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*Technical Report*

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

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# Contents

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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" V1.2.1 (2001-08), august 2001.

[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: "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

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 specificing 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"

<for further study>

### 4.2.2 transmitter signal model for "ISDN.MMS.43"

<for further study>

### 4.2.3 transmitter signal model for "Proprietary.SymDSL.CAP.QAM"

<for further study>

## 4.3 Cluster 3 transmitter signal models

### 4.3.1 Transmitter signal models for "HDSL.2B1Q"

<for further study>

### 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 1. 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 $R_s$=135Ω.

| **HDSL.CAP/2** | 2 pair<br>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 | -90 |
| 1,188 M | -120 |
| 30 M | -120 |

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

**Table 1. 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 1 and the associated table 2.

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.

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

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

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

$R_s$ = 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 2*

**Expression 1. 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 | $0.5683 \cdot 10^{-4}$ |
| Sym | ≥ 2048 | both | (R+ 8 kbit/s)/3 | $f_{sym}$ | $f_X/2$ | 5 | 1 | 6 | 9.90 | $0.5683 \cdot 10^{-4}$ |
| Asym | 2048 | LTU | (R+ 8 kbit/s)/3 | $2 \times f_{sym}$ | $f_x \times 2/5$ | 5 | 1 | 7 | 16.86 | $0.5683 \cdot 10^{-4}$ |
| Asym | 2048 | NTU | (R+ 8 kbit/s)/3 | $f_{sym}$ | $f_x \times 1/2$ | 5 | 1 | 7 | 15.66 | $0.5683 \cdot 10^{-4}$ |
| Asym | 2304 | LTU | (R+ 8 kbit/s)/3 | $2 \times f_{sym}$ | $f_x \times 3/8$ | 5 | 1 | 7 | 12.48 | $0.5683 \cdot 10^{-4}$ |
| Asym | 2304 | NTU | (R+ 8 kbit/s)/3 | $f_{sym}$ | $f_x \times 1/2$ | 5 | 1 | 7 | 11.74 | $0.5683 \cdot 10^{-4}$ |

**Table 2. Parameter values for the SDSL templates, as defined in expression 1.**

### 4.3.4 Transmitter signal model for "Proprietary.SymDSL.CAP.A::Fn"

<for further study>

### 4.3.5 Transmitter signal model for "Proprietary.SymDSL.CAP.B::Fn"

<for further study>

### 4.3.6 Transmitter signal model for "Proprietary.SymDSL.CAP.C::Fn"

<for further study>

### 4.3.7 Transmitter signal model for "Proprietary.SymDSL.PAM::Fn"

<for further study>

### 4.3.8 Transmitter signal model for "Proprietary.SymDSL.2B1Q::Fn"

<for further study>

### 4.3.9 Transmitter signal model for "Proprietary.PCM.HDB3.2M.SR"

<for further study>

### 4.3.10 Transmitter signal model for "Proprietary.PCM.HDB3.2M.SQ"

## 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"

<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 datarate. When the internal receiver noise is zero and the echo cancellation is infinite, quantities like noise margin and signal margin become equal.

Performance models are implementation and linecode specific. Performance modeling becomes more convenient when broken down into a cascade of smaller submodels:

- a linecode independent *input* (sub)model that evaluates the effective SNR from received signal, received noise, and various receiver imperfections.

- a linecode dependent *detection* (sub)model that evaluates the performance (e.g. the noise margin at specified bit rate) from the effective SNR.

This clause describes various sub models, being used for evaluating the performance of receivers under noise conditions.

NOTE Generic models are defined with various parameters, to express various receiver properties. They include parameters to express the amount of echo suppression, receiver noise level, and SNR gap. 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

This clause describes (sub)models for xDSL performance that enable the description of the line code independent behavior of xDSL receivers. They describe how to evaluate the effective SNR, as intermediate result, from various input quantities and linear imperfections. When combined with a (sub)model of a line code dependent detection block a complete performance model can be formed (see succeeding subclauses).

### 5.1.1 Linear input model for effective SNR

This model is restricted to linear evaluations of the effective SNR. When non-linear behavior of the input block is relevant, such as for gain controlled analog frontends, more advanced input models may be required.

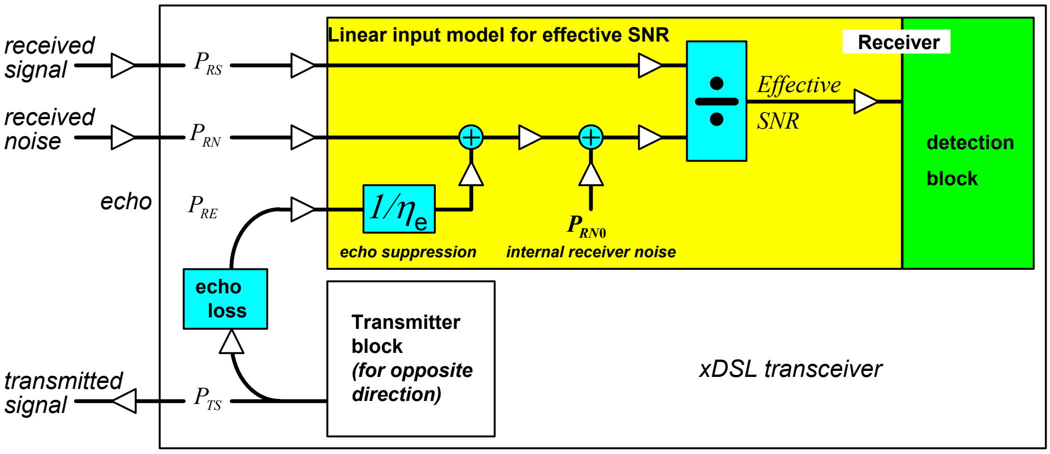



**Figure 1: Flow diagram of a transceiver model, that incorporates a linear model for effective SNR.**

*On input*, the linear 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 which is identified as echo. The echo loss building block models this effect.

The transfer function of echo loss can be modeled by one of the models described in 7.2 (see expression 6 ), and is related to the cable characteristics and the transceiver termination impedances on both ends of the cable.

*On output*, the linear input model for effective SNR evaluates a quantity called SNR (Signal to noise Ratio) that indicates to what degree the received signal is deteriorated by noise and residual echo. Due to signal processing by the receiver the *input* SNR (the ratio between signal power, and the powersum 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, while the A/D-converter produces quantization noise. The combination of all these individual noise sources deteriorates the effective SNR.

The flow diagram of figure 1 illustrates how this effective SNR is evaluated by this model of the input block. It incorporates two parameters: (a) an *echo suppression factor* $\eta_e$ that indicates how effective echo cancellation is implemented, and (b) an equivalent *receiver noise power* $P_{RN0}$ that indicates how much noise is added by the receiver electronics. This input model evaluates the effective SNR as follows:

$$SNR(P_{RS}, P_{RN}, P_{RE}, P_{RN0}, \eta_e) = \frac{P_{RS}}{P_{RN} + P_{RN0} + P_{RE} / \eta_e^2}$$

In principle all parameters of the effective SNR can be assumed as frequency dependent, but this dependency has been omitted here. In addition, external change of signal and noise levels will modify the value of this effective SNR.

To simplify further analysis of performance quantities like noise margin and signal margin, a short-cut is used for the effective SNR by applying dedicated offset formats. The simplified SNR formula is now parameterized by a single offset parameter $m$ and an optional frequency parameter $f$. The offset effective SNR is the effective SNR, evaluated when the received signal or the received noise power has been modified by a factor $m$. 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. Two offset formats for this SNR are identified in expression 2.

| | |
|---|---|
| ***Noise* offset format:** | $SNR_{ofs,N}(m,f) = SNR(P_{RS}(f), P_{RN}(f) \times m, P_{RE}(f), P_{RN0}(f), \eta_e(f))$ |
| ***Signal* offset format:** | $SNR_{ofs,S}(m,f) = SNR(P_{RS}(f)/m, P_{RN}(f), P_{RE}(f), P_{RN0}(f), \eta_e(f))$ |

**Expression 2: Shortcuts for SNR, resulting from the linear input model, using offset formats.**

These shortcuts are used for modeling the detection block of a receiver. Mark that when the receiver noise becomes zero and the echo suppression infinite, the noise offset and signal offset formats become the same.

## 5.1.2 Advanced input models for effective SNR

<left for further study>

ED NOTE These input models may address imperfections that cannot be represented by simple linear modelling. For example the non-linear aspects of gain controlled analog frontends

## 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 3 summarizes the naming convention for input and output quantities.

| Input quantities | linear | In dB | remarks |
|---|---|---|---|
| Signal to Noise Ratio | SNR | $10\times\log_{10}(\text{SNR})$ | 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 3. Symbols used for input and output quantities of detection models**

<u>On input</u>, 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 (see expression 2), it will also be a function of the offset parameter *m*.

<u>On output</u>, 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 3. 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 4.

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 expression 2), 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 3: Equation of the Shifted Shannon detection model, for solving the margin *m*.**

| Model Parameters | linear | In dB | remarks |
|---|---|---|---|
| SNR gap | $\Gamma$ | $10\times\log_{10}(\Gamma)$ | |
| 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 4. Parameters used for Shifted Shannon detection models.**

The various parameters used within this generic detection model are summarized in table 4. The model can be made specific by assigning values to all these model parameters.

- The SNR-gap ($\Gamma$) 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 4. This model assumes ideal decision feedback equalizer (DFE) margin calculations. The associated parameters are summarized in table 5.

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 expression 2), 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 4: Equation of the PAM-detection model, for solving the margin *m*.**

The SNR gap $\Gamma$, being used in the above expression 4, is a combination of various effects. This $\Gamma$ parameter is often split-up into the following three parts:

- Its theoretical value $\Gamma_{PAM}$ (in the order of 9.75 dB, at BER=$10^{-7}$)
- A theoretical coding gain $\Delta\Gamma_{coding}$ (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 $\Delta\Gamma_{impl}$ (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 $\Gamma$ is split-up into the above three parts, its value shall be evaluated as follows:

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

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

| Model Parameters | linear | In dB | remarks |
|---|---|---|---|
| SNR gap (effective) | $\Gamma$ | $10\times\log_{10}(\Gamma)$ | $= SNR_{req} / (2^{2\cdot b}-1)$ |
| SNR gap in parts: | $\Gamma_{PAM}$<br>$\Delta\Gamma_{coding}$<br>$\Delta\Gamma_{impl}$ | $10\times\log_{10}(\Gamma_{PAM})$<br>$10\times\log_{10}(\Delta\Gamma_{coding})$<br>$10\times\log_{10}(\Delta\Gamma_{impl})$ | Theoretical linecode value<br>Coding gain<br>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 5. Parameters used for PAM detection models.**

The various parameters in table 5 used within this generic detection model have the following meaning:

- The SNR-gap ($\Gamma$) and required SNR ($SNR_{req}$) are equivalent parameters and can be converted from one to the other. The advantage of using $\Gamma$ over $SNR_{req}$ is that $\Gamma$ 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 $\Gamma$ 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 5. This model assumes ideal decision feedback equalizer (DFE) margin calculations. The associated parameters are summarized in table 6.

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 expression 2), 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 5: Equation of the CAP/QAM-detection model, for solving the margin *m*.**

The (effective) SNR gap $\Gamma$, being used in the above expression 5, is a combination of various effects. This $\Gamma$ parameter is often split-up into the following three parts:

- Its theoretical value $\Gamma_{CAP}$ (in the order of 9.8 dB for BER=$10^{-7}$)
- A theoretical coding gain $\Delta\Gamma_{coding}$ (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 $\Delta\Gamma_{impl}$ (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 $\Gamma$ is split-up into the above three parts, its value shall be evaluated as follows:

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

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

| Model Parameters | linear | In dB | remarks |
|---|---|---|---|
| SNR gap (effective) | $\Gamma$ | $10\times\log_{10}(\Gamma)$ | $= SNR_{req} / (2^b-1)$ |
| SNR gap in parts: | $\Gamma_{CAP}$<br>$\Delta\Gamma_{coding}$<br>$\Delta\Gamma_{impl}$ | $10\times\log_{10}(\Gamma_{PAM})$<br>$10\times\log_{10}(\Delta\Gamma_{coding})$<br>$10\times\log_{10}(\Delta\Gamma_{impl})$ | Theoretical linecode value<br>Coding gain<br>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 6. Parameters used for CAP/QAM detection models.**

The various parameters in table 6 used within this generic detection model have the following meaning:

- The SNR-gap ($\Gamma$) and required SNR ($SNR_{req}$) are equivalent parameters and can be converted from one to the other. The advantage of using $\Gamma$ over $SNR_{req}$ is that $\Gamma$ 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 $\Gamma$ 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>

## 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.

> ED NOTE This will be the main portion of the document. The validity of each model that get the predicate "ETSI compliant" must be demonstrated by showing how close it can predict the ETSI performance requirements specified in the associated ETSI xDSL standard.
> For instance SDSL: Gap=6.6 dB, Echo=-50dB, Noise=-110 dBm, BitDensity=3 bits/symbol, Overhead= …, etc.

## 6.1 Receiver performance model for "HDSL.2B1Q"

## 6.2 Receiver performance model for "HDSL.CAP"

## 6.3 Receiver performance model for "SDSL"

## 6.4 Receiver performance model for "ADSL over POTS"

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

## 6.6 Receiver performance model for "ADSL over ISDN"

## 6.7 Receiver performance model for "ADSL.FDD over ISDN

## 6.8 Receiver performance model for "VDSL"

# 7 Transmission and reflection models

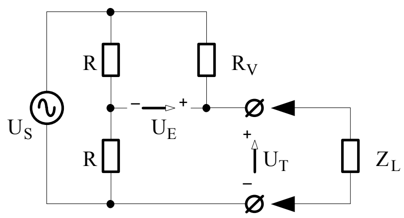
## 7.1 Summary of test loop models

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 all here.

## 7.2 Basic model for echo loss

A model for echo loss describes a property of the hybrid in a transceiver, and models what portion of the transmitted signal reflects directly into the receiver. When the hybrid is perfectly balanced, no echo will flow into the receiver. When the cable impedance differs from the value where the hybrid is designed for, the hybrid will be out of balance and some transmitted signal reflects into the receiver.

The basic model for echo loss assumes that (a) the output impedance of the transceiver equals some value $R_v$, that (b) the hybrid is balanced when terminated with a load impedance $Z_L$ equal to $R_v$, and that the hybrid can be represented by a Wheatstone bridge. This is illustrated in figure 2. The associated transfer function $H_E$ is specified in expression 6.




$$\frac{U_E}{U_T} = \frac{U_T - U_s/2}{U_T} = 1 - \tfrac{1}{2} \times \frac{R_V + Z_L}{Z_L} = \frac{Z_L - R_V}{2 \cdot Z_L}$$

**Figure 2: Flow diagram of the basic model for echo loss**

| | |
|---|---|
| $H_E(j\omega) = \dfrac{Z_L(j\omega) - R_V}{2 \cdot Z_L(j\omega)}$ | $\dfrac{P_{RE}}{P_{TS}} = \lvert H_E(j\omega) \rvert^2$ |

**Expression 6: Transfer function 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 $Z_L$ 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 $R_V$.

# 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 3 and 4. 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 3 and 4 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 3 and 4 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 3 and 4.

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 7.

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

**Expression 7: Evaluation of the cross talk power levels, that flow into the noise injection blocks of the two-node topology models in figure 3 and 4.**

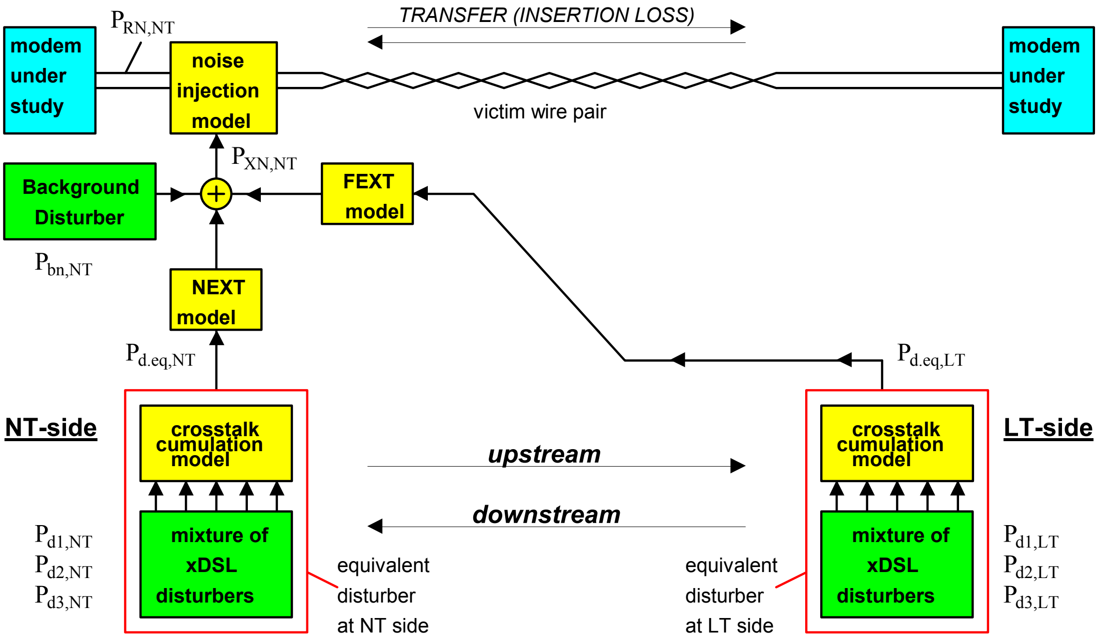



**Figure 3: Flow diagram of the basic model for two-node topologies, for evaluating downstream performance**

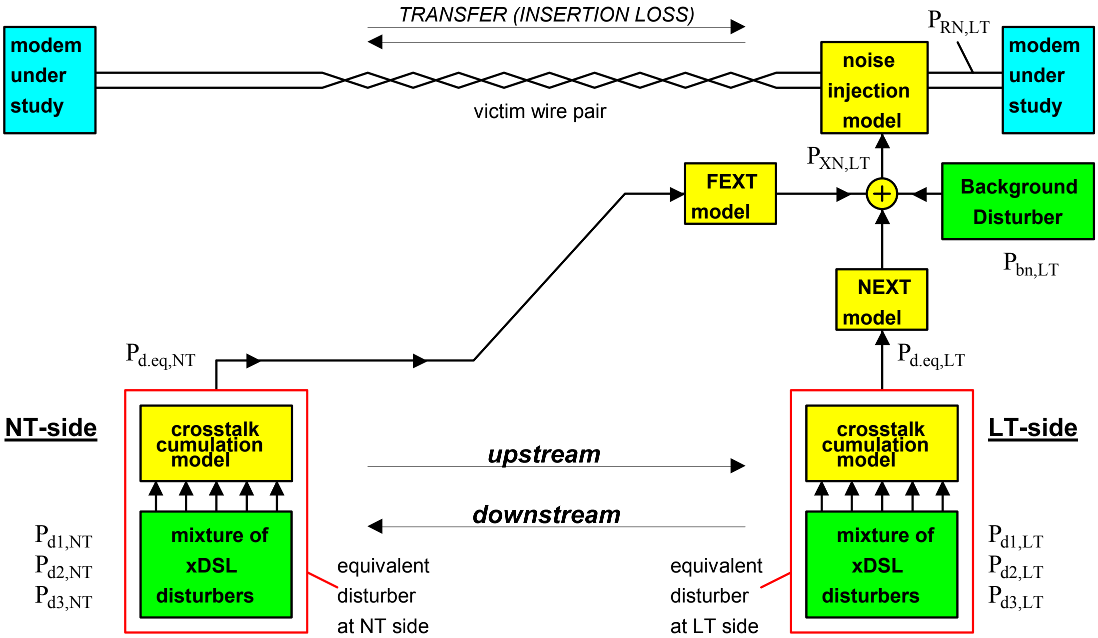



**Figure 4: Flow diagram of the basic model for two-node topologies, for evaluating upstream performance**

## 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 8. 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 8: 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 9 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 $H_{fext}(f, L)$.

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

**Expression 9: 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 3 and 4.

The transfer function of the cross talk injection block identified as $H_{xi}$, and is frequency and impedance dependent. Expression 10 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 10: 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 11.

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

**Expression 11: 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 5. The associated transfer function is shown in expression 12.

- 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.

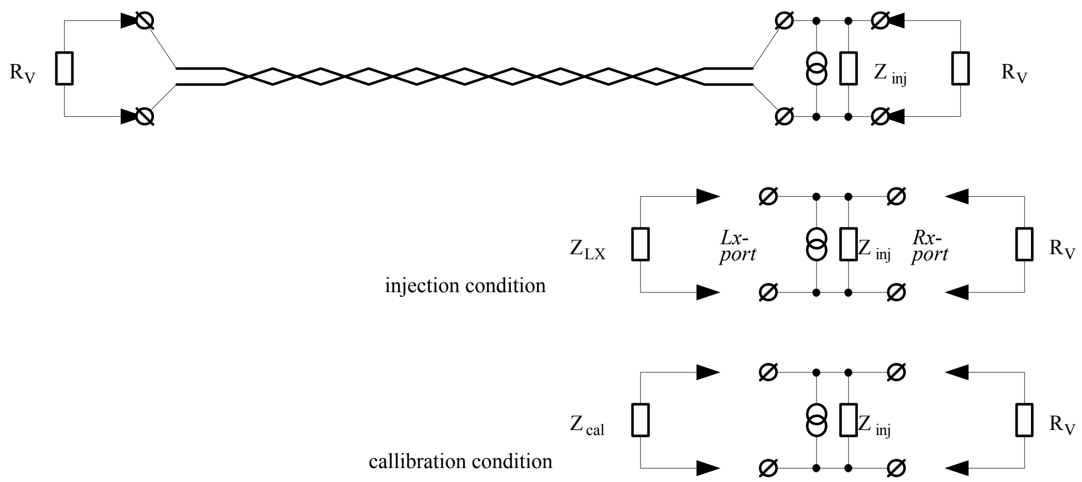



**Figure 5: Current injection enables modeling of the impedance dependent behavior of cross talk noise levels.**

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:

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

**Expression 12: 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.

<for further study>

# 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)<br>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<br>$K_{xn}$=–45 dB @ 1 MHz, 1 km |
| | Current injection model (real) | $Z_{line}$= 135 ohm<br>Rv=135 (HDSL/SDSL/ISDN)<br>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 |
| | 15 march 2002 | Textual refinement of Scope, TOC and introductory text |
| | 6 june 2002 | Minor refinement on definitions |
| | 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 |
| | 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. |