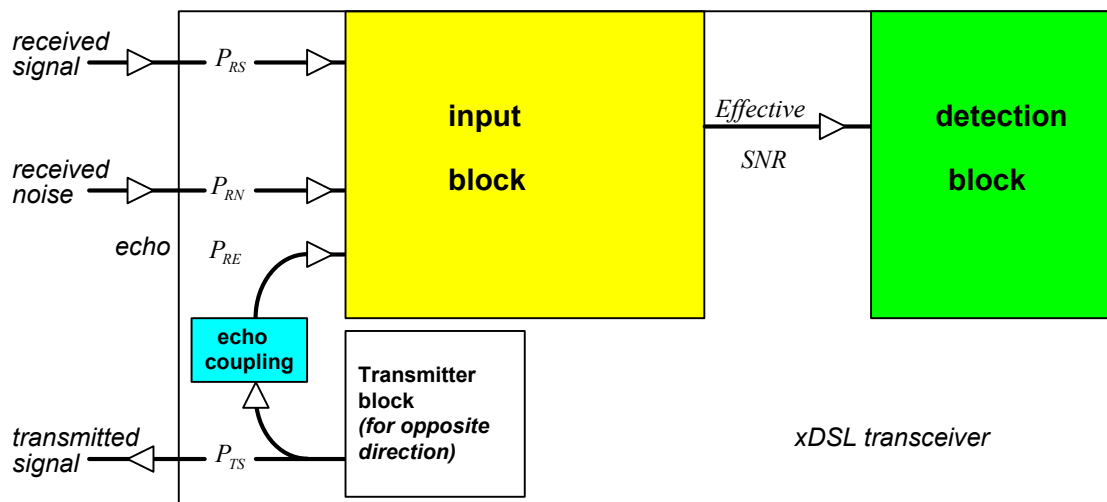

Figure 1: Flow diagram of a transceiver model, build up from individual sub models.

On input, the input model for effective SNR requires values for _signal_, _noise_ and _echo_. The flow diagram in figure 1 illustrates this for an xDSL transceiver that is connected via a common wire pair to another transceiver (not shown).

- 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 en 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.

On output, the input model 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.

Effective SNR, in offset format for margin evaluations

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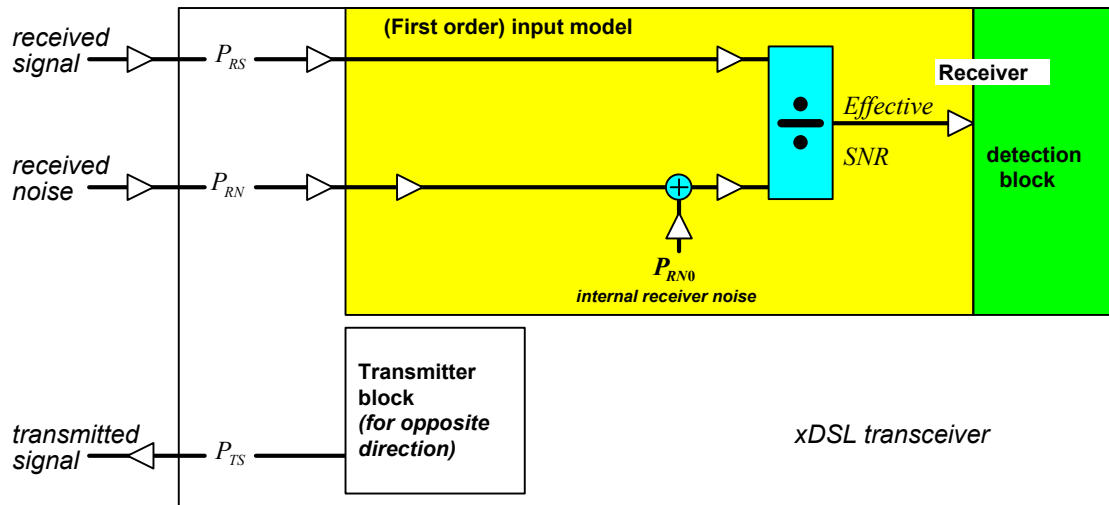
**Figure 2: Flow diagram of a transceiver model that incorporates a linear
first order input model for the determination of the effective SNR.**

Expression 4 summarizes how to evaluate the effective SNR for this model, and it has been specified in plain and offset formats. Table 5 summarizes the involved parameters.

$$\text{Plain format:} \qquad SNR(f) \qquad = \frac{P_{RS}}{P_{RN} + P_{RN0}}$$

$$\text{Noise offset format:} \quad SNR_{\text{ofs,N}}(m, f) = \frac{P_{RS}}{P_{RN} \times m + P_{RN0}}$$

$$\text{Signal offset format:} \quad SNR_{\text{ofs,S}}(m, f) = \frac{P_{RS} / m}{P_{RN} + P_{RN0}}$$

Expression 4: Effective SNR, in various formats, for a first order input model

Input quantities	linear	In dB	remarks
Received signal power	P_{RS}	$10 \times \log_{10}(P_{RS})$	Frequency dependent
Received crosstalk noise	P_{RN}	$10 \times \log_{10}(P_{RN})$	External noise
Model Parameters			
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 5: Involved parameters and quantities for a first order input model.

5.1.2   Second order input model (with residual distortion)

<for further study>

5.1.3   Second order input model (with residual echo)

5.1.4   Third order input model

<for further study; may be removed>

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 6 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 6. 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 <u>in</u>crease before the transmission becomes unreliable.

- The *Signal Margin m_s* indicates how much the received signal power can <u>de</u>crease 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 5. 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 7.

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.

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

**Expression 5: 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		= DateRate + overhead bitrate
Bandwidth	B		Width of most relevant spectrum

Table 7. Parameters used for Shifted Shannon detection models.

The various parameters used within this generic detection model are summarized in table 7. 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 linerate is usually higher then the data rate (0…30%) to transport overhead bits for error correction, signaling and framing.

- The Bandwidth is a parameter that indicates what portion of the received spectrum is relevant for data transport. The model assumes that this portion passes 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 6. This model assumes ideal decision feedback equalizer (DFE) margin calculations. The associated parameters are summarized in table 8.

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.

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

Expression 6: Equation of the PAM-detection model, for solving the margin *m*.

The SNR gap Γ, being used in the above expression 6, is a combination of various effects. This Γ parameter is often split-up into the following three parts:

- Its theoretical value 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): $\qquad\qquad \Gamma = \Gamma_{PAM} \ / \ \Gamma_{coding} \ \times \ \Gamma_{impl}$

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

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}	$10\times\log_{10}(\Gamma_{PAM})$	Modulation gap for PAM
	Γ_{coding}	$10\times\log_{10}(\Gamma_{coding})$	Coding gain
	Γ_{impl}	$10\times\log_{10}(\Gamma_{impl})$	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		= DateRate + overhead bitrate
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 8. Parameters used for PAM detection models.

The various parameters in table 8 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, signaling 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 correspond 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 7. This model assumes ideal decision feedback equalizer (DFE) margin calculations. The associated parameters are summarized in table 9.

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.

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

Expression 7: Equation of the CAP/QAM-detection model, for solving the margin *m*.

The (effective) SNR gap Γ, being used in the above expression 7, is a combination of various effects. This Γ parameter is often split-up into the following three parts:

- Its theoretical value 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		= DateRate + overhead bitrate
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 9. Parameters used for CAP/QAM detection models.

The various parameters in table 9 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, signaling 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

<left for further study>

5.3   Generic models for echo coupling

5.3.1   Linear echo coupling model

ED. NOTE This text was moved from clause 7.2, because it is more appropriated, and a slightly rephrased for clarity. The modeling of echo in general is subject of discussion within ETSI-TM6, and may be kept out completely of SpM-2 when TM6 has come to a conclusion.

This model describes a property of linear hybrids in transceivers, and models what portion of the transmitted signal reflects 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.

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

$$\frac{P_{RE}}{P_{TS}} = \left| H_E(j\omega) \right|^2$$
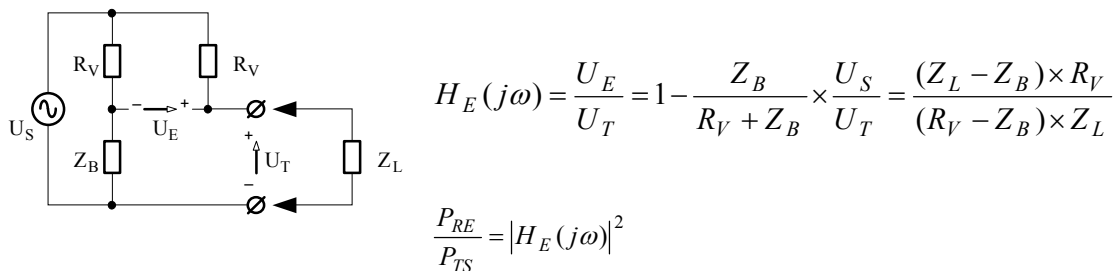
Figure 3: Flow diagram of the basic model for echo loss. The identifiers P_{RE} and P_{TS} refer to power flow values used in figure 1.

When using this basic model for echo loss 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 clause 6 defines parameter values for the generic performance models of the previous clause 5, to provide 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

Building blocks of the receiver performance model.

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 10 of the succeeding clause.

Parameters, of the receiver performance model.

The parameter values, used in the receiver performance model for ETSI compliant HDSL-CAP, are summarized in table 10. 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)	Γ_{dB}	6.8 dB	6.8 dB
SNR-Gap in parts	Γ_{CAP_dB}	<TBD>	<TBD>
	Γ_{coding_dB}	<TBD>	<TBD>
	Γ_{impl_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	N_H	+3	+3
CAP/QAM model	N_L	0	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 10. 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

Building blocks of the receiver performance model.

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.

- The generic PAM detection model, specified in clause 5.2.2

- The parameter values specified in table 11 of the succeeding clause.

Parameters, of the receiver performance model.

The parameter values, used in the receiver performance model for ETSI compliant SDSL, are summarized in table **11**. Part of them are 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}	9.75 dB	9.75 dB
	Γ_{coding_dB}	4.4 dB	5.1 dB
	Γ_{impl_dB}	1.6 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	N_H	+1	
PAM model	N_L	–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 11. 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.

6.4.1   Building blocks of the receiver performance model

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 imperfections (front-end noise, residual echo and equalization errors), in one virtual noise source.

- The generic DMT detection model, specified in clause 5.2.4.

- The parameter values specified in table 12 of the succeeding clause.

6.4.2   Parameters of the receiver performance model

The parameter values, used in the receiver performance model for ETSI compliant "ADSL over POTS" modems, are summarized in table 12. 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	64 … 640 kb/s	64 … 6144 kb/s	
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	*tones*	[7:31] = [k_1 : k_2]	[7:63 , 65:255] = [k_1 : k_2 , k_3 : k_4] Tone 64 = pilot tone	DMT tone k = 64 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 sub-carrier
Maximum bit-loading	b_{max}	15	15	Bits per sub-carrier

Table 12: Values for the performance parameters extracted from the ETSI performance requirements under ETSI stress conditions.

ED NOTE: Once the template PSDs of these ADSL variants are agreed, it is recommended to verify if the above effective SNR gap is still adequate, or needs a minor modification

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

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.

6.6.1   Building blocks of the receiver performance model

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 imperfections (front-end noise, residual echo and equalization errors), in one virtual noise source.

- The generic DMT detection model, specified in clause 5.2.4.

- The parameter values specified in table 13 of the succeeding clause.

6.6.2   Parameters of the receiver performance model

The parameter values, used in the receiver performance model for ETSI compliant "ADSL over ISDN" modems, are summarized in table 13. 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}	9.75 dB	9.75 dB	
	Γ_{coding_dB}	4.25 dB	4.25 dB	
	Γ_{impl_dB}	2.3 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	64 … 640 kb/s	64 … 6144 kb/s	
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	*tones*	[33:63] = [$k_1 : k_2$]	[33:95 , 97:255] = [$k_1 : k_2$, $k_3 : k_4$] 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 sub-carrier
Maximum bit-loading	b_{max}	15	15	Bits per sub-carrier

Table 13: Values for the performance parameters extracted from the ETSI performance requirements under ETSI stress conditions.

ED NOTE: Once the template PSDs of these ADSL variants are agreed, it is recommended to verify if the above effective SNR gap is still adequate, or needs a minor modification

6.7   Receiver performance model for "ADSL.FDD over ISDN

<left for further study>

6.8   Receiver performance model for "VDSL"

7   Transmission and reflection models

7.1   Summary of test loop models

ED NOTE   This clause refers to various testloops for ADSL, SDSL, VDSL, as defined in published documents like standards.
If required references to additional cable models can be added, but when possible we should try to keep this clause as short as possible. In practice, each country will favor its own cable models, and they are too numerous (and too proprietary) to mention them here all .

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 lineair 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 provide several cross talk models to predict how much noise is coupled into a victim wire pair.

8.1   Overview of different network topologies

<for further study>

8.2   Validity limitations of cross talk modeling

<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.

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 modeling *cross talk cumulation* as an isolated building block. This is because the cumulation from different disturbers cannot be modeled by a simple *linear* power sum of all individual disturbers. Since each wire pair couples at different ratio to the victim wire pair, the cumulation requires some *weighed* power sum that accounts for the statistical distribution of all involved cross talk coupling ratios.
  By modeling 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 modeled 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 modeled 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 8.

$$P_{XN,NT} = P_{d.eq,NT} \times |H_{next}|^2 + P_{d.eq,LT} \times |H_{fext}|^2 + P_{bn,NT}$$

$$P_{XN,LT} = P_{d.eq,LT} \times |H_{next}|^2 + P_{d.eq,NT} \times |H_{fext}|^2 + P_{bn,LT}$$

Expression 8: 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.
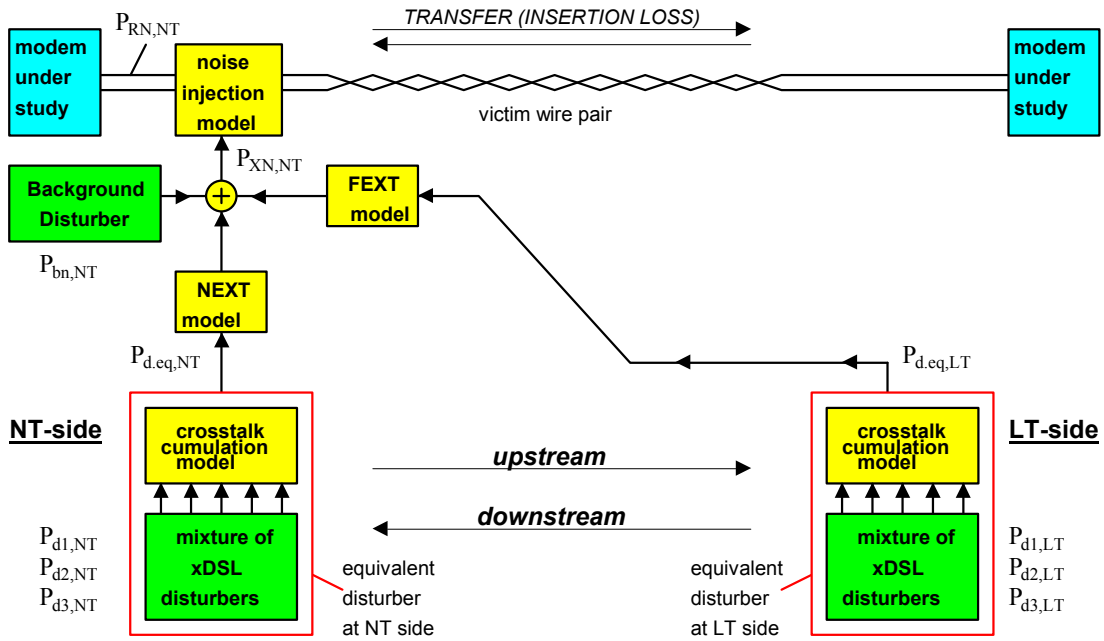
**Figure 4: Flow diagram of the basic model for two-node topologies,
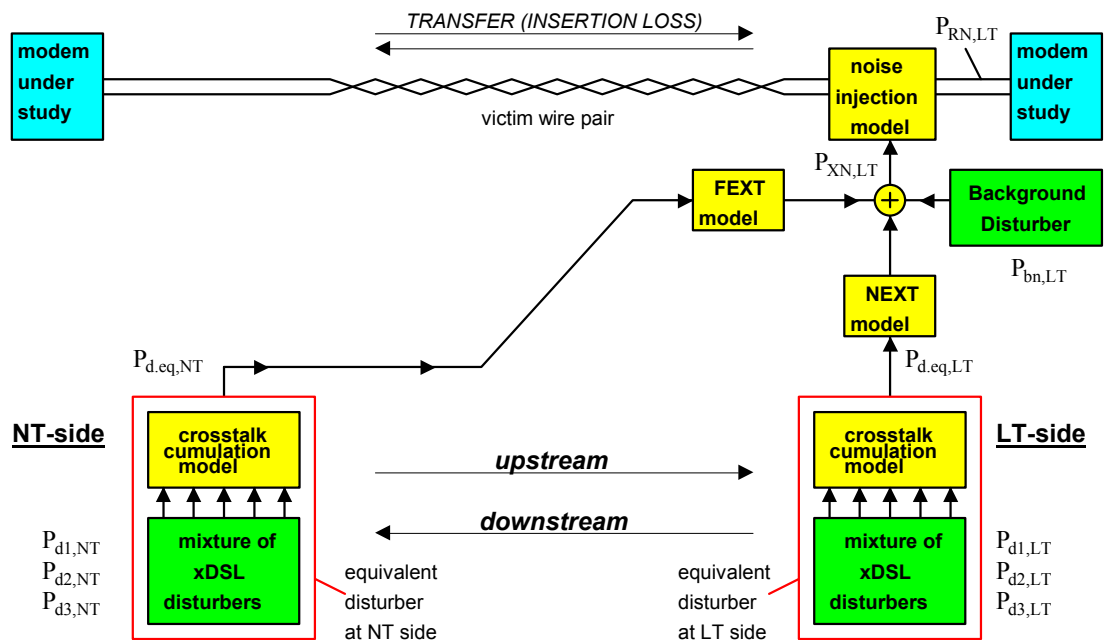for evaluating downstream performance**

**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 modeled 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 *weighed* 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 9. The (frequency dependent) power level of the equivalent disturbance, that cumulates from *M* individual disturbers, is expressed below.

The factor K_n weighs 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 9: 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 $P_{d.eq} = P_d \times M^{1/K_n}$.

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 10 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 **f** identifies the frequency.

- Constant $\mathbf{f_0}$ identifies a chosen reference frequency, commonly set to f_0 = 1 MHz.

- Variable **L** identifies the physical length of the cable between the two nodes in meters. Constant $\mathbf{L_0}$ identifies a chosen reference length, commonly set to L_0 = 1 km.

- Function $\mathbf{s_T(f, L)}$ 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 $\mathbf{K_{xn}}$ identifies an empirically-obtained number that scales the NEXT transfer function $H_{next}(f, L)$.

- Constant $\mathbf{K_{xn}}$ identifies an empirically-obtained number that scales the FEXT transfer function $H_{fext}(f, L)$.

$$H_{next}(f,L) \;=\; K_{xn} \times \left(f \middle/ f_0 \right)^{0.75} \times \sqrt{1 - \left| s_T(f,L) \right|^4}$$

$$H_{fext}(f,L) \;=\; K_{xf} \times \left(f \middle/ f_0 \right) \times \sqrt{L/L_0} \times \left| s_T(f,L) \right|$$

Expression 10: 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 11 illustrates how to use this transfer function for evaluating the power level P_{RN} from power level P_{XN}.

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

Expression 11: 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 summarize 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 12.

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

Expression 12: 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 13.

- 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.
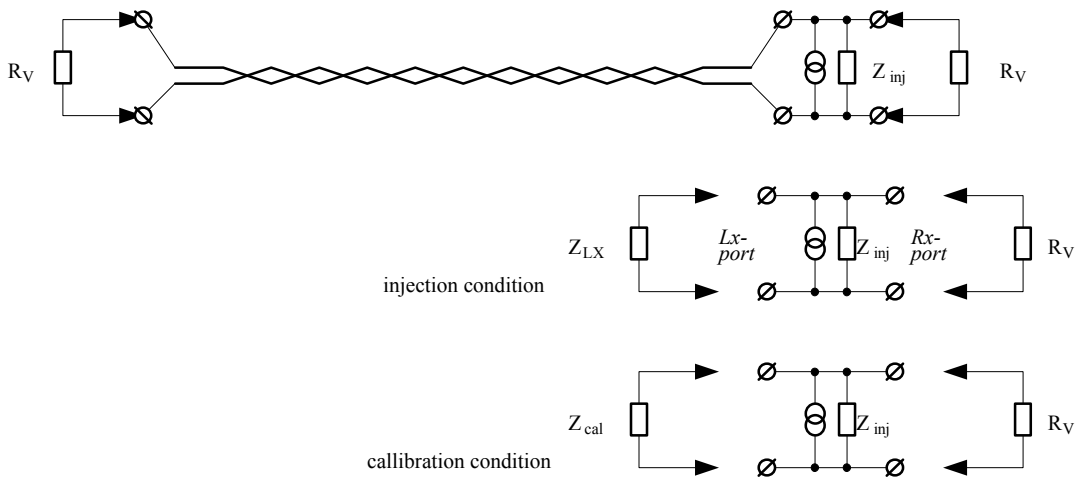

Figure 6: Current injection enables modeling of the impedance dependent behavior of cross talk noise levels.

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:

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

Expression 13: Transfer function to model impedance dependency according to the current injection method.

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

ED NOTE This clause provides the common calculation approach for deploying xDSL from subloop location (like HDSL repeaters and VDSL). For these calculations, the access network is simplified as if is a single cable but with multiple LT and NT-nodes distributed along the cable.

9   Measurement methods

ED NOTE This clause has been included here on explicit request, as a placeholder for using measurements instead of calculations. Currently, there is no detailed guidance for this approach, so this will be contribution driven.

10   Examples of evaluating various scenarios

ED NOTE This section should demonstrate how to define a full scenario in less that one page of paper, be referring as much as possible to the described reference models

These scenario's are examples only, and enable for each scenario to calculate the performance of each involved system. If, for a specific purpose, one of these scenarios is labeled as "reference" and another one as "modified" then the change in performance is a nice demonstration of what the consequences are of changing for instance the technology mix. This can be a basis in what context (= specific scenario) the word "spectral compatibility" has got a meaning.

10.1   Example scenario A

ED NOTE (this example is FSAN noise model B for ADSL)

10.1.1   Assumed configuration

Disturber assumptions

Technology mix	Number of wire pairs	Transmitters/disturbers model
ISDN.2B1Q	10	ETSI default model "ISDN.2B1Q"
HDSL.2B1Q (2-pair)	2×2	ETSI default model "HDSL.2B1Q/2"
ADSL over ISDN (E.C.)	15	ETSI default model "ADSL over ISDN"
SDSL (2.3 Mb/s; sym)	15	ETSI default model "SDSL"

Performance assumptions

Technology	Target noise margin	Performance model
ISDN.2B1Q	6 dB	ETSI default model "ISDN.2B1Q"
HDSL.2B1Q (2-pair)	6 dB	ETSI default model "HDSL.2B1Q/2"
ADSL over ISDN (E.C.)	6 dB	ETSI default model "ADSL over ISDN"
SDSL (2.3 Mb/s; sym)	6 dB	ETSI default model "SDSL"

10.1.2   Assumed conditions

property	Model name	Parameter values
Transmision models	ETSI testloop model "ADSL#2"	-
	ETSI default echo-loss model	Rv=135 (HDSL/SDSL/ISDN) Rv=100 (ADSL)
Cross talk models	Basic two-node topology model	-
	FSAN cumulation model	K_n=0.6
	Basic NEXT & FEXT model	K_{xn}=–50 dB @ 1 MHz K_{xn}=–45 dB @ 1 MHz, 1 km
	Current injection model (real)	Z_{line}= 135 ohm Rv=135 (HDSL/SDSL/ISDN) Rv=100 (ADSL)

10.1.3   Evaluated performance for scenario A

ED NOTE:
Margin of technology "HDSL.2B1Q" as a function of cable length
Margin (or bitrate) of technology "ADSL over ISDN" as a function of cable length
Margin (or bitrate) of technology "SDSL" as a function of cable length

10.2   Example scenario B

<for further study>

10.3   Example scenario C

<for further study>

10.4   Example scenario D

<for further study>

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 crosstalk 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 contrtibution is received nor expected. These placeholders can be reinserted as soon as a proposal for the associated template PSDs is contributed to ETSI-TM6.