ETSI WG TM6
(ACCESS TRANSMISSION SYSTEMS ON METALLIC CABLES)

Permanent Document

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Living List for Spectral Management SpM - part 2 creation of TR 101 830-2

This document is the living list of current issues connected with ETSI's spectral management report TR 101 830, part 2 (*Technical methods for performance evaluations*).
This work item is focussed on the creation of "Part 2", dedicated to calculation and measurement methods for evaluating what the performance of xDSL systems will be for various scenarios.
The target is to achieve working group approval by the end of the ETSI-TM6 meeting in march 2003. This means that the first version of SpM part 2 can be published by ETSI before summer 2003. Issues that are (still) unsolved by that time, may be scheduled for a succeeding revision.
The issues related to the revision of "Part 1" or the creation of "Part 3" are beyond the scope of this living list.

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mark the above changes, since feb 2001

2. STUDY POINTS PART 2 (TECHNICAL METHODS FOR PERFORMANCE EVALUATIONS)

SP	Title	Owner	Status
2-1	~~Spectral management aspects of non-stationary signals.~~	~~Reuven Franco (Tioga)~~	Deleted
2-2	Basic model of input block	Ragnar Jonsson (Conexant)	Prov Agreed
2-3	Basic model of 2-node cross talk	Rob van den Brink (KPN)	Prov Agreed
2-4	Generic detection models	Rob van den Brink (KPN)	Under Study
2-5	Transmitter/Disturber models	Rosaria Persico (TI-labs)	Under Study
2-6			
2-7			
2-8			
2-9			
2-10			

The current agreed procedure for changing the status of living list items is in Annex A of TM6 working methods.

Part 2 study points

SP 2-1. Spectral management rules for non-stationary signals.

It was observed that the combined impairment from modems that are rapidly switching on and off over a period of time is much more destructive to ADSL then when these modems are continuously transmitting their signals. This is identified as "non stationary noise". The effect of non-stationary transmission in general on ADSL modems has not been fully understood. Is it a performance issue, related to the way a victim xDSL modem is implemented, or is it a spectral management issue that requires a way to bound the amount of non-stationary behaviour of signals that are injected into the Local Loop Wiring.

This study point is dedicated to the analysis of the impact of non-stationary cross talkers on legacy systems, and to find a way to model and bound the amount of non stationary noise.

Status: Deleted

Related Contributions:

- *TD25, TD26,TD35,TD53, Montreux 2000 - Alcatel*
- *TD24, Helsinki 2000, Impact of non-stationary cross talk on legacy ADSL modems - Orckit*
- *TD52, Vienna - Alcatel*
- *TD53, Vienna 2000, Stationarity requirements for spectral compatibility - Tioga*

SP 2-2. Basic model of input block.

Part 2 of SpM requires a range of calculation blocks, to enable performance evaluations. One of them is the evaluation of SNR, as interim result of an xDSL performance model (receiver). This study point explores possible improvements to the calculation blocks proposed in TD35 (021t35) of the Torino meeting, dedicated to the input block and the associated echo loss model.

Status: Provisionally Agreed

Related Contributions:

- *TD35, Torino 2002 - Model of basic input block, within xDSL receivers - KPN*

SP 2-3. Basic model of 2-node cross talk.

Part 2 of SpM requires a range of calculation blocks, to enable performance evaluations. One of them is the evaluation of cross talk noise levels in a scenario, in the special case that all disturbers are virtually co-located at no more than 2 nodes. This study point explores possible improvements to the calculation block proposed in TD36 (021t36) of the Torino meeting.

Status: Provisionally Agreed

Related Contributions:

- *TD36, Torino 2002 - Generic cross talk models for two-node co-location - KPN*

SP 2-4. Generic Detection models.
Part 2 of SpM requires a range of calculation blocks, to enable performance evaluations. One of them is the evaluation of the performance (in terms of noise margin or max bitrate) when a received signal is deteriorated by noise. Models for PAM and CAP/QAM and a linecode independent ("Shifted Shannon") model have been proposed. This study point explores possible improvements of the proposed models, and to study additional models dedicated to DMT.
Status: Under study
Related Contributions:

- *TD35, Sophia 2002 - Generic detection models for performance modelling - KPN*

SP 2-5. Transmitter/Disturber models.
Part 2 of SpM requires a range of calculation blocks, to enable performance evaluations. One of them is the evaluation of the expected signal levels of the "modem under study" as well as modems acting as disturber for the "modem under study". The PSD ***masks*** from "part 1" cover worst *case values* and are too pessimistic for this purpose and related to some resolution bandwidth. Performance modelling requires the definition of PSD ***templates*** representing *expected* values, being independent from any resolution bandwidth.
Status: Under study
Related Contributions:

- *TD36, Sophia 2002 - Transmitter models for performance evaluations - KPN*
- *TD22, Sophia 2002 - FSAN noise models are too pessimistic for SpM - Alcatel*
- *TD23, Sophia 2002 - PSD of ADSL is too pessimistic in FSAN noise models - Alcatel*

Text proposals, being candidate for inclusion into the Draft .

The text fragments below have been proposed for inclusion in the draft version of SpM part 2, but are still in the "under study" status. If agreement is achieved, they will be moved into the Draft

Text portion proposed for inclusion into clause 4

4.2 Cluster 2 transmitter models

4.2.1 Transmitter model for "ISDN.2B1Q"

The PSD template for modeling the "ISDN.2B1Q" transmit spectrum is defined in terms of break frequencies, as summarized in table 1. The associated values are constructed with straight lines between these break frequencies, when plotted against a <u>logarithmic</u> frequency scale and a <u>linear</u> dBm scale. The source impedance equals 135Ω.

ISDN 2B1Q	***135 W***
[Hz]	**[dBm/Hz]**
1	-31.8
15k	-31.8
30k	-33.5
45k	-36.6
60k	-42.2
75k	-55
85k	-55
100k	-48
114k	-48
300k	-69
301k	-79
500k	-90
1.4M	-90
3.637M	-120
30M	-120

Table 1 PSD template values at break frequencies for modeling "ISDN.2B1Q"

NOTE: This PSD template is constructed for in-band frequencies from a piece-wise approximation of a (theoretical) sync-shape of 2B1Q encoded signals. For out-of-band frequency the PSD template is guided by the PSD mask. The resulting envelope power of that PSD-template is close to the maximum power is allowed by the ISDN standard.

4.2.2 Transmitter model for "ISDN.MMS.43"

<This model is left for further study>

4.2.3 Transmitter model for "Proprietary.SymDSL.CAP.QAM"

<This model is left for further study>

4.3 Cluster 3 transmitter models

4.3.1 Transmitter model for "HDSL.2B1Q/1"

<This model is left for further study>

4.3.2 Transmitter model for "HDSL.2B1Q/2"

The PSD template for modeling the "HDSL.2B1Q/2" transmit spectrum is defined in terms of break frequencies, as summarized in table 2. 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 135Ω.

HDSL 2B1Q	***2 pair 135 W***
[Hz]	**[dBm/Hz]**
1	-40.2
100k	-40.2
200k	-41.6
300k	-44.2
400k	-49.7
500k	-61.5
570k	-80
600k	-80
650k	-72
755k	-72
2.92M	-119
30M	-119

Table 2 PSD template values at break frequencies for modeling "HDSL.2B1Q/2"

NOTE: This PSD template is constructed for in-band frequencies from a piece-wise approximation of a (theoretical) sync-shape of 2B1Q encoded signals. For out-of-band frequency the PSD template is guided by the PSD mask. The resulting envelope power of that PSD-template is close to the maximum power is allowed by the HDSL standard.

4.3.3 Transmitter model for "HDSL.2B1Q/3"

<This model is left for further study>

4.3.4 Transmitter model for "HDSL.CAP/1"

<This model is left for further study>

4.3.5 Transmitter model for "HDSL.CAP/2"

The PSD template for modeling the "HDSL.CAP/2" transmit spectrum is defined in terms of break frequencies, as summarized in table 3. 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 135Ω.

HDSL.CAP/2	***2 pair 135** W*
[Hz]	**[dBm/Hz]**
1	-57
3.98k	-57
21.5k	-43
39.02k	-40
237.58k	-40
255.10k	-43
272.62k	-60
297.00k	-90
1.188M	-120
30M	-120

Table 3. PSD template values at break frequencies for modeling "HDSL.CAP/2"

NOTE: This PSD template is taken from the nominal shape of the transmit signal spectrum, as specified in the ETSI HDSL standard (TS 101 135)

4.3.6 Transmitter model for "SDSL"

The PSD template for modeling the "SDSL." transmit spectrum is defined in three distinct frequency bands, as described in table 4. The break frequency f_{int} is the frequency where the curves for $P_1(f)$ and $P_2(f)$ intersect. The source impedance equals 135Ω.

$f<f_{int}$	$P_1(f) = \frac{K_{SDSL}}{R_s} \times \frac{2 \cdot f_0}{f_{sym}} \times \mathrm{sinc}^2(f/f_{sym}) \times \frac{1}{1+(f/f_H)^{2 \cdot N}} \times \frac{1}{1+(f_L/f)^2}$	[W/Hz]
$f_{int} \le f \le 1.5$ MHz	$P_2(f) = K_X \times (f/f_0)^{1.5}$	[W/Hz]
$f > 1.5$ MHz	$P_3(f) = -110$	[dBm/Hz]
	$R_s = 135\ \Omega$; $\mathrm{sinc}(x) = \sin(\pi \cdot x) / (\pi \cdot x)$	

Data Rate R	**SymbolRate f_{sym}**	**K_{SDSL}**	**K_X**	**N**	**f_H**	**f_L**	**f_0**
R < 2.024 Mb/s	(R+ 8 kbit/s)/3	7.86 V^2	$0.5683 \cdot 10^{-4}$ W	6	$f_{sym}/2$	5 kHz	1 Hz
R ≥ 2.024 Mb/s	(R+ 8 kbit/s)/3	9.90 V^2	$0.5683 \cdot 10^{-4}$ W	6	$f_{sym}/2$	5 kHz	1 Hz

Table 4. PSD template expressions for modelling "SDSL"

NOTE: This PSD template is taken from the nominal shape of the transmit signal spectrum, as specified in the ETSI SDSL standard (TS 101 524)

4.3.7 Transmitter model for "Proprietary.XXXXX"

<all proprietary models are left for further study>

4.4 Cluster 4 transmitter models

4.4.1 Transmitter model for "ADSL over POTS" (echo cancelled)

The PSD template for modeling the (echo cancelled) "ADSL over POTS" transmit spectrum is defined in terms of break frequencies, as summarized in table 5. The associated values are constructed with straight lines between these break frequencies, when plotted against a _logarithmic_ frequency scale and a _linear_ dBm scale. The source impedance equals 100Ω.

ADSL over POTS	*Up 100 W*
[Hz]	[dBm/Hz]
1	-97.5
3.99k	-97.5
4k	-92.5
25.875k	-38
138k	-38
307k	-90
1.221M	-90
1.630M	-110
30M	-110

ADSL over POTS	*Down 100 W*
[Hz]	[dBm/Hz]
1	-97.5
3.99k	-97.5
4k	-92.5
25.875k	-40
1.104M	-40
3.093M	-90
4.545M	-110
30M	-110

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

NOTE: This PSD template is based on a combination of the nominal PSD value for in-band frequencies, and the PSD mask for out-of-band frequencies, as specified in the ETSI ADSL standard.

4.4.2 Transmitter model for "ADSL.FDD over POTS"

The PSD template for modeling the "ADSL.FDD over POTS" transmit spectrum is defined in terms of break frequencies, as summarized in table 6. 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 100Ω.

ADSL.FDD over POTS	*Up 100 W*
[Hz]	[dBm/Hz]
1	-97.5
3.99k	-97.5
4k	-92.5
25.875k	-38
138k	-38
307k	-90
1.221M	-90
1.630M	-110
30M	-110

ADSL.FDD over POTS	*Down 100 W*
[Hz]	[dBm/Hz]
1	-97.5
3.99 k	-97.5
4 k	-92.5
80 k	-72.5
138.0 k	-44.2
138.1 k	-40
1.104 M	-40
3.093 M	-90
4.545 M	-110
30 M	-110

Table 6. PSD template values at break frequencies for modelling "ADSL.FDD over POTS"

NOTE: This PSD template is based on a combination of the nominal PSD value for in-band frequencies, and the PSD mask for out-of-band frequencies, as specified in the ETSI ADSL standard.

4.4.3 Transmitter model for "ADSL over ISDN" (echo cancelled)

The PSD template for modelling the (echo cancelled) "ADSL over ISDN" transmit spectrum is defined in terms of break frequencies, as summarized in table 7. 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 100Ω.

ADSL over ISDN	***Up* 100 *W***
[Hz]	**[dBm/Hz]**
1	-90
50k	-90
80k	-81.8
138k	-38
276k	-38
614k	-90
1.221M	-90
1.630M	-110
30M	-110

ADSL over ISDN	***Down* 100 *W***
[Hz]	**[dBm/Hz]**
1	-90
50k	-90
80k	-81.8
138k	-40
1.104M	-40
3.093M	-90
4.545M	-110
30M	-110

Table 7. PSD template values at break frequencies for modeling "ADSL over ISDN"

NOTE: This PSD template is based on a combination of the nominal PSD value for in-band frequencies, and the PSD mask for out-of-band frequencies, as specified in the ETSI ADSL standard.

4.4.4 Transmitter model for "ADSL.FDD over ISDN"

The PSD template for modelling the "ADSL.FDD over ISDN" transmit spectrum is defined in terms of break frequencies, as summarized in table 8. The associated values are constructed with straight lines between these break frequencies, when plotted against a <u>*logarithmic*</u> frequency scale and a <u>*linear*</u> dBm scale. The source impedance equals 100Ω.

ADSL.FDD over ISDN	***Up* 100 *W***
[Hz]	**[dBm/Hz]**
0.001	-90
50 k	-90
80 k	-81.8
120 k	-38
276 k	-38
614 k	-90
1.221 M	-90
1.630 M	-110
30 M	-110

ADSL.FDD over ISDN	***Down* 100 *W***
[Hz]	**[dBm/Hz]**
0.001	-90
93.1	-90
209	-62
253.99	-48.5
254	-40
1104	-40
3093	-90
4545	-110
30000	-110

Table 8. PSD template values at break frequencies for modeling "ADSL.FDD over ISDN"

NOTE: This PSD template is based on a combination of the nominal PSD value for in-band frequencies, and the PSD mask for out-of-band frequencies, as specified in the ETSI ADSL standard.

Text portion proposed for inclusion into clause 5

5.1 Basic model for the input block (for effective SNR)

This clause describes a linear (sub)model for xDSL performance that enables the description of the linecode independent behavior of an xDSL receiver. It describes how to evaluate the effective SNR, from various input quantities, as intermediate result. When combined with a (sub)model of a linecode dependent detection block a complete performance model can be formed (see succeeding subclauses).
When non-linear behavior of the input block is relevant, such as for gain controlled analog frontends, more advanced modeling may be required.

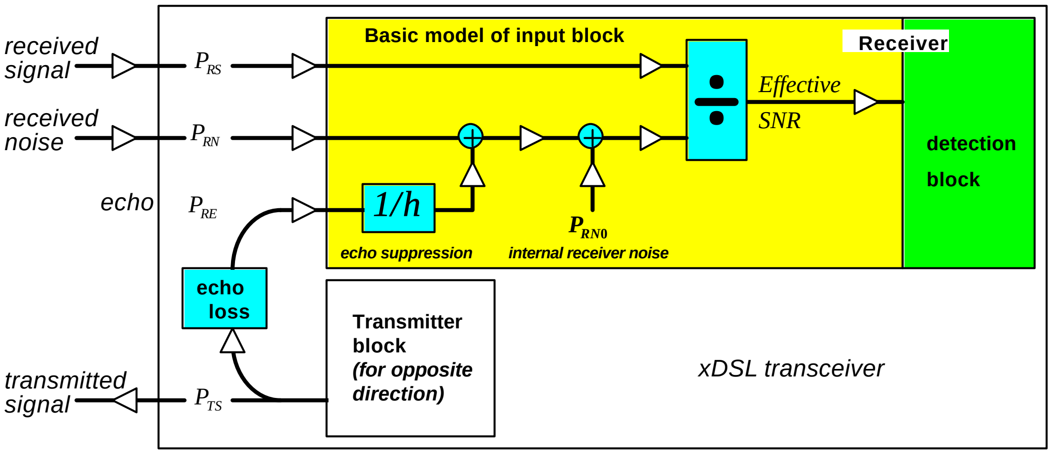


Figure 1: Flow diagram of a transceiver model, that incorporates the basic model for the input block.

On input, the basic model for the input block 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 echo loss can be modeled by the transfer function in expression 5, and is related to the cable characteristics and the transceiver termination impedances on both ends of the cable.

On output, the basic model for the input block 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 the basic model of the input block. It incorporates two parameters: (a) a *suppression factor* h 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. The basic input model evaluates the effective SNR as follows:

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

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 1.

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

Expression 1: Shortcuts for SNR, resulting from the basic model of the input block, 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.

5.2 Generic detection models

This clause identifies several generic (sub) models for the detection block: one linecode independent model derived from the Shannon capacity limit, and various linecode dependent models dedicated to PAM, CAP/QAM or DMT linecoding.
Table 9 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 9. 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 (see expression 1), it will also be a function of the offset parameter *m.*

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

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

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

NOTE From an xDSL deployment point of view, analyzing the 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 2. It has been derived from Shannon's capacity theorem, by reducing the effective SNR ("shifting" on a dB scale) by a factor Γ, to account for the imperfections of practical detectors. The associated parameters are summarized in table 10. Depending on what offset format is used for the SNR expression (see expression 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 2: 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 relevant spectrum

Table 10. Parameters used for Shifted Shannon detection models.

The various parameters used within this generic detection model are summarized in table 10. 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 datarate (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 3. The associated parameters are summarized in table 11. Depending on what offset format is used for the SNR expression (see expression 1), the calculated margin *m* will represent the noise margin m_n or the signal margin m_s.
This model assumes optimal decision feedback equalizer (DFE) margin calculations.

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

Expression 3: Equation of the PAM-detection model, for solving the margin m.

Model Parameters	*linear*	*In dB*	*remarks*
SNR gap	Γ	$10\times\log_{10}(\Gamma)$	$= SNR_{req} / (2^{2\cdot b}-1)$
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 11. Parameters used for PAM detection models.

The various parameters in table 11 used within this generic detection model have the following meaning:

- The SNR-gap (Γ) and required SNR (SNR_{req}) are similar parameters and can be converted into each 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 linerate is usually higher then the datarate (0…30%) to transport overhead bits for error correction, signaling and framing. The symbol rate is usually significantly lower when multiple bits are packed together in a single symbol.
- The summation range for n is from $n=N_L$ to $n=N_H$, 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, but 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 4. The associated parameters are summarized in table 12. Depending on what offset format is used for the SNR expression (see expression 1), the calculated margin m will represent the noise margin m_n or the signal margin m_s.
This model assumes optimal decision feedback equalizer (DFE) margin calculations.

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

Expression 4: Equation of the CAP/QAM-detection model, for solving the margin m.

Model Parameters	*linear*	*In dB*	*remarks*
SNR gap	Γ	$10\times\log_{10}(\Gamma)$	$= SNR_{req} / (2^{\cdot b}-1)$
Required SNR	SNR_{req}	$10\times\log_{10}(SNR_{req})$	$= \Gamma\times(2^{\cdot b}-1)$
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=0 and N_H=+3

Table 12. Parameters used for CAP/QAM detection models.

The various parameters in table 12 used within this generic detection model have the following meaning:

- The SNR-gap (Γ) and required SNR (SNR_{req}) are similar parameters and can be converted into each 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 linerate is usually higher then the datarate (0..30%), to transport overhead bits for error correction, signaling and framing. The symbol rate is usually significantly lower when multiple bits are packed together in a single symbol.
- The summation range for n is from $n=N_L$ to $n=N_H$, Commonly used values for CAP/QAM systems using oversampling 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).

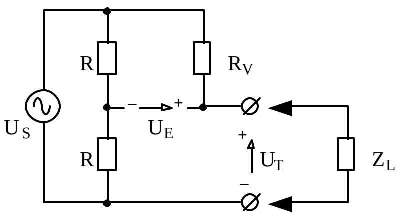
5.2.4 Generic DMT detection model

<left for further study>

Text portion proposed for inclusion into clause 7

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 5.



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

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

$H_E(jw) = \dfrac{Z_L(jw) - R_V}{2 \cdot Z_L(jw)}$	$\dfrac{P_{RE}}{P_{TS}} = \left\lvert H_E(jw) \right\rvert^2$

Expression 5: 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.

Text portion proposed for inclusion into clause 8

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. 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 6.

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

Expression 6: 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.

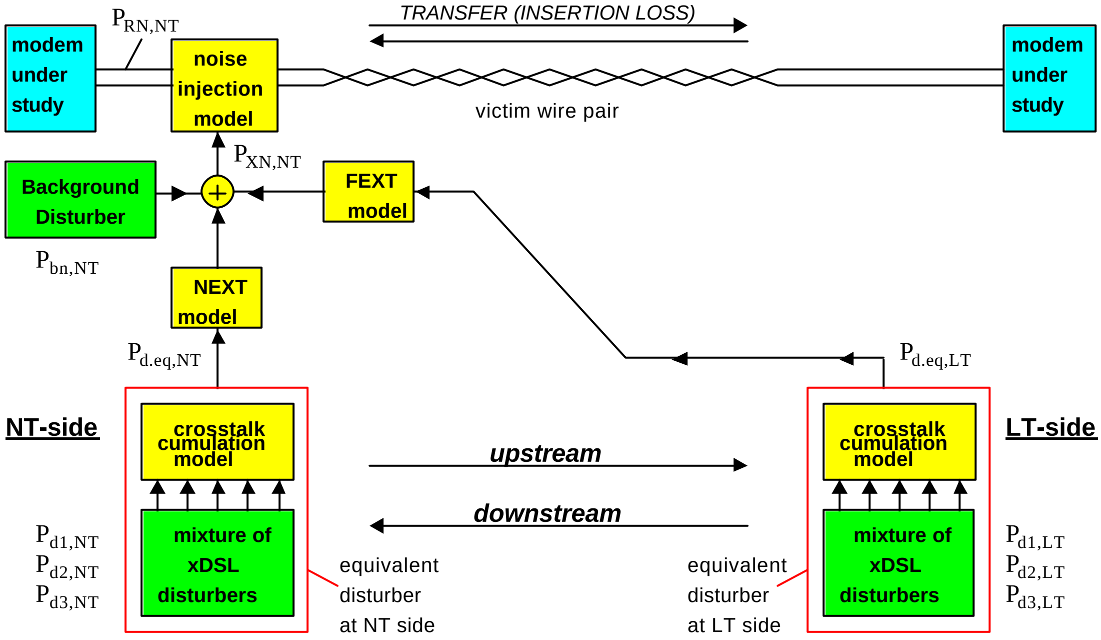


Figure 3: Flow diagram of the basic model for two-node topologies, for evaluating downstream performance

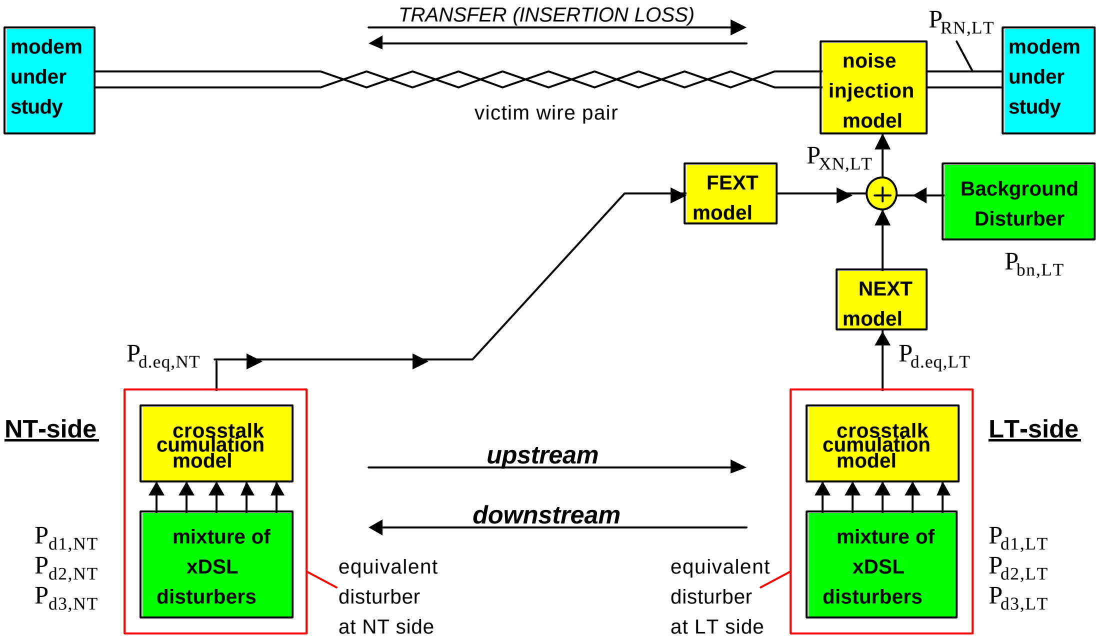


Figure 4: Flow diagram of the basic model for two-node topologies, for evaluating upstream performance

8.1.2 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.

8.1.2.1 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 7. 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 7: 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), 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.

8.1.3 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 8 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 $H_{fext}(f, L)$.

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

Expression 8: 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 9 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 9: 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 clause summarize a few of these approximations.

8.1.4.1 Forced noise injection

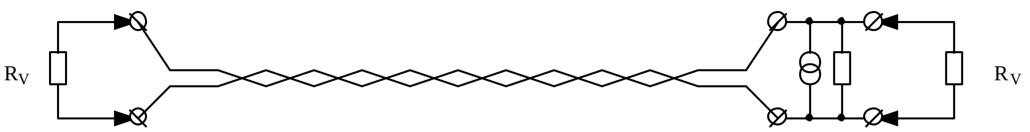
When cross talk is modelled by means of *forced* noise injection, then all impedance and frequency dependency of noise injection is ignored. The associated transfer function is shown in expression 10.

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

Expression 10: Transfer function for forced noise injection.

8.1.4.2 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 in figure [*].



ED NOTE. This issue is for further study, but should follow the ETSI-TM6 agreements for ADSL testing about noise injection and the way the injected noise level is defined by means of a complex calibration impedance.

End of literal text proposals