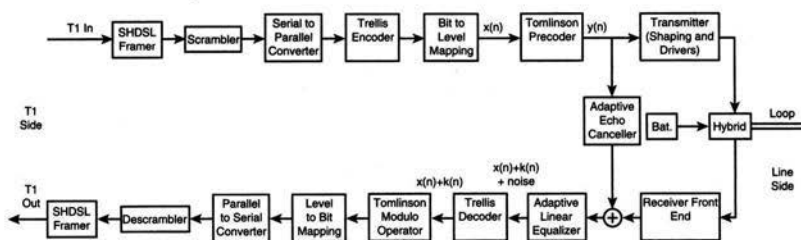


Figure 14.8 shows a single-pair HDSL (SHDSL) transceiver structure with Trellis coding.¹³

Figure 14.8 Trellis-Coded SHDSL Transceiver Block Diagram



In the top transmitter path for the Trellis-coded SHDSL system, the process is as follows:

1. The SHDSL framer converts from the 1.544 Mbps T1 format to the 1.552 Mbps SHDSL serial data format.
2. The scrambler randomizes the incoming data stream.
3. The serial-to-parallel converter produces three information bits per symbol time.
4. The Trellis encoder adds redundancy in the form of a Trellis bit, resulting in a total of four coded bits per symbol.
5. The bit-to-level mapper maps the four bits of output from the Trellis encoder to one of 16 levels in a one-dimensional PAM constellation.
6. The constellation symbols are passed through a Tomlinson precoder to counter the effect of the loop intersymbol interference.

In the bottom receiver path for the Trellis-coded SHDSL system, the process is as follows:

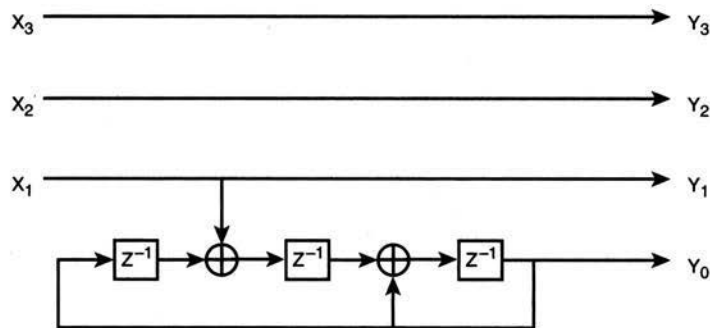
1. The adaptive linear equalizer suppresses channel precursors and whitens the noise.
2. The Trellis decoder picks the most likely received constellation sequence.
3. The Tomlinson modulo operator removes the modulo sequence added by the transmitter's Tomlinson precoder.
4. The level-to-bit mapper maps each of the 16 possible constellation points to a unique 4-bit number, then discards the Trellis bit leaving the 3-tuple information sequence.
5. The parallel-to-serial converter serializes the 3-tuple sequence.

6. The descrambler removes the randomizing sequence, leaving the original 1.552 Mbps sequence.
7. The SHDSL framer converts from the 1.552 Mbps SHDSL format to the 1.544 Mbps T1 format.

There is also an adaptive echo canceller connected in between the output of the Tomlinson precoder and the input of the adaptive linear equalizer for the separation of the receive path from the transmit signal.

Figure 14.9 shows an eight-state Trellis encoder with an asymptotic coding gain of 4 dB in the systematic feedback form.

Figure 14.9 An Eight-State Trellis Encoder in Systematic Feedback Form

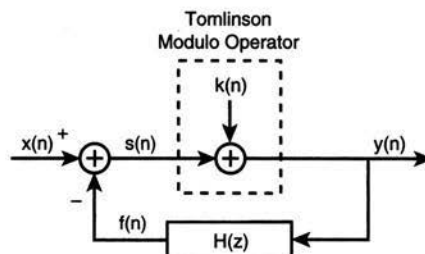


The effective coding gain is approximately 3.2 dB. The three input bits are passed directly to the output, and the Trellis bit becomes the least significant bit of the four output bits.

Tomlinson Precoding

Figure 14.10 shows the Tomlinson precoder.

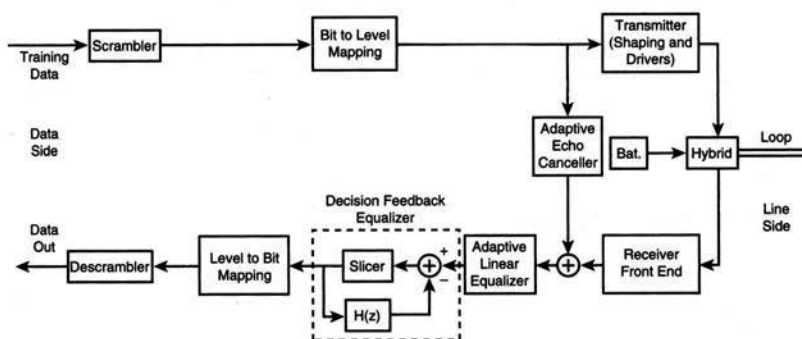
Figure 14.10 Tomlinson Precoder



The Tomlinson precoder consists of a feedback filter, $H(z)$, which is the replica of the feedback filter of the DFE, and a Tomlinson modulo operator. The input to the Tomlinson precoder is the 16-PAM signal, 8-PAM with the additional bit generated from the Trellis coding. The output of the Tomlinson precoder is the modulo remainder of the input sequence plus the feedback signal. The output of the Tomlinson precoder is also bounded between -16 and $+16$, assuming signal levels of $-15, -13, \dots, -1, 1, \dots, 13, 15$.

Coefficient values of the feedback filter are obtained during the startup training by the receiver using a random uncoded PAM sequence. These coefficients are sent to the transmitter at the other end of the twisted-pair loop to be used in the Tomlinson precoder prior to the regular data transmission. Compared with the regular data transmission, the feedback filter during the startup is in place of the Trellis decoder and Tomlinson modulo operator. Figure 14.11 illustrates the startup mode functional block diagram of an SHDSL transceiver.

Figure 14.11 Startup Mode Functional Block Diagram



14.2.2 Viterbi Decoding

A sophisticated coding scheme is needed to achieve a reasonable NEXT performance margin on all CSA test loops for SHDSL. A high coding gain usually requires a more complicated Trellis decoder. Table 14.4 summarizes a comparison of complexities of rate $3/4$ (three bits to four bits) Viterbi decoders of various numbers of states.¹⁴

Table 14.14 Rate $3/4$ Viterbi Decoder's Relative Complexity

Number of States	Relative Complexity	Relative Memory
8	0.08	0.07
16	0.16	0.17
32	0.30	0.41

Number of States	Relative Complexity	Relative Memory
64	0.60	0.96
128	1.18	2.18
256	2.36	4.89
512	4.71	10.82
1024	9.40	23.74

A Viterbi decoder with its number of states between 64 and 128 has the same complexity as the combination of the echo canceller and the equalizer in a regular HDSL transceiver. In other words, a rate 3/4 Viterbi decoder with 64 or 128 states and an asymptotic coding gain of 4.94 dB or 5.01 dB at the error rate of 10^{-6} is economically feasible in a current (1997) CMOS (Complementary Metal Oxide Semiconductor) technology.

14.2.3 Sequential Decoding

The primary characteristic of the sequential decoding algorithm is that its complexity is essentially independent of the number of states in the encoder, but rather is dependent on the signal-to-noise ratio.¹⁵

PairGain's T1E1 contribution studied the possibility of using sequential decoding for Trellis codes to obtain a performance margin of 6 dB for SHDSL.¹⁶ A real coding gain of at least 5.6 dB can be obtained at an error rate of 10^{-7} using the long constraint length rate 2/3 (two bits to three bits) codes with the sequential decoder whose buffer size is dictated by the 1 ms HDSL delay requirement. The computational requirements of a sequential decoder are substantially less than what would be required for a Viterbi decoder.

For a general Trellis-coded modulation system, the effective code rate of $k/(k+1)$ is established by using a rate $m/(m+1)$ convolution encoder, $m \leq k$, in conjunction with $k-m$ uncoded bits, where the convolutional encoder selects one of 2^{m+1} partitioned subsets, each consisting of 2^{k-m} constellation points. The $k-m$ uncoded bits are used to select the subset point.

Figure 14.12 shows the Trellis encoder part of a rate $k/(k+1)$ Trellis-coded modulation system.

Figure 14.12 Encoder and Channel

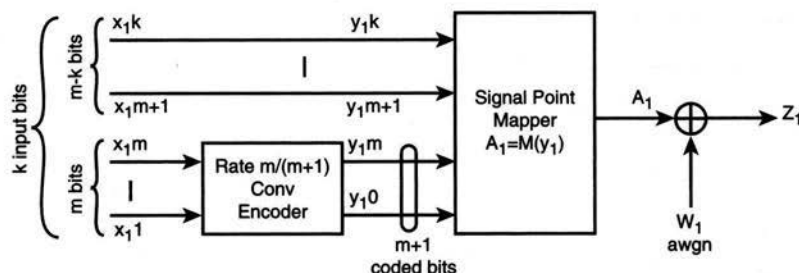
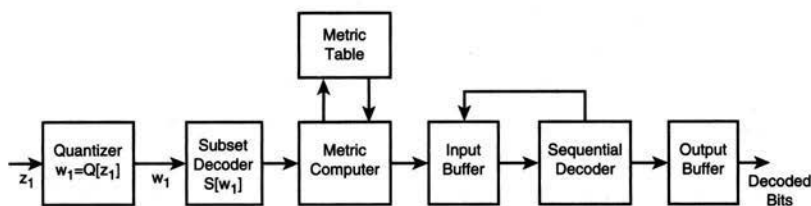


Figure 14.13 shows the sequential decoder part of a Trellis-coded modulation system.

Figure 14.13 Decoder Function



Fano Algorithm

The Fano algorithm¹⁷ can be used for the sequential decoder. The Fano algorithm examines only one path of the code Trellis at any time, resulting in substantially reduced memory requirements and metric computations. A sequence of metric thresholds spaced at intervals of a basic increment threshold offset of Δ is used to determine if the correct path is being followed by the decoder. If the metric computed by the decoder is increasing sufficiently on a forward search, the threshold is tightened (incremented) by Δ . During backward searches, the threshold is relaxed (lowered). The search is implemented in a way that prevents any node in the trellis from being searched forward twice with the same threshold setting.

The Fano algorithm employs a metric that is optimal in the sense that the path of maximum Fano metric is the most likely path explored so far. For Trellis Coded Modulation (TCM), the metric computed for each new branch in the path is given by¹⁸

$$\begin{aligned}
 M_B(A_1, z_1) &= \log_2 \frac{Kp\{z_1|A_1\}}{\sum_{i=0}^{K-1} \{z_1|A_i\}} - k \\
 &= \log_2 \frac{\exp(-|z_1 - A_1|^2 / 2\sigma^2)}{\sum_{i=0}^{K-1} \exp(-|z_1 - A_i|^2 / 2\sigma^2)} + n(1 - R)
 \end{aligned}$$

The values are as follows:

$$n = m + 1 \text{ and } R = n / m = m / (m + 1) \text{ so that } n(1 - R) = 1.$$

z_l is the received signal sample.

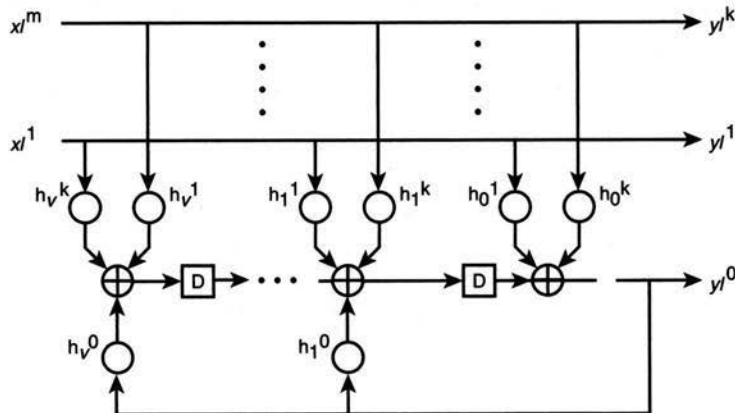
A_l is the branch label associated with the branch being extended.

The denominator is the summation of the probability that the sample z_l is received given A_i in the constellation. In practice, some method of recovery from buffer overflow must be implemented. The recovery method is necessary because, for the powerful codes under consideration, the dominant source of decoder errors becomes the overflow of the receiver symbol buffer. Data is framed for sequential decoding to ensure re-synchronization in a reasonable period of time.

The performance of a constraint length $\nu = 13$ corresponding to 8192-state code according to the code-combining approach of Porath and Aulin¹⁹ was examined. The code parity check vectors are $h^0=20310$, $h^1=11002$, and $h^2=04404$ all in octal representation. Figure 14.14 shows the correlation between parity check bits and encoder structure.

For a 16-QAM system, the coding gain for this $\nu = 13$ code is about 5.6 dB in conjunction with a 256-symbol buffer and a backsearch depth of 128 symbols. Similar coding gain has also been confirmed for the 64-QAM system.²⁰ When compared to Viterbi decoding of $\nu = 6$ rate 2/3 Ungerboeck code, the computational requirements for the sequential decoder are trivial.

Figure 14.14 Systematic Encoder with Feedback



The ADTRAN T1E1 contribution also examined the performance of a 64-QAM SHDSL with Trellis coding and a sequence decoder.²¹ The code parity check vectors for a rate 2/3 $v = 13$ code according to Pottie and Taylor²² are $h^0=020311$, $h^1=011002$, and $h^2=004404$ all in octal representation. Figure 14.15 shows the correlation between parity check bits and the encoder structure. Figure 14.16 illustrates the decoder.

Because the computational load of the Fano algorithm is a random variable from symbol to symbol, input and output queues are used to produce a constant bit stream at the output. The input and output queues are complementary, such that when one is full, the other is empty, and such that the sum of their depths is a constant. The coset slicer rounds the received signal plus noise to the closest point in each of the eight cosets of the rate 2/3 code. The branch metric calculator computes the Euclidean distance between each of the points from the coset slicer and the received signal, then applies each distance to the Fano metric.

The backsearch buffer of size BS symbol holds the past BS symbol duration worth of branch metrics and slicer outputs. When a new element is placed in the buffer, the Fano algorithm moves backward and forward through this buffer until a path can be found through the Trellis that will include this new element and the previous BS-1 elements. The output of the backsearch buffer is the most likely slicer output corresponding to the oldest element in the backsearch buffer.

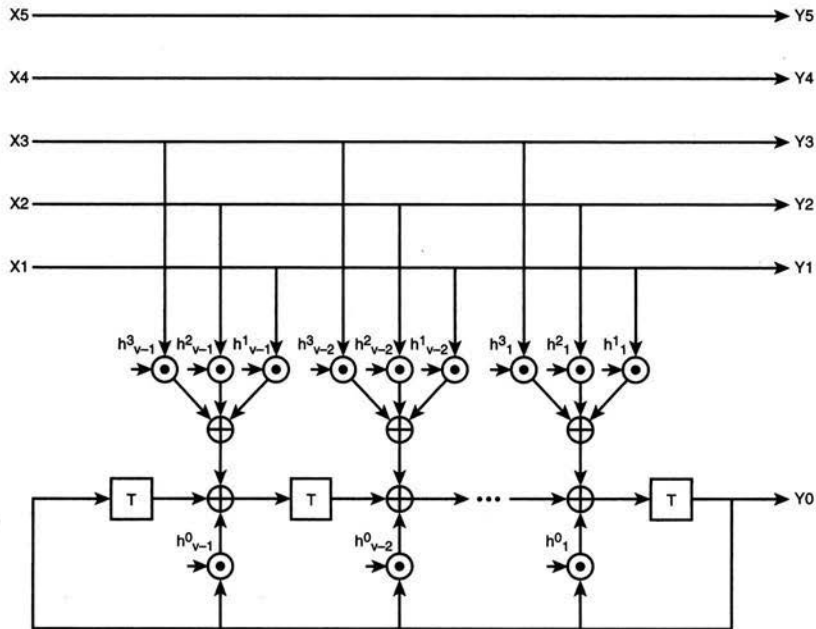
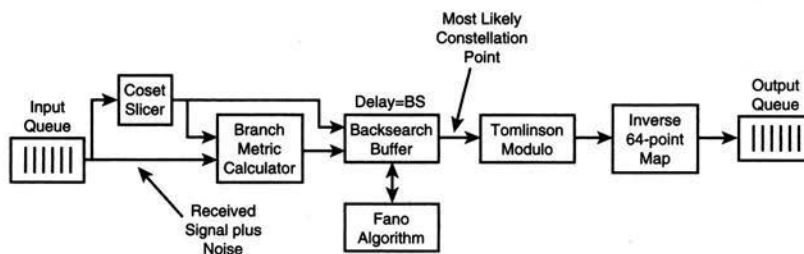
Figure 14.15 The Structure of a Rate 3/4 Trellis Encoder**Figure 14.16** Block Diagram of the Sequential Decoder

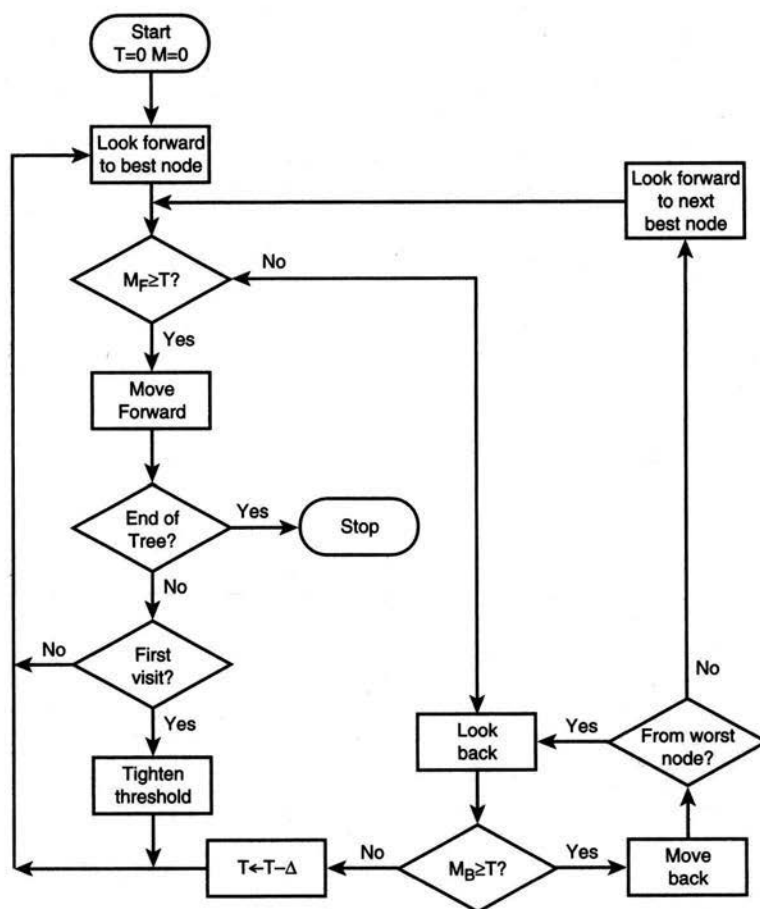
Figure 14.17 illustrates a block diagram of the Fano algorithm.

The variables for the Fano algorithm are as follows:

- The threshold T
- The cumulative path metric M
- The cumulative path metric after looking forward M_f

- The cumulative path metric after looking backward M_B
- The threshold offset Δ .

Figure 14.17 Block Diagram of the Fano Algorithm



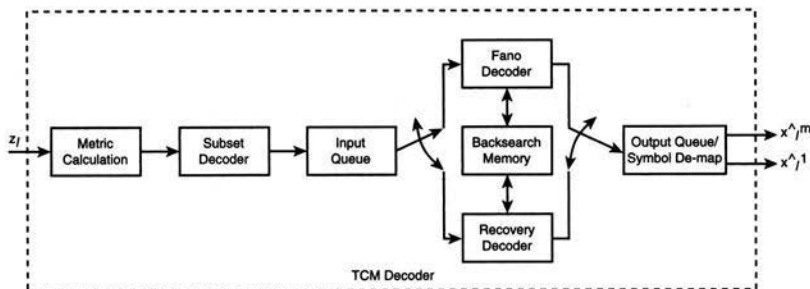
A coding gain of 5.1 dB has been observed and a degradation of 0.3 dB in coding gain can be caused by a constellation slicer compared with a slicer of infinite resolution. Similar coding gain can also be obtained for a 16-QAM Trellis-coded system with a similar sequential decoder.²³

PairGain's T1E1 contribution also indicates that the maximum coding gain achievable with a rate 2/3 code using 64-QAM line code is 6.0 dB.²⁴ However, smaller speed factors,

defined as the number of decoder operations per symbol interval, coupled with smaller input buffer and backsearch memory sizes, increase the probability of frame loss. The performance simulation results of a 64-QAM with Trellis coding and a sequence decoder show coding gain degradation of 0.7, 0.9, and 1.2 dB for speed factors of 128, 64, and 32, respectively. The code parity check vectors for this rate $2/3$ $v = 17$ code according to Wang and Costello²⁵ are $h^0=0627325$, $h^1=0275752$, and $h^2=0007350$, all in octal representation.

Figure 14.18 shows the structure of the simulated sequential decoder with an input queue after the metric operation.

Figure 14.18 Another Implementation of the Sequential Decoder



Further performance optimization can be obtained through the fine tuning of the decoder implementation.²⁶ The performance of a Trellis code transmission system with sequential decoder was also verified under the influence of typical channel impulse noise wave forms.²⁷

Computational Cutoff Rates

The number of computations per unit time in a sequence decoder is a random variable. The peak number of computations is unbounded. The average number of computations increases as the SNR decreases and becomes unbounded at the SNR at which the computational cutoff rate, R_0^* ,²⁸ falls below the information rate, R .

Table 14.15 summarizes computational cutoff rates for different line codes in addition to illustrating the difference between the computational cutoff rate SNR and the SNR for uncoded 10^{-7} error probability. This difference is not *coding gain* in the traditional sense because, given infinite memory, the sequential decoder does not make errors. The SNR difference does, however, provide a reference for achievable coding gain with limited memory size and computation power.

Table 14.15 Computational Cutoff Rates

Line Code	SNR (dB) where $R_o \cdot * < R$	SNR (dB) at 10^{-7} BER	SNR Difference
8-PAM	14.5	21.5	7.0
16-PAM	21.0	27.7	6.7
32-PAM	27.1	33.6	6.5
16-QAM	11.1	18.5	7.4
64-QAM	17.8	24.7	6.9
256-QAM	24.1	30.8	6.7

Turbo Codes for SHDSL

The communication within 1 dB of Shannon lemma (channel capacity) has been reported²⁹ with the use of *Turbo codes*. Turbo codes are also called *parallel concatenated codes*. An encoder of a Turbo code consists of parallel concatenation of two convolution codes with a random interleaver at the input of each encoder.

There are also proposals for *low latency Turbo codes*³⁰ and *Turbo-Trellis-Coded Modulation*.³¹ An example of 4 bits/sec/Hz QAM was shown to achieve better than 10^{-5} BER at between 7.4 and 7.6 dB Eb/No, suggesting a coding gain of between 7.8 and 8 dB. The gain of such a code would be sufficient for SHDSL. Figure 14.19 shows the general structure of the encoder.

At a smaller latency of about 90 symbols, the encoder can only generate an equivalent coding gain of about 5.5 dB at 10^{-7} BER.³² Preliminary study results show that the performance of Turbo code is inferior to the more mature sequential decoding technique under the condition of low latency and low BER. Such techniques^{33,34} as using structured interleaver instead of random interleaver might be able to make the Turbo code more suitable for the application of SHDSL.

14.2.4 Coding Gain and Latency

Table 14.16 summarizes coding gain estimates and corresponding latencies for different line codes and from different resources.³⁵

Figure 14.19 Parallel Concatenated Trellis-Coded Modulation

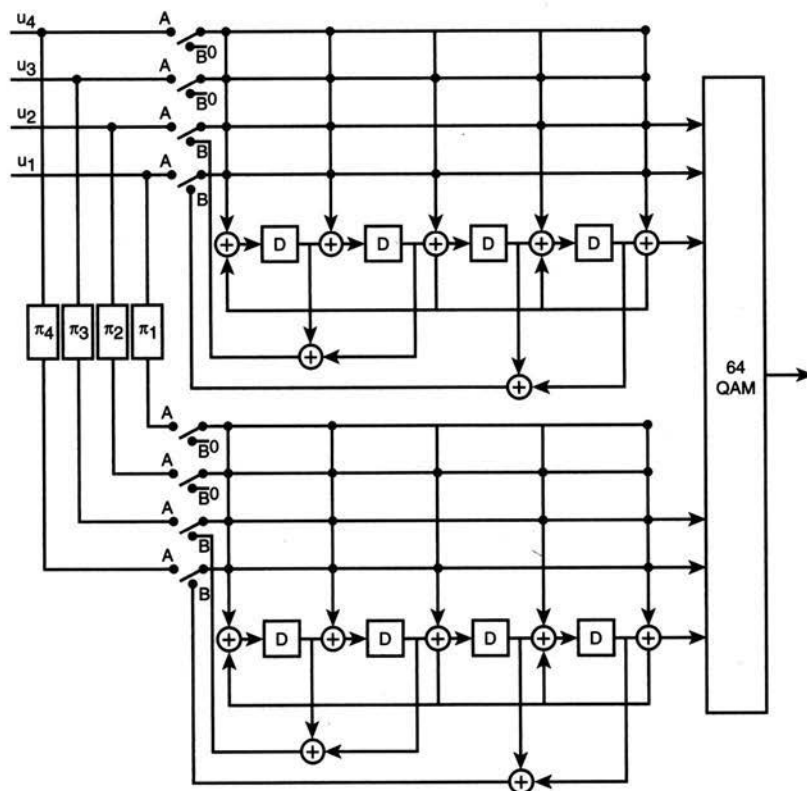


Table 14.16 Summary of FEC Techniques for SHDSL

Technique	Source	Line Code	Latency	Coding Gain (dB)
RS Block code	Metalink	2B1Q	300 μ s–1.5 ms	1.25–2.25
TCM	ADTRAN	16-PAM	25 μ s	3.20
TCM/sequential	PairGain	16-QAM	1 ms	5.60
TCM/sequential	Level One/ Heegard	64-QAM	483 μ s	5.50
TCM/sequential	ADTRAN	64-QAM	500 μ s–500 μ s	5.40–5.10
Turbo coding	PairGain	64-QAM	464 μ s–21 ms	4.50–7.50
Turbo coding	Globespan	64-QAM	324 μ s–469 μ s 639 μ s–834 μ s	5.00–5.20 5.30–5.50

The results of Table 14.16 indicate that coding gains of up to 5.2 dB are achievable if the 500 ms system latency is met. The code proposed by Level-One/Heegard with a coding gain of 5.5 dB and a delay of 483 μ s is a little too close to the delay requirement. Coding gains of up to 5.5 dB are achievable if the system latency can be relaxed to more than 500 μ s. The advantages of using long constraint length convolution codes decodable via sequential decoding for SHDSL can be summarized as follows:³⁶

- Currently achievable gains exceeding 5 dB
- Potential for gains of 6 dB and greater with future increases in VLSI speed or density
- Flexibility for multiple applications without specifying additional codes within the standard

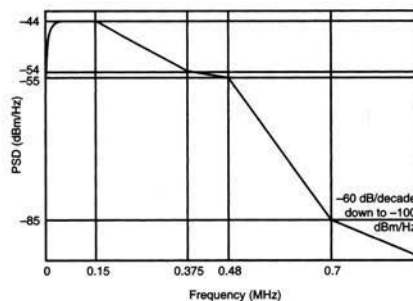
14.3 Minimizing Self and Foreign NEXT with Spectrum Shaping

Most SHDSL proposals are echo cancellation based. The Bellcore T1E1 contribution also studied the possibility of a Frequency Division Multiplex (FDM) approach SHDSL.³⁷ Both echo cancellation and FDM approaches fell short of the desired performance margin on CSA test loops, even with about 5 dB of coding gain. The echo cancellation-based systems are self-NEXT limited. On the other hand, the FDM systems have no self-NEXT, but are limited by NEXT from previously defined DSL systems.

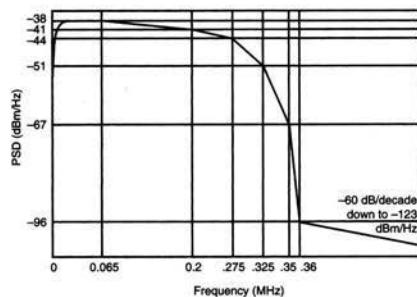
14.3.1 POET-PAM Systems

The ADTRAN T1E1 contribution proposed a Partially Overlapped Echo-Canceled Transmission (POET) approach to further enhance the performance of SHDSL.^{38,39,40} The POET proposal uses the combination of non-symmetric baud rates and spectral shaping of the transmitter PSDs to achieve both spectral compatibility and CSA reach performance requirements.

For a POET-PAM system, a coded 2B1Q line code is used. A 2D Trellis code can be used, resulting in a coding overhead of 0.5 and an information rate of 1.5 bits per symbol. The downstream baud rate is 1.0347 MHz, resulting in a non-filtered 3 dB corner frequency of 517.35 kHz. A shaping filter is used to reduce the transmit power density level further by 10 dB at above 375 kHz, resulting in a power spectrum density as shown in Figure 14.20. The total downstream transmit power is 9.7 dBm.

Figure 14.20 POET-PAM Downstream Power Spectrum Density

A coded 3B1O line code is used in the upstream direction. A 2D Trellis code can also be used, resulting in a coding overhead of 0.5 and an information rate of 2.5 bits per symbol. The upstream baud rate is 620.8 kHz, resulting in a non-filtered 3 dB corner frequency of 310.4 kHz. A gradual power spectrum density reduction of about 6 dB is created at 275 kHz by using a shaping filter. Figure 14.21 shows the upstream transmit power spectrum density. The total upstream transmit power is 14.7 dBm.

Figure 14.21 POET-PAM Upstream Power Spectrum Density

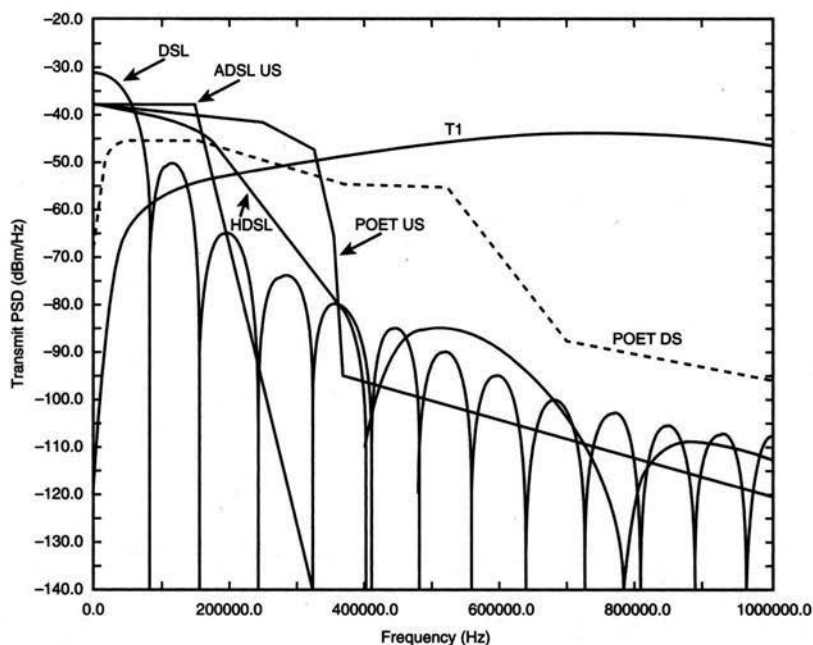
Simulations of Trellis-coded modulation techniques suitable for sequential decoding show that the 10^{-7} error rate can be achieved at an SNR of 13.4 dB for the downstream receiver and 19.6 dB for the upstream receiver. Table 14.17 summarizes the performance margins above those required. Notice that 5 dB of coded margin is achieved on CSA loop 6, and 6 dB of coded margin is achieved for both loops for all interferers except the EC ADSL.

Table 14.17 The SHDSL Performance Margin (dB)

Interferers	CSA Loop 6		CSA Loop 4	
	(US)	(DS)	(US)	(DS)
39 Self NEXT/FEXT	9.7	10.1	8.5	8.8
39 EC-ADSL NEXT/FEXT	5.0	16.5	3.9	16.2
49 HDSL NEXT	8.6	12.3	7.4	10.5
25 T1 NEXT	19.6	8.7	18.4	7.6
49 FDM-ADSLD NEXT, 49 ADSLU FEXT	10.3	15.9	9.0	15.6
49 CAP/QAM RADSL NEXT/FEXT	23.6	16.2	22.4	15.5
49 DSL NEXT	24.4	23.5	23.2	21.6

The performance of the POET system was examined for multiple type interference cases.⁴¹ Figure 14.22 compares the transmit power spectra of T1 and different DSLs.

Figure 14.22 Comparison of Transmit Power Spectra



It was found that the coexistence of T1 and SHDSL could make the performance of POET very poor in the downstream direction, as indicated in Table 14.18.

Table 14.18 The HDSL2 Performance Margin (dB)

Interferers	CSA Loop 6		CSA Loop 4	
	(US)	(DS)	(US)	(DS)
39 Self NEXT/FEXT	4.9	4.9	3.7	3.7
39 EC-ADSL NEXT/FEXT	0.1	11.0	-1.0	10.8
49 HDSL NEXT	3.7	7.0	2.4	5.3
39 Self NEXT/FEXT +10 T1 NEXT	4.9	-5.6	3.7	-6.7

14.3.2 TCSFDM Systems

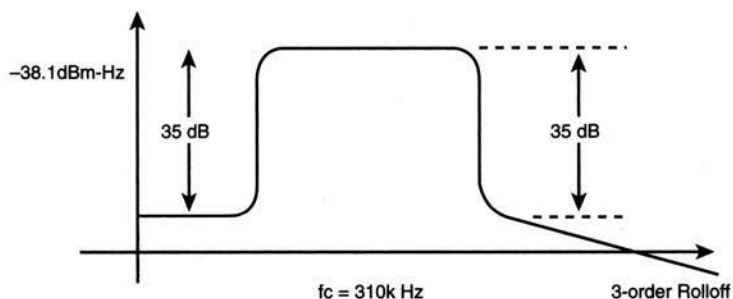
The PairGain Technologies T1E1 contribution proposes a Trellis-Coded Staggered FDM (TCSFDM) in conjunction with a 64-QAM line code.⁴² The total transmit power for both upstream and downstream is 16.8 dBm. Table 14.19 lists the starting, ending, and center frequencies.

Table 14.19 TCSFDM Spectral Placement

Direction	Frequency (kHz)		
	Starting	Ending	Center
Upstream	0	356.96	178.5
Downstream	131.52	441.92	310.0

Figure 14.23 shows the downstream transmit power spectrum.

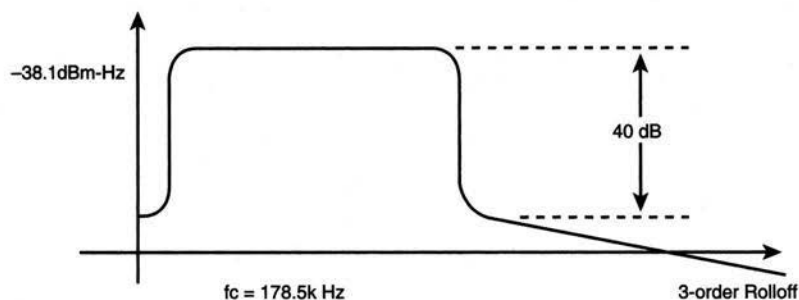
Figure 14.23 Downstream Transmit Power Spectrum



The downstream transceiver uses a square root-raised cosine spectrum with 15 percent excess bandwidth. At the first null of the square root-raised cosine, the attenuation flattens at 35 dB down, on the lower band edge. On the upper frequency band, there is a third-order lowpass filter. The rolloff frequency is 511 kHz.

Figure 14.24 shows the upstream transmit power spectrum.

Figure 14.24 Upstream Transmit Power Spectrum



Like the downstream transceiver, the upstream transceiver also uses a square root-raised cosine spectrum with 15 percent excess bandwidth. Here, the attenuation following the first null is masked at 40 dB down. On the upper frequency band, there is, same as in the downstream case, a third-order lowpass filter. The corner frequency is 389.6 kHz.

Table 14.20 summarizes the performance estimate of this TCSFDM system. With 5.1 dB coding gain, minimum performance margin is expected to be 4.6 dB.

Table 14.20 Performance Estimate

Interferer	DFE SNR Margin
49 ADSL	-5.07
25 T1	1.22
39 Self-NEXT/FEXT	-0.384

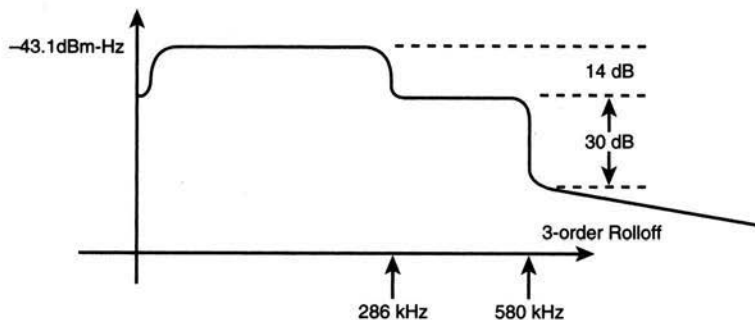
14.3.3 OverCAPped Systems

PairGain Technologies modified their original TCSFDM proposal with asymmetric spectrum shaping similar to that of POET-PAM.⁴³ Table 14.21 summarizes the transmit power spectrum characteristics of the modified system, also called OverCAPped.⁴⁴

Table 14.21 Transmit Power Spectrum Characteristics

	Upstream	Downstream
Transmit Power	16 dBm	11 dBm
Maximum PSD	-38.1 dBm/Hz	-43.1 dBm/Hz

It was believed that a passband line code could avoid signal losses at the near DC region. The line code still has a 64-point constellation size, including one coding overhead bit. Figure 14.25 shows the downstream transmit power spectrum.

Figure 14.25 Downstream Transmit Power Spectrum

On the lower frequency band, the downstream transceiver might use a square root-raised cosine spectrum with 15 percent excess bandwidth with 3 dB frequency at 25 kHz. On the upper frequency band, a third-order lowpass filter is used and the corner frequency is 580 kHz.

Figure 14.26 shows the upstream transmit power spectrum.

On the lower frequency band, the upstream transceiver might use a square root-raised cosine spectrum with 15 percent excess bandwidth with 3 dB frequency at 25 kHz. On the upper frequency band, there is, same as in the downstream case, a third-order lowpass filter with a corner frequency of 356 kHz.

Uncoded performance margins are estimated for the two most difficult CSA loops. Table 14.22 lists the results of these uncoded performance margin estimates.

Figure 14.26 Upstream Transmit Power Spectrum

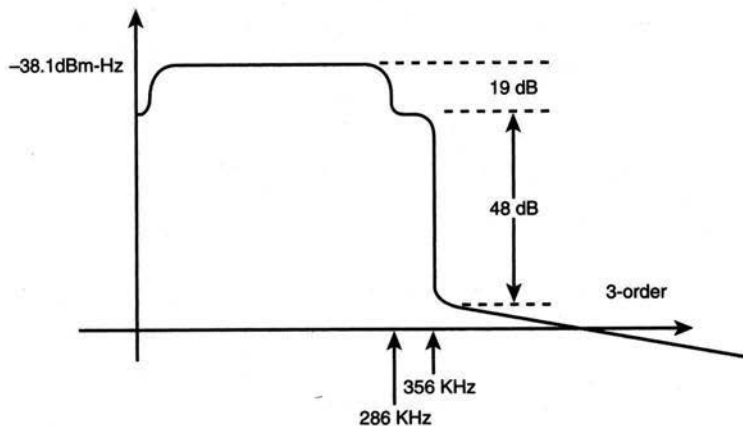


Table 14.22 The SHDSL Performance Margin (dB)

Interferes	CSA Loop 6		CSA Loop 4	
	(US)	(DS)	(US)	(DS)
39 Self NEXT/FEXT	4.05	3.68	3.14	2.64
39 EC-ADSL NEXT/FEXT	0.50	9.85	-0.38	10.2
49 EC-ADSL NEXT/FEXT	-0.06	9.26	-0.94	9.66
49 HDSL NEXT	4.05	3.68	3.14	2.64
25 T1 NEXT	16.80	11.65	15.88	10.72

The OverCAPped transmit spectrum is very similar to that of POET-PAM for both upstream and downstream. Figures 14.27 and 14.28 reveal the differences between OverCAPped and POET-PAM upstream and downstream power spectra, respectively.

Table 14.23 compares the difference in performance between OverCAPped and POET-PAM SHDSL systems, both without coding gain.

Figure 14.27 Upstream Transmit Power Spectra

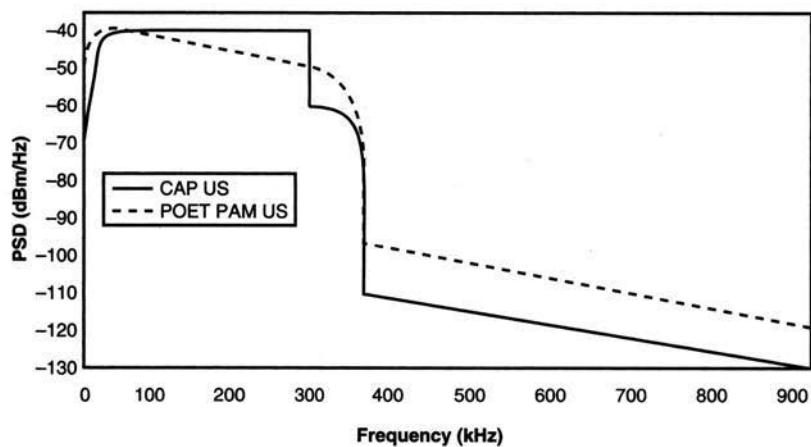


Figure 14.28 Downstream Upstream Transmit Power Spectra

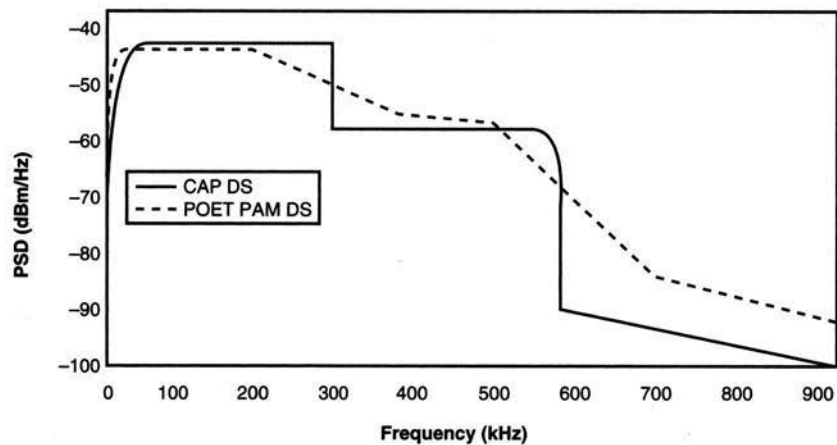


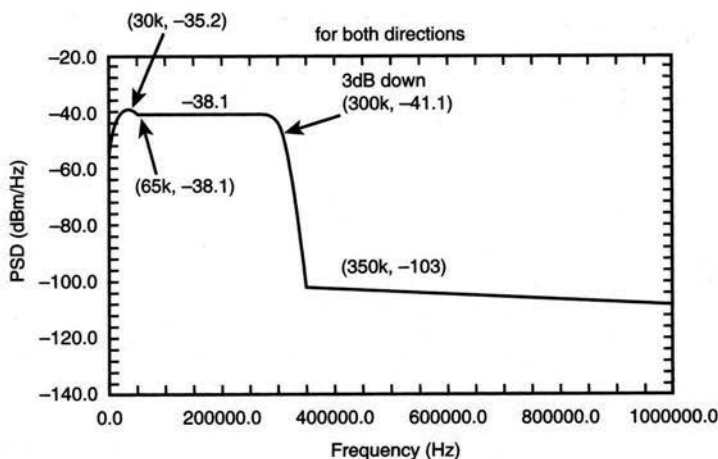
Table 14.23 SHDSL Performance Margin (dB)

Interferes	CSA Loop 4 Upstream		CSA Loop 4 Downstream	
	POET PAM	OverCAPped	POET PAM	OverCAPped
39 Self NEXT/FEXT	3.28	2.58	3.99	4.85
39 EC-ADSL NEXT/FEXT	-1.55	-0.94	11.26	9.32
49 EC-ADSL NEXT/FEXT	-2.10	-1.45	10.69	8.74
49 HDSL NEXT	2.10	2.18	5.47	2.21
25 T1 NEXT	13.49	13.85	1.85	9.17
49 FDM-ADSL NEXT/FEXT	2.70	3.35	10.70	8.76

14.3.4 SST Systems

The Level One Communications T1E1 contribution proposes a symmetric spectrum echo canceled system, called Spectral Shaped Passband Transmission (SST).⁴⁵ The baud rate for both upstream and downstream is 310.4 kHz with a coded 64-QAM/CAP line code. The total transmit power is 16.8 dBm. The transmit power density is raised to -35.2 dBm/Hz at around 30 kHz while the remaining power spectrum density, starting at 65 kHz, is -38.1 dBm/Hz, as illustrated in Figure 14.29.

Figure 14.29 SST Transmit Power Spectrum



The increased power density will not cause harm to any existing DSL systems because the raised portion matches that of ISDN Basic Rate Access lines. Table 14.24 summarizes the

estimated performance of the SST SHDSL for different interferer conditions. The SST system has a 3.2 dB performance margin on CSA loop 4 and a 4.3 dB margin on CSA loop 6 considering a coding gain of 5 dB.

Table 14.24 The SHDSL Performance Margin (dB)

Interferers	CSA Loop 6		CSA Loop 4	
	(US)	(DS)	(US)	(DS)
39 Self NEXT/FEXT	4.3	4.3	3.2	3.2
39 EC-ADSL NEXT/FEXT	7.0	20.4	5.9	20
49 HDSL NEXT	10.7	10.7	9.5	9.5
39 Self NEXT/FEXT +10 T1 NEXT	4.7	4.7	3.5	3.5

14.4 Status of the HDSL2 Standard

The agreement of key transceiver parameters among major proponents is a strong signal of the convergence of a possible SHDSL, also called HDSL2, standard. During the December 1997 T1E1.4 meeting, the basics of an SHDSL standard were agreed upon by co-authors from ADC Telecommunications, ADTRAN, Level One Communications, PairGain Technologies, and Siemens EZM of a joint T1E1.4 contribution.⁴⁶

This contribution recommends the following items for SHDSL:

- A performance margin of 5 dB on all CSA loops
- A programmable encoder for rate 3/4 Trellis codes with 512 states
- The OPTIS (Overlapped PAM Transmission with Interlocking Spectra) asymmetrical transmit power spectral shapes
- The standards project name of Next Generation HDSL for SHDSL

14.4.1 The Rationale for a 5 dB Performance Margin

With the 16-PAM, 512-state, Trellis-coded modulation and the OPTIS transmit spectral shape, 6 dB of performance margin can be achieved on all CSA loops according to computer performance simulation except for one particular loop and noise combination. Table 14.25 summarizes the estimated uncoded OPTIS performance margins for the two worst CSA loops, loops #4 and #6, under different crosstalk noise conditions.⁴⁷

Table 14.25 The OPTIS Uncoded Performance Margin (dB)

	CSA #4 Downstream	CSA #4 Upstream	CSA #6 Downstream	CSA #6 Upstream
39 SELF	9.8	1.7	9.8	2.7
49 HDSL	10.0	1.6	11.4	2.7
25 T1	15.9	18.8	16.6	19.8
49 FDM ADSL	16.1	8.0	14.2	8.9
39 EC ADSL	14.6	1.4	14.2	2.4
24 SELF + 24 T1	1.1	3.8	1.8	4.9
24 FDM ADSL +24 HDSL	9.9	0.8	10.2	1.9

For CSA loop #4 with 24 FDM ADSL and 24 HDSL, the performance margin is only 0.1 dB short considering a Trellis coding gain of 5.1 dB. On the other hand, the required 5 dB performance margin can provide enough room for hardware implementation tolerance.

14.4.2 The Programmable Encoder and 512-State Trellis Codes

The 512-state, 16-PAM, Trellis-coded modulation proposed by PairGain has been considered as a primary encoder for achieving the required 5.1 dB coding gain.⁴⁸ Figure 14.30 shows the structure of this 512-state convolution encoder.

Figure 14.30 Encoder Block Diagram

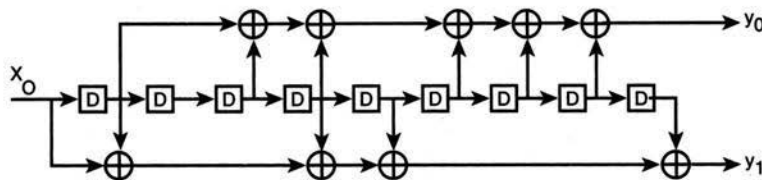


Figure 14.31 shows the structure of this Trellis encoder.

Figure 14.31 512-State Feedforward Convolutional Encoder

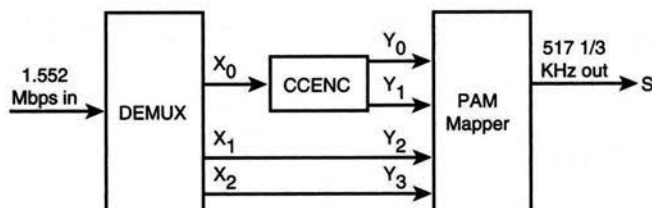


Figure 14.32 shows the bit-to-symbol mapping for the 16-PAM line code.

Figure 14.32 16-PAM Bit-to-Symbol Mapping

-15	-13	-11	-9	-7	-5	-3	-1	1	3	5	7	9	11	13	15	
0	1	0	1	0	1	0	1	0	1	0	1	0	1	0	1	y_0
0	0	1	1	0	0	1	1	0	0	1	1	0	0	1	1	y_1
0	0	0	0	1	1	1	1	1	1	1	1	0	0	0	0	y_2
0	0	0	0	0	0	0	0	1	1	1	1	1	1	1	1	y_3

Assuming Tomlinson precoding, this 16-PAM line code can achieve a coding gain of 5.1 dB with a latency of 217 ms. The generator polynomial for bit y_0 is denoted (in octal numbers) as $g_0=0556$. The generator polynomial for bit y_1 is denoted as $g_1=1461$. The corresponding generator matrix of this code is

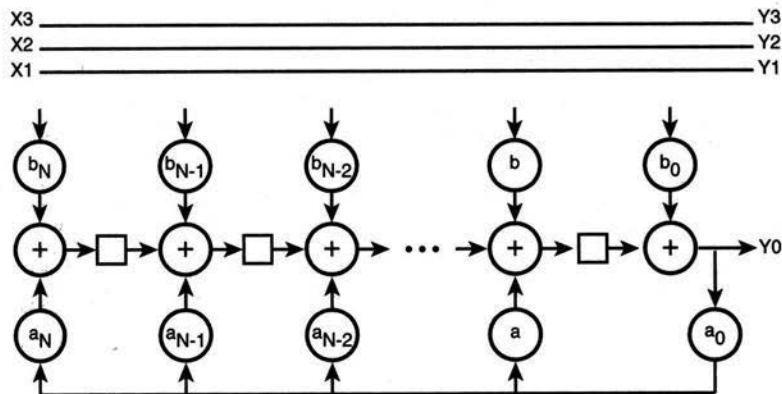
$$g = [g_0 \quad g_1 \quad g_2 \quad g_3] = \begin{bmatrix} 0 & 1 & 0 & 0 \\ 0 & 0 & 1 & 0 \\ 0 & 0 & 0 & 1 \\ 1 & 1 & 0 & 0 \\ 0 & 0 & 0 & 0 \\ 1 & 0 & 0 & 0 \\ 1 & 1 & 0 & 0 \\ 0 & 1 & 0 & 0 \\ 1 & 0 & 0 & 0 \\ 1 & 0 & 0 & 0 \\ 1 & 0 & 0 & 0 \\ 0 & 1 & 0 & 0 \end{bmatrix}$$

or

$$g(D) = \begin{bmatrix} D + D^3 + D^4 + D^6 + D^7 + D^8 & 1 + D + D^4 + D^5 + D^9 & 0 & 0 \\ 0 & 0 & 1 & 0 \\ 0 & 0 & 0 & 1 \end{bmatrix}$$

This Trellis code can also be implemented in a feedback format. Figure 14.33 shows the general structure of a programmable encoder in the feedback format.

Figure 14.33 Block Diagram of a Programmable Encoder



It has also been observed that many different codes can provide similar performance.⁴⁹ The adoption of a programmable encoder is proposed by ADTRAN⁵⁰ for accommodating a variety of coding schemes allowing differentiation and innovation among different vendors.

14.4.3 The OPTIS Transmit Spectra

Figure 14.34 shows the transmit power spectral density of the OPTIS line code in the downstream direction.

The downstream power spectrum consists of the base spectrum and a part of its mirror image. The bandwidth of the base spectrum is about 256 kHz. The mirror image extends the downstream bandwidth further to about 400 kHz. The power density of the mirror image is about 3 dB higher compared with that of the base spectrum. Figure 14.35 shows the transmit power spectral density of the OPTIS line code in the upstream direction.

The upstream power spectrum has only its base spectrum; however, the power spectrum density is enhanced by about 6 dB between 210 kHz and 245 kHz.

The downstream power spectrum can be implemented with two methods:

- Zero insertion and extension
- Expansion of single sideband

Figure 14.34 OPTIS Downstream Mask

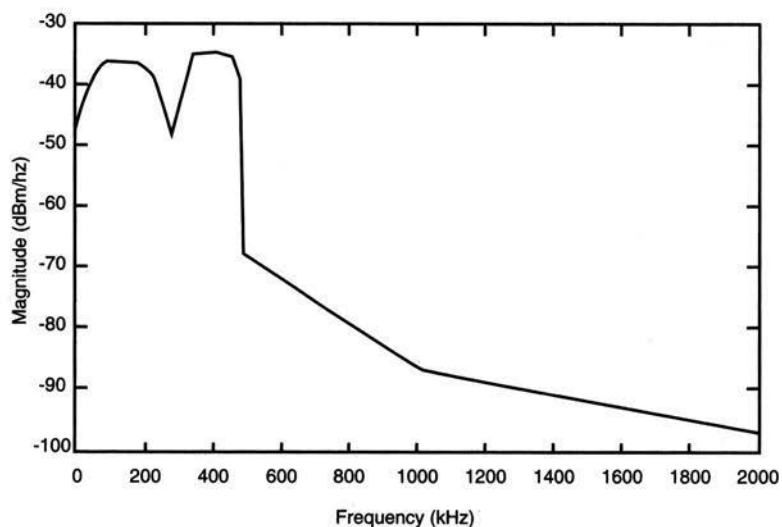


Figure 14.35 OPTIS Upstream Mask

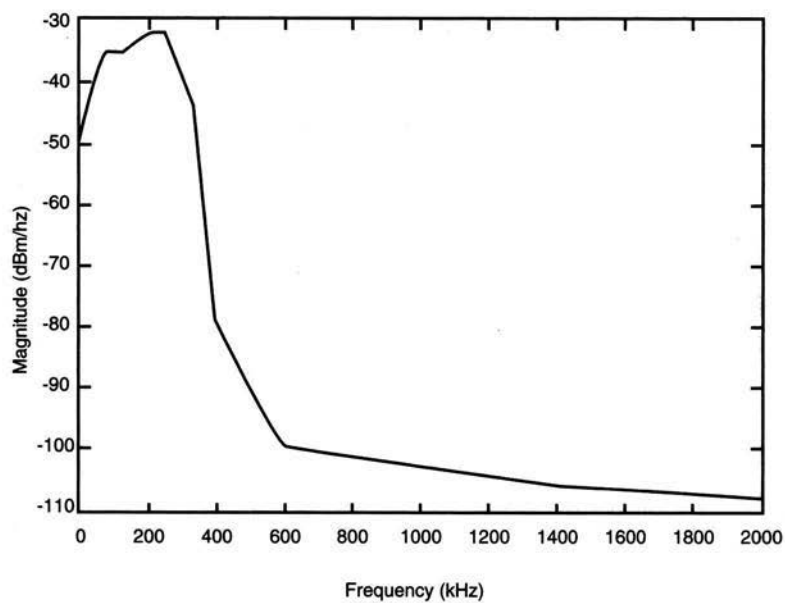
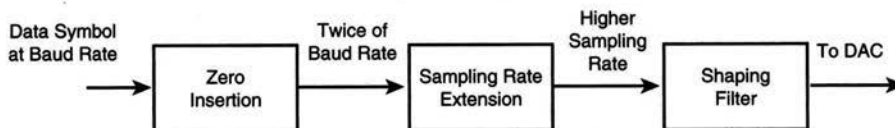


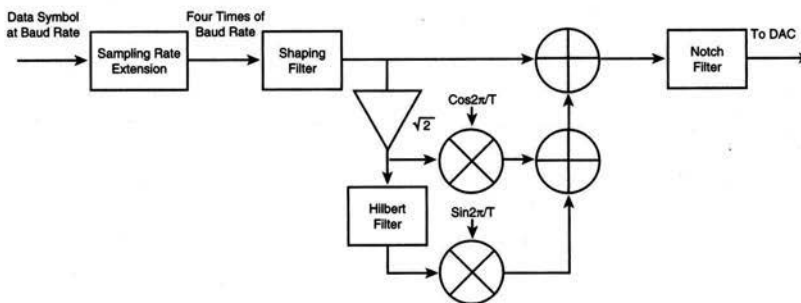
Figure 14.36 shows the functional block diagram of the zero insertion and extension approach.

Figure 14.36 Implementation with Zero Insertion and Extension



In Figure 14.36, a zero is inserted after every data symbol to double the baud rate. A higher sampling rate for spectral shaping filter is then obtained by extending symbol as well as zero points through sample and hold. For example, assuming a symbol sequence of $[-15, 5, -5, 1]$, the sequence becomes $[-15, 0, 5, 0, -5, 0, 1, 0]$ after zero insertion and becomes $[-15, -15, 0, 0, 5, 5, 0, 0, -5, -5, 0, 0, 1, 1, 0, 0]$ after sampling rate extension. Figure 14.37 shows the functional block diagram of the single sideband approach.

Figure 14.37 Implementation with the Single Sideband Approach

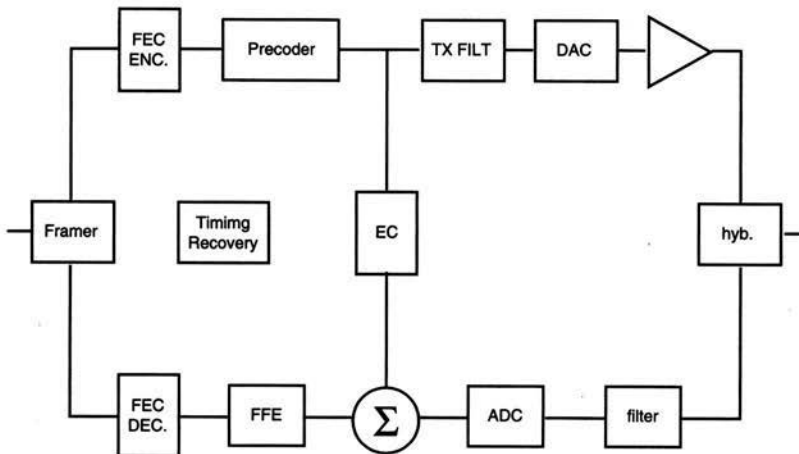


In Figure 14.37, data symbols are first extended to sampling rate through sample and hold. A shaping filter is used to control the excessive bandwidth and the energy of sidelopes. The partial mirror image is added through the single sideband modulation based on the digital Hilbert filter. The specific design of the digital Hilbert filter can control the mirror bandwidth. A notch filter creates the null between the base spectrum and the mirror image defined by the downstream power spectral density mask.

14.4.4 Analyzing SHDSL Transceiver Complexity

Figure 14.38 shows major functional blocks of an SHDSL transceiver.⁵¹

Figure 14.38 SHDSL Block Diagram



The implementation complexity of an SHDSL transceiver can be estimated by examining its digital computation intensity and analog front-end dynamic range and sampling rate requirements in comparison to those of an HDSL transceiver.

HDSL Transceiver Complexity

The digital computation requirements for an HDSL transceiver can be estimated as follows. Major computations are for the DFE and the echo canceller. The feedback filters (FBF) of a DFE and echo canceller are similar in terms of a low data resolution, 2-bit, adaptive filter with 128 filter coefficients.

The required calculation per second for a DFE or an echo canceller is described as

$$2 \times 128 \times 400 \text{ kHz} = 102.4 \text{ MIPS}$$

The value 400 kHz is the HDSL baud rate, 128 is the length of the adaptive filter, the number 2 accounts for both filtering and coefficient adapting operations, and MIPS stands for Millions of Instructions Per Second. The FBF and echo canceller combined calculation requirement is described as follows:

$$2 \times 102.4 = 204.8 \text{ MIPS}$$

In addition, the coefficient number for the feedforward filter (FFF) of DFE is about 8 and the data path resolution is about 12 bits. The FFF calculation requirement is described as follows:

$$2 \times 8 \times 400 \text{ kHz} = 6.4 \text{ MIPS}$$

This is equivalent to

$$6.4 \times 12 / 2 = 38.4 \text{ MIPS}$$

of 2-bit resolution filtering. The total computation requirement for an HDSL transceiver (in terms of 2-bit resolution filtering) is as follows:

$$204.8 + 38.4 = 243.2 \text{ MIPS}$$

The sampling rate of HDSL is 400 kHz. The peak-to-average ratio for a DAC without digital filtering is about 3.8 dB requiring a DAC resolution of only two bits. The required DAC resolution is seven bits even with a transmit filter. The ADC for an HDSL transceiver is about 13 bits as calculated in section 7.4.5, “ADC Precision Requirements Considering Channel Loss,” in Chapter 7, “Analog Front-End Precision.”

SHDSL Transceiver Complexity

The digital computation requirements for an SHDSL transceiver can be estimated as follows. Major computations are for the fractionally spaced equalizer (similar to FFF of a DFE), the Tomlinson precoder (similar to FBF of a DFE), the Viterbi decoder, and the echo canceller. The feedback filter (FBF) of a DFE and echo canceller are similar in terms of a low data resolution, 2-bit, adaptive filter with 128 filter coefficients.

For the same CSA loop channel, the Tomlinson precoder and echo canceller length is increased to

$$128 \times 517.3 / 400 = 165 \text{ taps}$$

due to a higher baud rate, assuming echo cancellation is performed at the baud rate.

The required calculation per second for a Tomlinson precoder is described as

$$165 \times 517.3 \text{ kHz} = 85.35 \text{ MIPS}$$

assuming the real-time updating is not necessary for the Tomlinson precoder.

The required calculation per second for an echo canceller (assuming real-time echo canceller coefficient adaptation) is described as follows:

$$2 \times 165 \times 517.3 \text{ kHz} = 170.7 \text{ MIPS}$$

In addition, the coefficient number for the fractionally spaced equalizer is about 20 due to the higher sampling rate and $T/2$ fractional operation. The fractionally spaced equalizer calculation requirement is described as follows:

$$2 \times 20 \times 1034.6 \text{ kHz} = 41.384 \text{ MIPS}$$

Because of the Tomlinson precoder, all these calculations require a full data resolution of between 12 bits (for filtering) and 14 bits (for echo cancellation). The total computation requirement for the combined equalization and echo cancellation operation (in terms of full resolution filtering) is

$$85.35 + 170.7 + 41.384 = 297.43 \text{ MIPS}$$

The computational complexity of a rate $3/4$ Viterbi decoder depends on the state number of the encoder.⁵² Table 14.26 summarizes Viterbi decoder complexities for convolutional encoders of 64 to 512 states in comparison to that required by the combined equalization and echo cancellation operation of an SHDSL transceiver.

Table 14.26 Rate $3/4$ Viterbi Decoder Complexity

States	Computation	Storage
64	0.6	0.96
128	1.18	2.18
256	2.36	4.89
512	4.71	10.82

The complexity of a 512-state Viterbi decoder is about 4.7 times that required by the combined equalization and echo cancellation operation. The total computation requirement for an SHDSL transceiver (in terms of full resolution filtering) is about

$$297.43 \times 5.71 = 1698.348 \text{ MIPS}$$

The sampling rate of SHDSL is 1034.6 kHz. The peak-to-average ratio for a DAC is about 18 dB, requiring a DAC resolution of about eight bits. The ADC for an SHDSL transceiver is about 14 bits because of the 16-PAM line code and a wider signal bandwidth.

SHDSL-to-HDSL Computation Complexity Ratio

The SHDSL-to-HDSL MIPS requirements ratio is

$$1698.348 / 243.2 = 6.983$$

The implementation complexity can be seven to eight times higher than this ratio because of the filtering data path resolution factor of between $12/2=6$ and $14/2=7$. The combined ratio is between 40 and 50.

On the other hand, the VLSI technology advances according to Moor's law that the complexity of the semiconductor circuit doubles every 18 months. Since the development of the HDSL chip from 1990 to 1998, we have an elapsed time of about five 18-month periods. Therefore, the advance of semiconductor technology can absorb a complexity increase of $2^5 = 32$ times. With that semiconductor technology advance, the development of an SHDSL chip can still be as challenging as the development of an HDSL chip about eight years ago.

Table 14.27 summarizes the SHDSL VLSI complexity estimate in comparison to that of HDSL by Level One Communications and PairGain Technologies.

Table 14.27 VLSI Complexity Comparison

	Framer	FEC	DFE	FFF	EC	MISC	Tot. Digi.	AFE
HDSL	2	NA	1.0	0.93	1.15	1.9	6.98	1.0
HDSL2	2	9.3	2.2	2.0	6.4	6.5	28.4	3.85

The estimate is based on estimated gate count of digital semiconductor circuits and the layout size of analog semiconductor circuits.

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VDSL (Very-High-Bit-Rate Digital Subscriber Lines)

If SHDL is the next generation HDSL, VDSL can also be considered as the next generation ADSL to some extent. The transmission throughput of VDSL is much higher than that of ADSL. The potential of a higher transmission throughput is achieved by expanding the signal bandwidth to the region of 10 to 30 MHz. At such a high frequency, a usable channel can only be realized on short twisted-pair telephone subscriber loops.

VDSL also has asymmetrical transmission throughput, although the symmetrical throughput could be an option. Due to a very high transmission bandwidth, which causes a very high sampling rate, the attempt of approaching the channel capacity has not been a top-priority concern in related standard activities.

The idea of signaling at a very high rate has been explored during the early 1990s.¹ Prototype *fiber-to-the-curb* (FTTC) systems with a physical layer similar to a VDSL system have been used in conjunction with a broadband optical distribution network for field tests. The Digital Audio-Visual Council (DAVIC) has referenced such an FTTC system.²

Since 1995, VDSL has been a regular discussion topic at ANSI T1E1.4 working group meetings. A VDSL system requirement document³ and a living list have been updated during these meetings. The European standards group, European Telecommunications Standardization Institute (ETSI), also addresses issues related to VDSL at its regular meetings. An ETSI VDSL system requirement document has also been created.⁴

VDSL is a natural extension of the evolving telecommunication network to the consumer market partially based on an existing loop plant infrastructure. The high transmission throughput telephone optical network can be initially terminated at the *Distribution Area* level and can be connected to end users through the VDSL technology.

The high throughput VDSL is easier to be made compatible with SONET and ATM-based services. VDSL can be used to interconnect business customers within a concentrated area through leased telephone lines for high-speed intranet use. VDSL can also be used in an enterprise or campus environment for providing high-speed digital transmission connections.

This chapter covers many topics related to VDSL, including the following:

- *Transmission environment.* Usable transmission bandwidths are examined for short telephone subscriber loops. Bases for slightly different high-frequency crosstalk models are reviewed. Behaviors of radio frequency ingress and egress through twisted pairs are studied.
- *General system architecture.* Channel capacities in conjunction with self-FEXT and background noise are calculated for short loops. Possible transmit power density levels are reviewed.
- *System proposals.* Key features of CAP/QAM and DMT VDSL line code proposals are reviewed.
- *Twisted-pair primary constants for VDSL.* Simplified primary parameter tables for different types of twisted pairs are provided for easy reference.

15.1 *Transmission Environment for VDSL*

I anticipate that the evolution of the existing telephone loop plant will bring more fiber links to telephone subscriber loops. Bigger portions of subscriber loops near a CO will be replaced by fiber links. The fiber link of a subscriber loop will be getting longer while the copper link will be getting shorter.

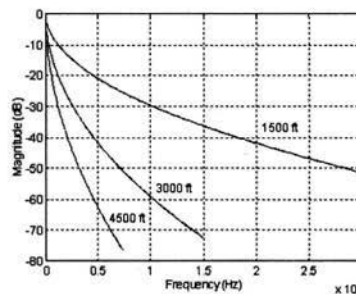
The size of a Distribution Area, where the distribution point can be connected to a CO through fiber links, will be shrinking from about 500 subscribers to a few dozen subscribers. The optical transceiver at the distribution point, also called the *Optical Network Unit* (ONU), can still be connected with remaining twisted-pair distribution cables and drop wires to utilize existing loop plant infrastructure.

Remaining copper twisted-pair loops from an ONU to a telephone subscriber are expected to be less than 5 kft. Consequently, a wider bandwidth becomes available for shorter twisted-pair loops. Primary parameters of twisted-pair cables, especially at higher frequencies, have been recently published^{5,6,7,8} to facilitate the feasibility studies of a Very-high-bit-rate Digital Subscriber Line (VDSL) on an emerging short-distance copper twisted-pair telephone loop plant.

15.1.1 VDSL Channel Model

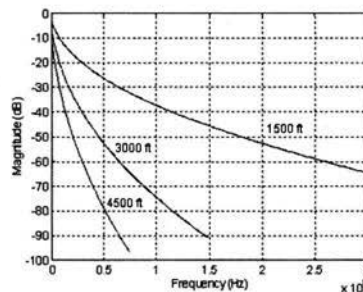
According to transmission characteristics defined by the recently published primary parameters of twisted-pair cables that make VDSL feasible on an emerging short-distance copper twisted-pair telephone loop plant, the 70 dB loss channel bandwidths are about 6.25 MHz, 14 MHz, and larger than 30 MHz for the 24-gauge 4500, 3000, and 1500 ft. twisted-pair loops,⁹ respectively, as shown in Figure 15.1.

Figure 15.1 Insertion Loss of 24-Gauge Twisted-Pair Loop



The transmit signal with a power density level of -60 dBm/Hz can maintain a signal-to-noise ratio of about 10 dB with respect to a receiver front-end background noise of -140 dBm/Hz at the 70 dB channel loss frequency point. The 70 dB loss channel bandwidths are 4 MHz, 9 MHz, and larger than 30 MHz for the 26-gauge 1500, 3000, and 4500 ft. twisted-pair loops, respectively, as shown in Figure 15.2.

Figure 15.2 Insertion Loss of 26-Gauge Twisted-Pair Loop



It has been observed that short bridged taps can cause deep notches for VDSL length loops.¹⁰ Figure 15.3 shows the effect of different lengths of bridged taps at the middle of a

2 kft, 26-gauge twisted-pair loop. Bridged tap lengths range from 8 ft. to 128 ft. The shortest bridged tap of 8 ft. causes the deepest frequency null at about 20 MHz.

Figure 15.3 The Effect of Bridged Taps on Short Twisted-Pair Loops

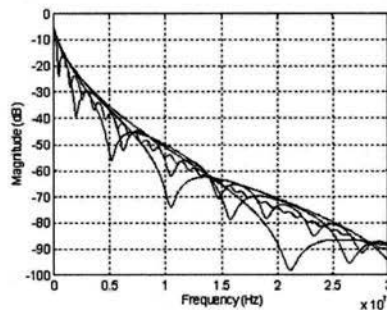
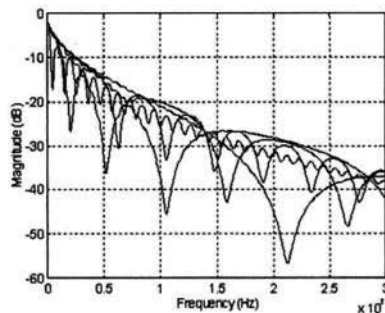


Figure 15.4 shows the effect of different lengths of bridged taps at the middle of a 1 kft, 24-gauge twisted-pair loop. Bridged tap lengths also range from 8 ft. to 128 ft. The notches could be as deep as 25 dB.

Figure 15.4 The Effect of Bridged Taps on Short Twisted-Pair Loops



15.1.2 VDSL Crosstalk Model

Recent measurements of 24-AWG PIC aerial distribution cables have indicated some slight differences compared with the 15 dB per decade simplified Unger model.¹¹ The NEXT power sum transfer function can still be described by

$$|H_{NEXT}(f)|^2 = \chi_n f^n$$

At the frequency range of 1.5 to 30 MHz, however, we have

$$x_{24} = 1.41 \times 10^{-13} \text{ and } n = 1.37$$

instead of

$$x_{24} = 5.937 \times 10^{-13} \text{ and } n = 1.5$$

as defined previously. The difference between coupling coefficients, x_{24} s, is about 6.24 dB while the loss differences are more than 5 dB.

The FEXT power sum transfer function is generally described by

$$|H_{FEXT}(f)|^2 = |H_{channel}(f)|^2 \times k \times l \times f^2$$

The values are as follows:

$H_{channel}(f)$ is the channel transfer function.

k is the coupling constant.

l is the twisted-pair length.

f is the frequency.

For $l = 1000$ ft. and at the frequency range of 1.5 to 30 MHz, we have

$$k = 4.33 \times 10^{-19} \text{ and } n = 1.82$$

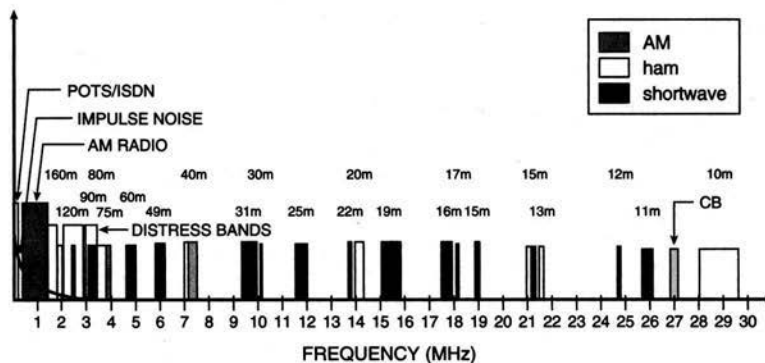
instead of

$$k = 5.21 \times 10^{-20} \text{ and } n = 2.0$$

as defined previously. The maximum loss difference is about 4 dB.

15.1.3 Radio Frequency Interference

For VDSL, the usable bandwidth is between 4 MHz and 30 MHz, depending on loop length and the twisted-pair cable type. Under such wide a bandwidth, there could be some interference between VDSL transceivers and traditional radio receivers. There are many Amplitude Modulated (AM) Medium Wave (MW) radio stations in the frequency band of between 550 kHz and 1700 kHz. Short Wave (SW) AM radio stations are located within a few frequency bands from 1.7 MHz and 30 MHz, as shown in Figure 15.5.

Figure 15.5 Frequency Bands for AM, Ham, Short Wave Radio, and Public Safety/Distress Bands

There are also amateur short wave frequency bands distributed in the same frequency range, as shown in Figure 15.5, and further summarized in Table 15.1. In addition, there are also some other specific-purpose radio transceivers.

Table 15.1 Amateur Ham Radio Frequency Bands

Wave Length (Meters)	Frequency Band (MHz)
180	1.8–2.0
80	3.5–4.0
40	7.0–7.3
30	10.10–10.15
20	14.00–14.35
17	18.068–18.168
15	21.00–21.45
12	24.89–24.99
10	28.00–29.70

The transmit power of AM MW radio stations is normally between 300 kW–500 kW. The high-power AM MW radio station, however, can have a transmit power of up to 5 MW.¹² The transmit power of AM SW radio stations is also normally between 300 kW and 500 kW. On the other hand, the transmit power of SW amateur radios is normally below 400 W or 1500 W, depending on local regulations.

The incident power at a location of distance d away from a transmit antenna of unit gain can be expressed as follows:¹³

$$P_{inc} = \frac{P_{trans}}{\pi d^2} |A_s|^2$$

in which

$$|A_s| = \frac{2 + 0.3p}{2 + p + 0.6p^2} - \sqrt{\frac{p}{2}} e^{-0.6p} \sin b$$

for

$$p \approx \frac{k_0 d \omega \epsilon_0}{2\sigma}$$

$$b \approx \tan^{-1} \frac{\kappa' \omega \epsilon_0}{\sigma}$$

With

$$\frac{\sigma}{\omega \epsilon_0} = \frac{1.8 \times 10^4 \sigma}{f|_{MHz}}$$

we have

$$p \approx \frac{2.78 \times 10^{-5} k_0 d f|_{MHz}}{\sigma}, \quad b \approx \tan^{-1} \frac{5.56 \times 10^{-5} \kappa' f|_{MHz}}{\sigma}$$

The dielectric constant κ' is usually in the range of 10 to 15. We usually have

$$\sigma = 10^{-2} S/m \text{ and } k_0 = \frac{2\pi}{\lambda_0}$$

The wavelength λ_0 can be calculated as follows:

$$\lambda_0 = \frac{3 \times 10^2}{f|_{\text{MHz}}}$$

Assuming $\kappa' = 10$, we therefore have

$$p \approx 5.82 \times 10^{-5} d(f|_{\text{MHz}})^2$$

$$b \approx \tan^{-1}(5.56 \times 10^{-2} f|_{\text{MHz}})$$

The radio electromagnetic field strength, in the unit of $\mu\text{V} / \text{m}$, is related to the incident power, in the unit of W / m^2 , through the intrinsic impedance

$$Z_0 = \sqrt{\frac{\mu_0}{\epsilon_0}} = 377\Omega$$

of the free space by

$$\frac{|E|^2}{2Z_0} = P_{inc}$$

For a dipole antenna with matched antenna and terminal impedance of R_a , we have

$$R_a = \frac{Z_0}{1.64\pi}$$

and

$$|E| = \sqrt{3.28\pi R_a P_{inc}}$$

With some transmission and coupling losses, these radio signals appear as noise interference at the front end of a VDSL transceiver. The VDSL noise power density level has been

expressed in the unit of dBm/Hz , which is 1 milliwatt per Hertz. On the other hand, the strength of radio electromagnetic field is expressed in the unit of $dB\mu V/m$, which is 1 microvolt per meter.

For a particular frequency band, a certain measuring distance, and a matched antenna and terminal impedance of 75 ohms, the equal power conversion factor between dBm and $dB\mu V$ is the following:

$$0dBm = 10 \times \log_{10} \frac{V^2}{100} \times \frac{1}{0.001} - > 0.2739 \text{ volts -}$$

$$> 20 \times \log_{10} \frac{0.2739}{0.000001} = 108.75dB\mu V$$

For using a particular type of antenna, the antenna gain should also be included in the conversion factor. At 10 MHz, an isotropic antenna has an antenna gain of -5 dB.¹⁴ For the isotropic antenna, the field power in the unit of $dB\mu V/m$ to antenna termination power in the unit of dBm conversion factor is $108.75-5=103.75$ dB. For a fixed field strength, the power delivered by an isotropic antenna is proportional to the square power of the wavelength. Therefore, the same field strength will produce 2.5 dB and 6 dB more power delivered by an isotropic antenna at frequencies of 7.5 MHz and 5 MHz, respectively. In the other direction, the same field strength will produce 6 dB and 14 dB less power at frequencies of 20 MHz and 50 MHz, respectively.

By comparing radio field strength with induced power at a cable termination, we can find the conversion factor. By comparing the conversion factor of a particular cable with that of an equal power conversion, we can find the cable *Radio Frequency Interference* (RFI) loss, which is equivalent to the negative antenna gain.

Table 15.2 summarizes power losses for different cables at different frequencies along with the average and worst cable power losses. Results in Table 15.2 are calculated based on induced power measurements made by Broadband Technologies for a known strength of radio field.¹⁵

Table 15.2 Cable RFI Power Loss (dB)

Frequency (MHz)	BKMA 100 Pair	ARTM 25 Pair	Steel Flat Drop	Twisted Drop	Twisted Customer
1.9	74	55.92	36.48	45.46	39.57
3.9	55.92	73.98	50.10	47.96	53.15
7.2	52.04	52.04	40.91	60.00	52.04
10.125	54.90	59.17	41.01	52.04	60.00
14.2	47.96	48.97	35.01	49.00	49.12
18.13	59.17	67.96	40.91	54.96	55.43
21.3	61.13	57.08	48.09	61.03	60.00
24.96	74.27	75.85	48.56	53.55	63.10
29.0	59.08	55.98	47.90	56.90	55.56
Average	59.83	60.77	43.22	53.43	5421
Worst	47.96	48.97	35.01	45.46	39.57

Through the ionospheric reflection, the SW radio interference field strength can be calculated by combining the ray-path distance loss and ionospheric absorption loss for a one-hop propagation. For a ray-path distance loss of between 80–120 dB and an ionospheric absorption loss between 20–40 dB, the combined loss is between 100–160 dB. For a SW radio transmit power of 300 kW, or 84.77 dBm, the incident power at a twisted pair could vary between -75.23 and -15.23 dBm/m² or 43.54 and 103.54 dBμV/m.

Assuming a radio station transmit power of 500 kW, a bandwidth of 10 kHz, and a cable power loss of 40 dB, we can calculate the RFI induced noise density. For d=10 miles =16090 m at a frequency of 1.9 MHz, we have the following:

$$p \approx 5.82 \times 10^{-5} \times 1.609 \times 10^4 (1.9)^2 = 3.38$$

$$b \approx \tan^{-1}(5.56 \times 10^{-2} \times 1.9) = \tan^{-1}(0.1056) = 6.03^\circ$$

$$|A_s| = \frac{2 + 0.3 \times 3.38}{2 + 3.38 + 0.6 \times 3.38^2} - \sqrt{\frac{3.38}{2}} e^{-0.6 \times 3.38} \sin 6.03^\circ = 0.228$$

$$P_{inc} = \frac{500 \times 10^3}{\pi \times 16090^2} 0.228^2 = 3.196 \times 10^{-5} \text{ Watts / m}^2 = -14.95 \text{ dBm / m}^2$$

$$|E| = \sqrt{2 \times 377 \times 3.196 \times 10^{-5}} = 0.1552 \text{ Volt} / \text{m} = 103.82 \text{ dB}\mu\text{V} / \text{m}$$

The RFI power density is

$$P_{RFI} = 103.82 - 108.75 - 40 - 10 \log_{10} 10^4 = -84.93 \text{ dBm} / \text{Hz}$$

Table 15.3 summarizes the RFI power density levels for other frequencies and distances.

Table 15.3 RFI Power Density (dBm/Hz)

Frequency (MHz)	0.01 Mile 500 W	0.1 Mile 500 W	1 Mile 500 kW	10 Miles 500 kW
1.9	-42.14	-62.31	-53.63	-84.93
3.9	-42.28	-63.04	-59.09	-98.96
7.2	-42.75	-65.44	-69.60	-111.02
10.125	-43.36	-68.83	-75.24	-117.32
14.2	-44.47	-74.78	-81.82	-123.41
18.13	-45.79	-79.63	-86.66	-127.74
21.3	-47.02	-81.69	-89.76	-130.57
24.96	-48.62	-82.92	-92.74	-133.36
29.0	-50.60	-84.28	-95.52	-135.98

AM MW and SW radio stations as well as some high-power amateur radio stations can generate radio frequency interference to a VDSL receiver. On the other hand, the VDSL signal radiated from telephone subscriber loops can also affect some sensitive radio receptions, such as that of an amateur radio receiver. With the worst and average cable power loss factors, we can also estimate VDSL signal-generated field strength. We have

$$|E| = 20 \log_{10} \frac{\sqrt{10^{\frac{P}{10}} \times 0.001 \times B_w R_t 10^{\frac{K_{cable}}{10}}}}{10^{-6} \sqrt{4\pi r}}$$

The values are as follows:

$|E|$ is the field strength in the unit of $\text{dB}\mu\text{V} / \text{m}$.

P is the VDSL transmit power density in the unit of dBm / Hz .

B_w is the bandwidth of interest in the unit of Hz .

R_t is the twisted-pair termination resistance.

r is the distance from the twisted-pair cable in the unit of m .

K_{cable} is the cable power loss in the unit of dB .

For $P = -70dBm/Hz$, $B_w = 2.2kHz$, $R_t = 100\Omega$, we have

$$\begin{aligned} |E| &= P - 30 + 10\log_{10} B_w R_t - K_{cable} + 120 - 10\log_{10}(4\pi) - 20\log_{10} r \\ &= -70 - 30 + 53.42 - K_{cable} + 120 - 10.99 - 20\log_{10} r \\ &= 62.43 - K_{cable} - 20\log_{10} r \end{aligned}$$

For $K_{cable} = 35dB$ and $r = 10m$, we have

$$|E| = 62.43 - 35 - 20\log_{10} 10 = 7.43dB\mu V/m$$

Table 15.4 shows estimated worst and average VDSL signal-caused field strength for different type of twisted-pair cables along with the worst-case measurements.¹⁶

Table 15.4 Field Strength Caused by a VDSL Signal (dB $\mu V/m$)

	BKMA 100 Pair at 10 m	ARTM 25 Pair at 6 m	Steel Flat Drop at 6 m	Twisted Drop at 6 m
Measured	1	-1	10	1
Worst Case	-5.53	-2.1	11.86	1.41
Average	-17.4	-13.9	3.65	-13.23

Measured values are also from the *Broadband Technologies T1E1.4 Contribution*. Estimations for worst VDSL-caused field strengths are very close to that of measured worst cases except for the BKMA 100 twisted pair. That big difference might be considered as a special case where the field strength is enhanced by an unknown cause. In comparison to twisted-pair cables, flat cables can induce a much stronger radio field.

15.2 General System Architecture

Before any specific VDSL transceiver parameters can be determined, some system issues need to be resolved. The understanding of the transmission potential under the VDSL

channel and noise condition is a very important aspect of defining the system architecture. The VDSL transmission throughput is mainly limited by the combination of loop plant background and self-NEXT noises.

A high transmit power can lead to a high transmission throughput. On the other hand, the transmission power level is limited not only by the capability of the hardware implementation, but also by the amount of egress from the twisted-pair loop. Using expressions developed in Chapter 4, “Transceiver Front-End Noise Models,” VDSL loop plant SNRs and channel capacities are calculated in this section.

The received signal power density is determined by the transmit power spectrum and the twisted-pair loop channel characteristics. We have

$$S(f) = Q(f)|H(f)|^2$$

$Q(f)$ is the transmit power spectral density, and $H(f)$ is the channel transfer function. The higher the received signal power, the higher the transmission throughput. The effectiveness of the received signal power, however, is relative to the noise power also present at the input of a receiver. For VDSL, we are interested in effects of FEXT, loop plant background, and RFI noise.

We consider the loop plant background noise as white noise because its spectrum is almost flat. The strength of a white noise of density W can be described by its total power σ^2 within a particular bandwidth of B . We have the received-signal-to-white-noise ratio as

$$\frac{S(f)}{W} = \frac{Q(f)|H(f)|^2}{W}$$

Figure 15.6 shows signal-to-noise ratios under the white noise condition for 24-gauge twisted-pair loop channel models, assuming a transmit power density of -70 dBm/Hz and a white noise power density of -140 dBm/Hz.

Under the self-FEXT condition for which the same transmit power spectrum also affects both received signal and noise power, we have the received-signal-to-noise ratio as follows:

$$\frac{S(f)}{N_F(f)} = \frac{Q(f)|H(f)|^2}{Q(f)FEXT(f)} = \frac{1}{klf^2}$$

The values are as follows:

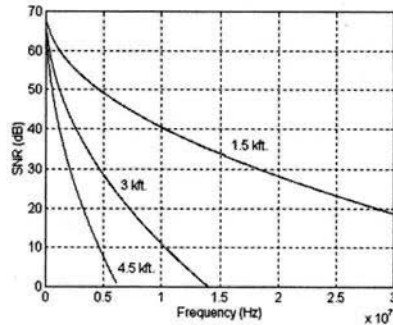
$$k = 8 \times 10^{-20}.$$

l is in units of ft.

f is in units of Hz.

The channel model becomes irrelevant.

Figure 15.6 Signal-to-White-Noise Ratio for 24-Gauge Twisted-Pair Loops



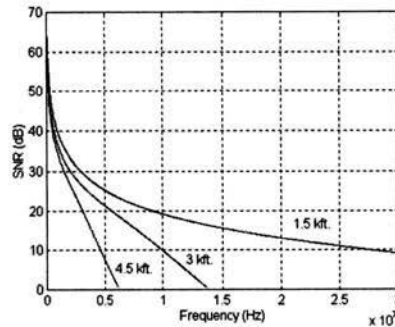
In practice, especially at a low-frequency region and a high-frequency region for which the FEXT noise power density is very small, the effect of white background noise also needs to be considered. By including the background noise power density, the received-signal-to-noise ratio becomes

$$\frac{S(f)}{N_F(f) + W} = \frac{Q(f)|H(f)|^2}{Q(f)FEXT(f) + W} = \frac{1}{klf^2 + \frac{W}{Q(f)|H(f)|^2}}$$

Figure 15.7 shows signal-to-noise ratios under the self-FEXT plus white noise condition for 24-gauge twisted-pair loop channel models with a transmit power density of -70 dBm/Hz and a background white noise level of -140 dBm/Hz.

For the receiver front-end white noise limited condition, the channel capacity for VDSL can be described as

$$C = \int_{f_1}^{f_2} \log_2 \left(1 + \frac{S(f)}{W(f)} \right) df$$

Figure 15.7 Signal-to-Self-FEXT Plus White Noise Ratio for 24-Gauge Twisted-Pair Loops

The values are as follows:

$f_1 = 1.5\text{ MHz}$ is the lower-corner frequency of the VDSL signal spectrum above that of ADSL.

f_2 is the higher-corner frequency.

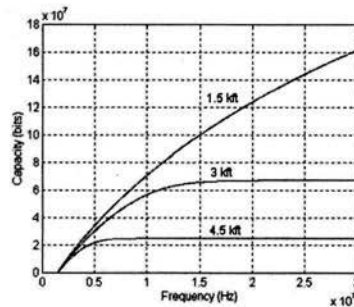
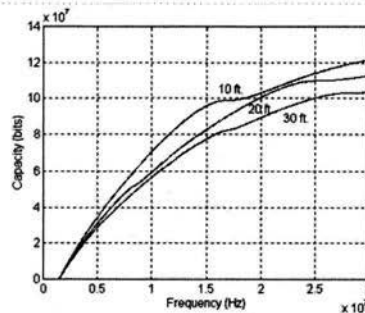
For the self-FEXT plus receiver front-end white noise limited condition, the channel capacity can be described as

$$C = \int_{f_1}^{f_2} \log_2 \left(1 + \frac{S(f)}{N_F(f) + W} \right) df$$

Figure 15.8 shows channel capacities under the self-FEXT plus white noise condition for 24-gauge twisted-pair loop channel models also assuming a transmit power density of -70 dBm/Hz and a white noise power density of -140 dBm/Hz.

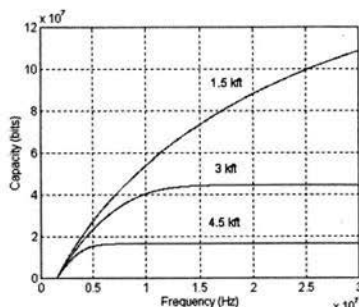
Figure 15.9 shows channel capacities also under the self-FEXT plus white noise condition for a 2 kft, 24-gauge twisted-pair loop with different lengths of bridged taps also assuming a transmit power density of -70 dBm/Hz and a white noise power density of -140 dBm/Hz.

According to channel capacity results, transmission throughputs for 1.5 kft, 3 kft, and 4.5 kft, 24-gauge twisted-pair loops are limited to 160 Mbps, 68 Mbps, and 25 Mbps, respectively, for a transmission bandwidth of less than 30 MHz. The frequency band below 1.5 MHz is also reserved for ADSL with concerns about spectral compatibility. We also observe that the inclusion of bridged taps will reduce the channel capacity. With advanced modulation and coding techniques, the transmission throughput of a DSL system can be made within 2 to 6 dB to less than the channel capacity.

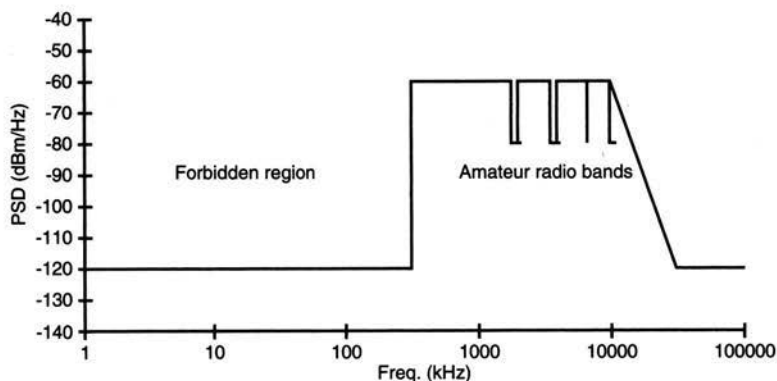
Figure 15.8 Self-FEXT Plus White Noise Channel Capacity for the 24-Gauge Twisted-Pair Loop**Figure 15.9** Self-FEXT Plus White Noise Channel Capacity for a 2 kft, 24-Gauge Twisted-Pair Loop

We might also reserve a 6 dB performance margin for VDSL to account for some non-characterized loop and system impairments. Figure 15.10 shows channel capacities under the same self-FEXT plus white noise condition for 24-gauge twisted-pair loop channel models except with a 6 dB performance margin. The effect of the 6 dB performance margin is emulated by reducing signal-to-noise ratio by a factor of 4 at all frequencies.

According to channel capacity results with 6 dB performance margin, transmission throughput for 1.5 kft, 3 kft, and 4.5 kft, 24-gauge twisted-pair loops is limited to 108 Mbps, 44 Mbps, and 18 Mbps, respectively, also for a transmission bandwidth of less than 30 MHz. The total of realized transmission throughput in both upstream and downstream directions might be less than what we have observed due to the allocation of a guardband between upstream and downstream channels if frequency division multiplex is involved.

Figure 15.10 Self-NEXT Channel Capacity with 6 dB Noise Margin

Symmetrical transmission throughput of 13 Mbps and 26 Mbps as well as asymmetrical transmission throughput of 52 Mbps/6.4 Mbps, 26 Mbps/3.2 Mbps, and 13 Mbp/1.6 Mbps have been considered for VDSL depending on loop and noise environment. Transmit power density of -60 dBm/Hz in non-HAM bands and -80 dBm/Hz in HAM bands has been considered. Figure 15.11 shows such a transmit power spectrum mask defined by the ETSI VDSL document.⁹ An alternative -70 dBm/Hz transmit power density was also considered.

Figure 15.11 VDSL Transmit Power Density with Notches

To take advantage of a better transmission environment at low frequencies, a staircase VDSL transmit power density mask has also been proposed, as shown in Figure 15.12.¹⁷ This proposed PSD mask has not been agreed to at any standard meetings.

Figure 15.12 A Staircase VDSL Transmit Power Density Mask

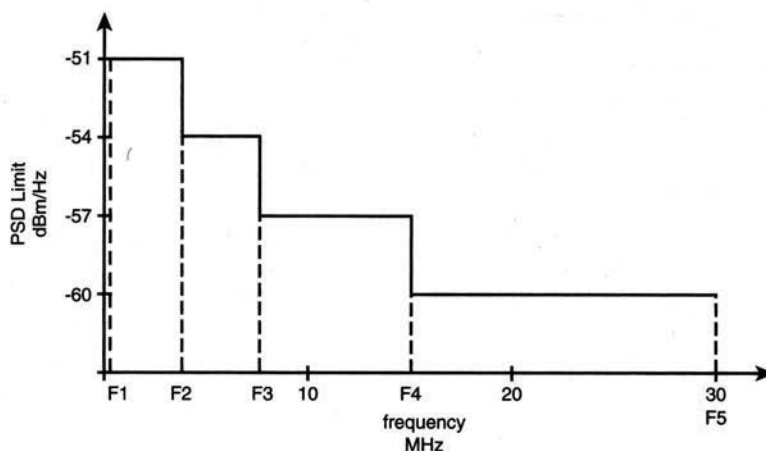


Table 15.5 defines the staircase VDSL PSD mask in more detail.

Table 15.5 The Definition of a Staircase VDSL PSD MASK

Frequency Band (MHz)	Power Density (dBm/Hz)
0.2–3.5	–51
3.5–7.0	–54
7.0–14	–57
14–30	–60
30–	–120

Table 15.6 specifies PSD at HAM bands for the same staircase VDSL mask.

Table 15.6 The Definition for HAM Bands

Frequency Band (MHz)	Power Density (dBm/Hz)
1.8–2.0	–71
3.5–4.0	–74
7.0–7.3	–77
10.1–10.15	–77
14–14.35	–80
18.068–18.168	–80
21.0–21.45	–80
28.0–29.7	–80

15.3 VDSL System Line Code Proposals

There are generally two proposals, CAP/QAM and DMT, for the VDSL line code. The CAP/QAM proposal is associated with the Frequency Division Multiplex for the upstream and downstream channels. The DMT proposal is associated with the Time Division Multiplex, although the division between upstream and downstream is independent from the choice of line code. In addition, there is also a multiple QAM VDSL proposal to avoid international amateur radio bands.¹⁸

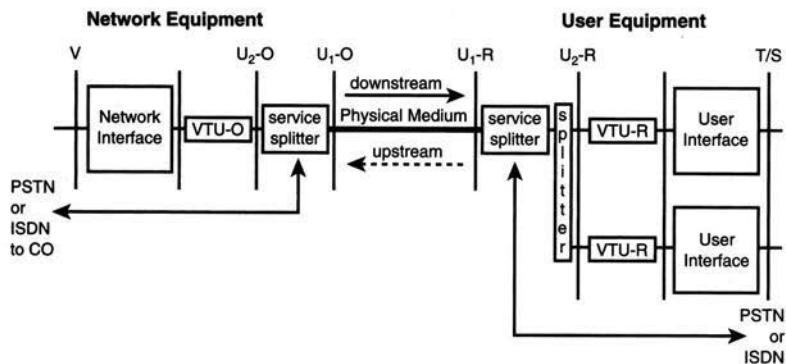
There are proposals to improve the performance of a DMT VDSL system using *window filters*.^{19,20} It was also considered possible to interlace downstream and upstream DMT subcarriers together for a FDM approach, but without any band splitting filter.²¹ It was shown that downstream and upstream frames can be aligned by extending the size of the cyclic prefix.

15.3.1 CAP/QAM Proposal

The CAP/QAM VDSL proposal is still under development. Currently, the transmission throughput, the constellation size, and the center frequency have been defined for different serving distance in both upstream and downstream directions.

Figure 15.13 shows the general architecture of a VDSL-based broadband transmission system.²²

Figure 15.13 VDSL Generic Reference Model



Service splitters, for either a POTS or an ISDN channel, terminate a VDSL subscriber loop at both ends. The VDSL at the network end, within an *Optical Network Unit* (ONU), is called *VDSL Terminal Unit-ONU* (VTU-O). The VDSL at the subscriber end within a residential premises is called *VDSL Terminal Unit-Remote* (VTU-R). Multiple deployment

of VTU-Rs was considered feasible with the use of a VDSL splitter. This is also sometimes called a *point-to-multipoint configuration*.

Figure 15.14 shows a VDSL protocol reference model, which further depicts the existence of a frame structure, separate upstream and downstream error control mechanisms, and some management and operational control functions.

Figure 15.14 VDSL Protocol Reference Model

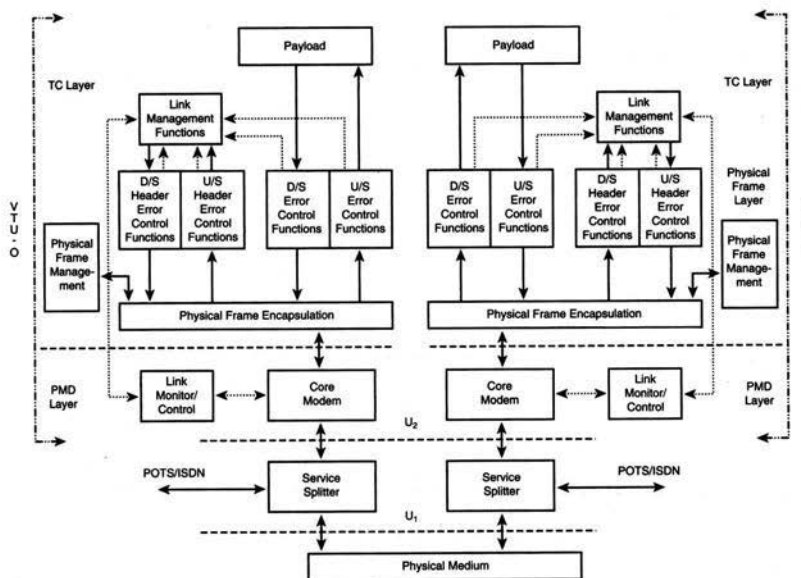


Figure 15.15 shows the general structure of a CAP VDSL transmitter. The CAP VDSL transmitter consists of a bit-to-symbol encoder, a pair of passband, in-phase and quadrature, transmit shaping filters, and an additional passband spectrum shaping filter.

Figure 15.16 shows the general structure of a QAM VDSL transmitter. The QAM VDSL transmitter also consists of a bit-to-symbol encoder, a pair of baseband transmit shaping filters followed by a pair of modulators implemented with the in-phase and quadrature multiplication of the carrier, and an additional passband spectrum shaping filter.

The impulse response of the baseband shaping filter for the QAM transmitter can be described as

$$g(t) = \frac{\sin\left(\pi \frac{4t}{5T}\right) + \left(\frac{4t}{5T}\right) \cos\left(\pi \frac{6t}{5T}\right)}{\left(\pi \frac{t}{T}\right) \left[1 - \left(\frac{4t}{5T}\right)^2\right]}$$

T is the baud period.

Figure 15.15 CAP VDSL Transmitter

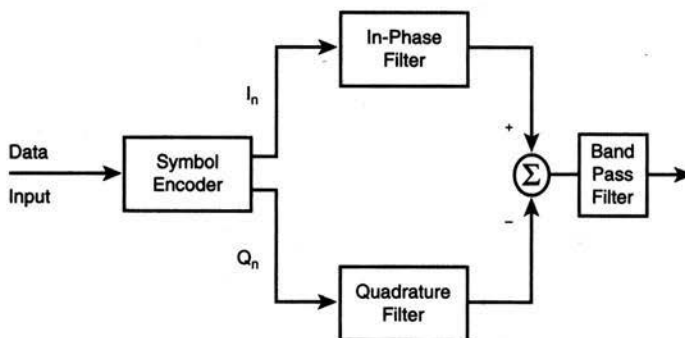
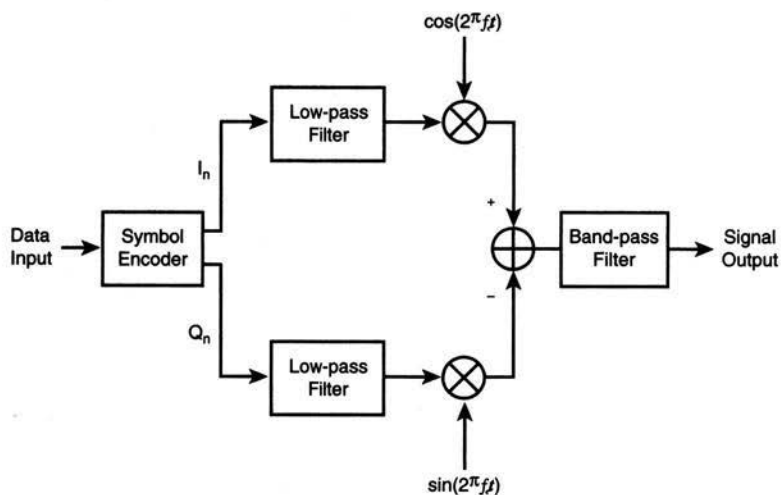


Figure 15.16 QAM VDSL Transmitter



The QAM transmit signal can be described as

$$S_{QAM}(t) = \left[\sum_n I_n g(t - nT) \right] \cos(2\pi f_c t) - \left[\sum_n Q_n g(t - nT) \right] \sin(2\pi f_c t)$$

The impulse response of the in-phase CAP transmit filter can be described as

$$f(t) = g(t) \cos(2\pi f_c t)$$

The value f_c is the center frequency.

The impulse response of the quadrature CAP transmit filter can be described as

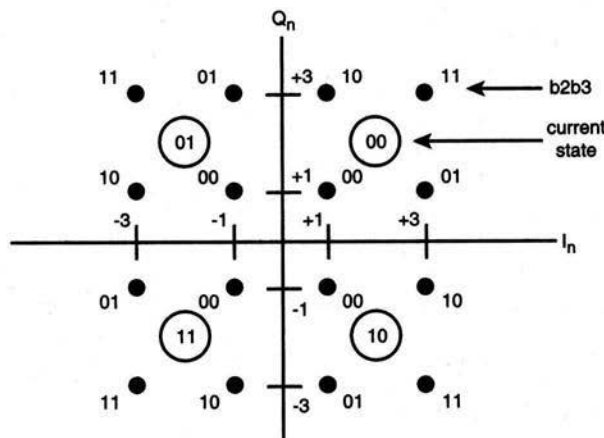
$$f'(t) = g(t) \sin(2\pi f_c t)$$

The CAP transmit signal can be described as

$$S_{CAP}(t) = \sum [I_n f(t - nT) - Q_n f'(t - nT)]$$

A 16-point constellation carrying four bits per symbol is used for both QAM and CAP line codes in the downstream direction, as illustrated in Figure 15.17.

Figure 15.17 CAP/QAM Downstream 16-Point Constellation



Even though the same constellation is adapted, VDSL CAP and QAM transceivers are generally not compatible, which means that a VDSL QAM receiver cannot be used to recover signals from a VDSL CAP transmitter and vice versa.

Table 15.7 summarizes the CAP/QAM VDSL downstream transmission throughput values, baud rates, and transmit spectrum characteristic frequencies.

Table 15.7 Downstream Transmission Parameters

Loop Length kft (km)	Downstream Rate (Mbps)	Symbol Rate (MHz)	Lowest Freq. (MHz)	Center Freq. (MHz)	Highest Freq. (MHz)
Long, 4.5 (1.5)	6.48	1.62	2	2.972	3.944
Long, 4.5 (1.5)	12.96	3.24	2	3.944	5.888
Medium, 3 (1)	25.92	6.48	2	5.888	9.772
Short, 1 (0.3)	51.84	12.96	2	9.776	17.552

The downstream baud rates of a VDSL transceiver can range from 1.62 MHz to 12.96 MHz. Transmit spectrum bandwidths defined by differences between highest and lowest frequencies are 120 percent of corresponding baud rates. Different transmit filter coefficients should be maintained for different loop conditions. The transmit spectrum of the VDSL downstream channel is allocated at above 2 MHz.

A 256-point constellation carrying eight bits per symbol is used for both QAM and CAP line codes in the upstream direction (see Figure 15.18). Again, VDSL CAP and QAM upstream transceivers are not compatible.

Table 15.8 summarizes CAP/QAM VDSL upstream transmission throughput values, baud rates, and transmit spectrum characteristic frequencies. The upstream baud rates of a VDSL transceiver can range from 202.5 kHz to 810 kHz. The transmit spectrum of the VDSL upstream channel is centered at 1.314 MHz.

Figure 15.18 CAP/QAM Upstream 256-Point Constellation

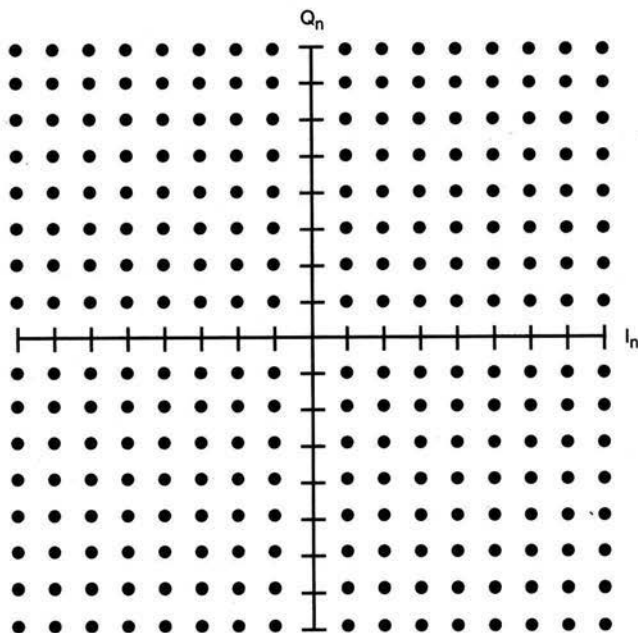


Table 15.8 Upstream Transmission Throughput

Loop Length kft (km)	Downstream Rate (Mbps)	Baud Rate (MHz)	Lowest Freq. (MHz)	Center Freq. (MHz)	Highest Freq. (MHz)
Long, 4.5 (1.5)	1.62	0.2025	1.1925	1.314	1.4355
Medium, 3 (1)	3.24	0.405	1.071	1.314	1.557
Short, 1 (0.3)	6.48	0.81	0.828	1.314	1.8

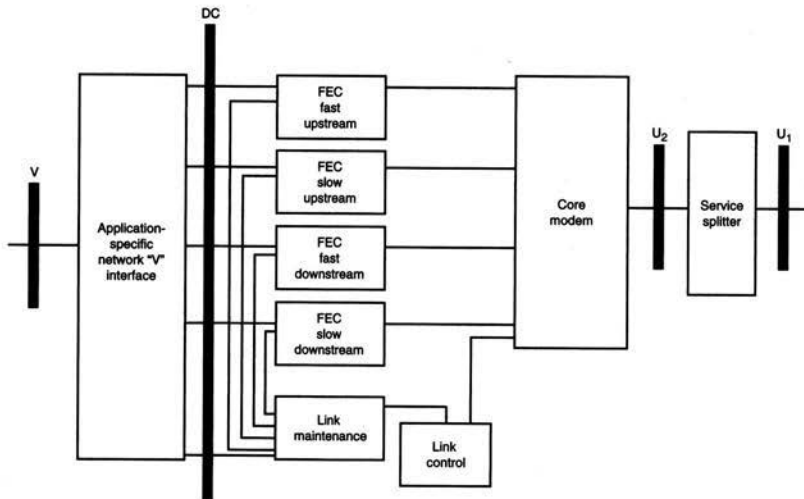
Symmetric transmission throughput values of 12.96 Mbps and 25.92 Mbps are considered possible for medium- and short-length VDSL loops, respectively. Detailed constellation and spectrum parameters, however, have not yet been decided.

15.3.2 DMT Proposal

The DMT VDSL proposal is also under further development. Currently, the IFFT/FFT size, the transmission throughput, the transmit power spectrum, and the super frame structure have been defined for different serving distances in both upstream and downstream directions.

Figure 15.19 illustrates the reference model of a DMT VDSL²³ transceiver, VTU-O. In addition to the core mode, physical layer of the DMT VDSL, Forward Error Correction (FEC)–based error control mechanisms are also depicted. Dual latency systems are defined for both downstream and upstream channels.

Figure 15.19 Reference Model



The reference model of a VTU-R is similar to that depicted in Figure 15.19, except that the name of α interface is replaced by the name β interface, and the V interface connecting to the backbone network is replaced by the S/T interface connecting to the premises wiring.

Table 15.9 summarizes the transmission parameters of the DMT VDSL. Three possible sampling rates (5, 10, and 20 times that of ADSL) are considered while IFFT/FFT size is maintained to be the same. Up to 11 bits can be allocated for each DMT subcarrier depending on the line-signal-to-noise ratio. A bit loading algorithm defining the constellation and corresponding bit mapping similar to that used for ADSL is also proposed.

Table 15.9 DMT VDSL Transmission Parameters

Sampling Rate	11.04 MHz	22.08 MHz	44.16 MHz
Baud Rate	20 kHz	40 kHz	80 kHz
Bandwidth	5.02 MHz	11.04 MHz	22.08 MHz
Cyclic Prefix Size	40 Samples	40 Samples	40 Samples
IFFT/FFT Size	256	256	256
Number of Tones	256	256	256
Tone Bandwidth	21.5625 kHz	43.125 kHz	87.25 kHz

Utilizing Time Division Multiplex (TDM), the transmit power spectrum mask shown in Figure 15.20 is defined for both upstream and downstream directions. The transmit power density level is -60 dBm/Hz with an allowed variation of ± 3 dB to realize the full potential of each subcarrier without the use of a fractional bit loading algorithm.

Figure 15.20 DMT VDSL Transmit Power Spectrum

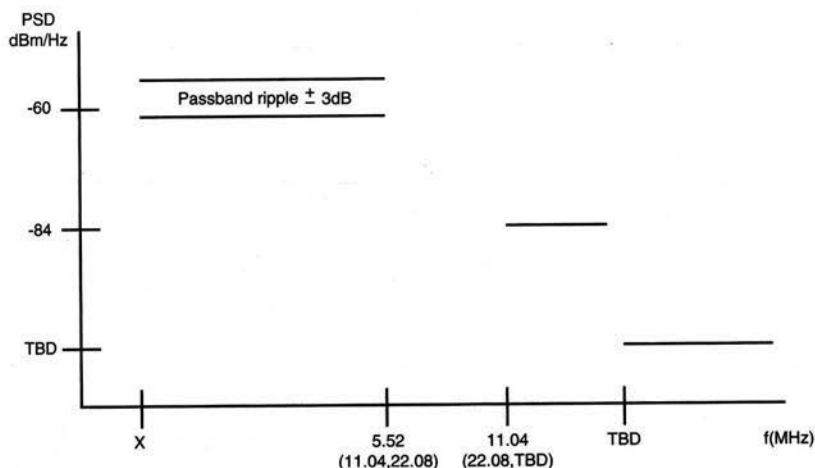
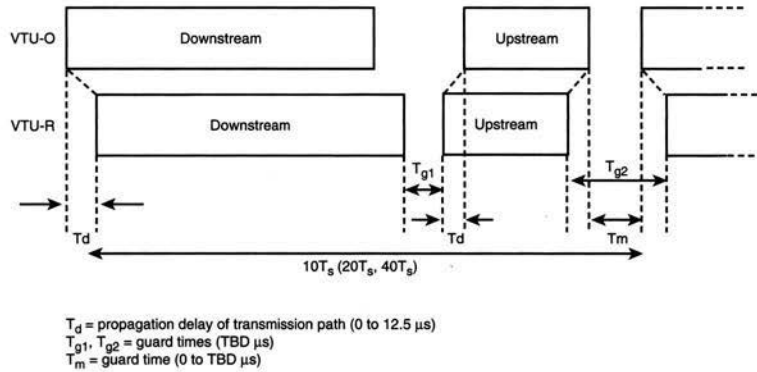


Figure 15.21 shows the timing relationship of downstream transmission, upstream transmission, propagation delay, and guard time.

A guard time period between the reception of the incoming packet and the transmission of the next outgoing packet is necessary to account for different round trip delays among all VDSL loops because the synchronization between all VDSL transceivers is required to avoid the effect of NEXT. There are two guard time periods, as shown in Figure 15.21. The total guard time ($T_{g1} + T_{g2}$) is a multiple of DMT symbols.

Figure 15.21 The DMT VDSL Time Division Multiplex System

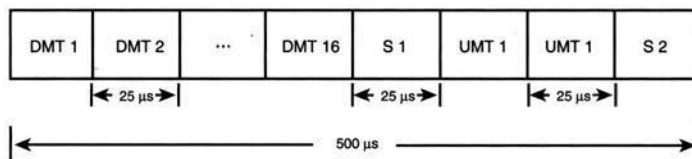


The transmission throughput ratio between downstream and upstream is regulated through the allocation of a different number of DMT data symbols for different directions. The transmission throughput is in multiples of 64 Kbps and is fully user programmable.

Asymmetric DMT VDSL Super Frames

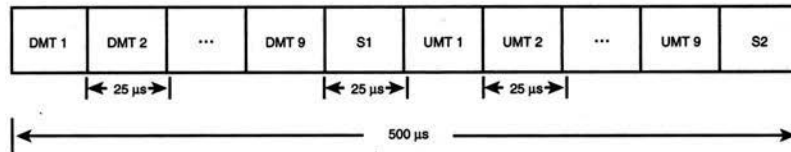
Figure 15.22 shows an asymmetric super frame with a downstream-to-upstream transmission throughput ratio of 8:1. 16 DMT symbols are allocated for the downstream channel (DMT), while only two DMT symbols are allocated for the upstream channel (UMT). One silent symbol is used following the group of downstream symbols and the group of upstream symbols.

Figure 15.22 An Asymmetric VDSL Super Frame



Symmetric VDSL Super Frames

Figure 15.23 shows a symmetric super frame. Nine DMT symbols are allocated for the downstream channel, and nine DMT symbols are allocated also for the upstream channel. One silent symbol is used also following the group of downstream symbols and the group of upstream symbols.

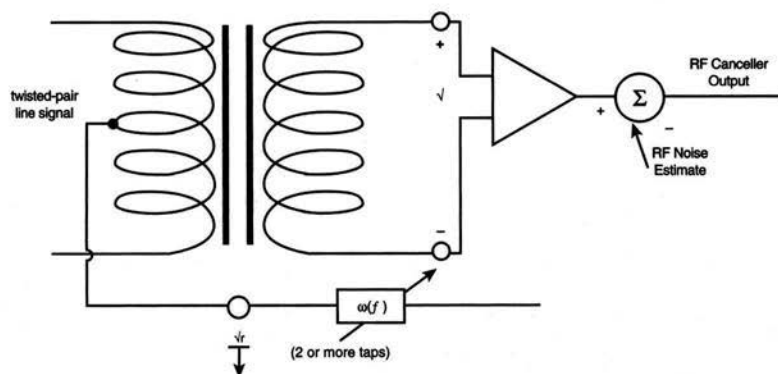
Figure 15.23 A Symmetric VDSL Super Frame

The mixing of asymmetric and symmetric transmission throughput services in the same VDSL is believed to be feasible. Under the all-VDSL transceiver synchronized condition, the implementation of mixed services will induce some extra NEXT noise for VDSL receivers located at the ONU side.

Minimizing RFI Noise

The effect of RFI can be minimized with some noise cancellation techniques.²⁴ A combination of analog and digital RFI cancellation as well as subcarrier nullifying techniques has been proposed for the DMT VDSL.²⁵

Figure 15.24 shows the general arrangement of an analog RFI canceller. The weight can be trained using adaptive signal processing algorithms during the silent periods. The weight can also be simplified to a frequency-independent scaling factor.

Figure 15.24 Analog Radio Interference Cancellation

15.4 Twisted-Pair Primary Constants for VDSL

Twisted-pair cable primary parameters presented by two Bellcore contributions^{2,3} are summarized in Tables 15.10 through 15.18 for a quick reference. The frequency range is up to 30 or 40 MHz. More frequency points are available from these contributions.

Table 15.10 0.4 mm (24-Gauge) Twisted Pair

Frequency (kHz)	R (ohm/km)	L (μH/km)	G (nMho/km)	C (μF/km)
5	180.98	724.63	74.72	130.66
10	181.14	721.82	72.89	236.51
15	181.40	719.42	71.88	334.65
20	181.76	717.27	71.19	428.10
30	182.79	713.46	70.26	605.74
50	185.96	707.05	69.15	938.00
70	190.44	701.66	68.46	1251.12
100	199.02	694.78	67.75	1697.87
150	216.34	685.43	66.97	2402.43
200	235.21	677.78	66.45	3073.31
300	272.90	665.70	65.74	4348.62
500	340.88	648.79	64.89	6733.92
700	399.36	637.08	64.35	8981.78
1000	474.74	624.65	63.81	12189.01
1500	579.72	611.03	63.22	17247.00
2000	668.71	601.99	62.81	22063.22
3000	818.39	590.45	62.27	31218.65
5000	1056.13	578.22	61.62	48342.67
7000	1249.50	571.59	61.21	64480.06
10000	1493.35	565.74	60.79	87504.68
15000	1828.92	560.39	60.34	123815.92
20000	2111.83	557.33	60.03	158391.52
30000	2586.43	553.87	59.61	224118.21

Table 15.11 0.5 mm (26-Gauge) Twisted Pair

Frequency (kHz)	R (ohm/km)	L (μ H/km)	G (nMho/km)	C (μ F/km)
5	174.62	616.69	50	29.88
10	174.81	615.96	50	77.77
15	175.12	615.17	50	136.10
20	175.55	614.35	50	202.42
30	176.76	612.66	50	354.21
50	180.49	609.16	50	716.83
70	185.67	605.63	50	1140.44
100	195.45	600.42	50	1865.68
150	214.64	592.18	50	3264.69
200	235.03	584.64	50	4855.77
300	274.83	571.60	50	8496.96
500	345.18	552.20	50	17195.51
700	405.11	538.85	50	27357.20
1000	482.06	525.44	50	44754.46
1500	588.99	512.26	50	78314.52
2000	679.53	504.61	50	116481.71
3000	831.75	496.23	50	203827.92
5000	1073.45	489.11	50	412491.45
7000	1270.01	486.02	50	656253.52
10000	1517.88	483.72	50	1073584.8
15000	1858.97	481.99	50	1878634.5
20000	2146.54	481.15	50	2794201.6
30000	2628.95	480.34	50	4889491.3

Table 15.12 DW10 Reinforced 0.5 mm Twisted Pair

Frequency (kHz)	R (ohm/km)	L (μ H/km)	G (nMho/km)	C (μ F/km)
5	286.22	673.73	49	16.70
10	286.33	672.27	49	27.13
15	286.53	670.88	49	36.04
20	286.80	669.55	49	44.07
30	287.58	667.01	49	58.54

Frequency (kHz)	R (ohm/km)	L (μ H/km)	G (nMho/km)	C (μ F/km)
50	290.04	662.29	49	83.70
70	293.60	657.93	49	105.93
100	300.77	651.94	49	135.98
150	316.47	643.08	49	180.60
200	335.14	635.31	49	220.90
300	376.06	622.20	49	293.40
500	457.03	602.52	49	419.51
700	530.29	588.28	49	530.93
1000	626.85	572.87	49	681.50
1500	763.06	556.00	49	905.17
2000	879.19	545.00	49	1107.11
3000	1075.10	531.41	49	1470.46
5000	1386.83	517.85	49	2102.56
7000	1640.56	511.01	49	2660.95
10000	1960.61	505.33	49	3415.61
15000	2401.10	500.51	49	4536.62
20000	2772.49	497.93	49	5548.68
30000	3395.54	495.20	49	7369.76

Table 15.13 DW8 1.14 mm Flat Telephone Cable

Frequency (kHz)	R (ohm/km)	L (μ H/km)	G (nMho/km)	C (μ F/km)
5	41.26	998.74	35.04	95.36
10	41.59	997.17	34.13	175.50
15	42.11	995.49	33.62	250.75
20	42.81	993.76	33.28	322.98
30	44.63	990.30	32.81	461.47
50	49.32	983.63	32.26	723.38
70	54.52	977.56	31.91	972.66
100	62.28	969.67	31.55	1331.30
150	74.18	959.34	31.16	1902.11
200	84.75	951.67	30.89	2450.09
300	102.99	941.30	30.53	3500.60

continues

Table 15.13 Continued

Frequency (kHz)	R (ohm/km)	L (μ H/km)	G (nMho/km)	C (μ F/km)
500	132.41	930.35	30.10	5487.44
700	156.49	924.83	29.83	7378.40
1000	186.92	920.41	29.55	10098.94
1500	228.86	916.87	29.25	14429.00
2000	264.23	915.12	29.04	18585.84
3000	323.59	913.40	28.76	26554.77
5000	417.74	912.10	28.43	41626.46
7000	494.27	911.58	28.22	55970.88
10000	590.76	911.21	28.00	76608.31
15000	723.53	910.94	27.77	109455.14
20000	835.46	910.81	27.61	140988.04
30000	1023.22	910.70	27.39	201438.54

Table 15.14 22-Gauge PIC Cable by Bellcore

Frequency (kHz)	R (ohm/km)	L (μ H/km)	G (nMho/km)	C (μ F/km)
0.3	197.30	0.537	96	50.27
0.5	236.91	0.528	161	50.94
0.7	289.54	0.518	226	51.41
1	357.25	0.506	323	51.61
1.5	422.08	0.494	485	51.80
2	494.48	0.488	653	51.89
3	604.82	0.481	977	51.92
5	773.35	0.474	1610	51.85
7	914.91	0.471	2265	51.75
10	1097.92	0.469	3262	51.63
15	1332.03	0.467	4809	51.50
20	1555.59	0.466	6441	51.38
30	1918.18	0.465	9624	51.27
40	2278.32	0.464	12866	51.19

Table 15.15 24-Gauge PIC Cable by Bellcore

Frequency (kHz)	R (ohm/km)	L (μ H/km)	G (nMho/km)	C (μ F/km)
0.3	269.87	0.581	98	51.30
0.5	335.32	0.559	162	51.20
0.7	395.73	0.544	226	51.47
1	465.11	0.528	324	51.70
1.5	557.87	0.514	486	51.84
2	640.57	0.506	653	51.91
3	776.39	0.497	977	51.90
5	990.41	0.489	1607	51.76
7	1173.18	0.485	2260	51.62
10	1407.60	0.482	3252	51.46
15	1712.96	0.480	4790	51.30
20	1998.94	0.479	6411	51.15
30	2484.74	0.477	9573	50.99
40	2935.00	0.477	12792	50.90

Table 15.16 26-Gauge PIC Cable by Bellcore

Frequency (kHz)	R (ohm/km)	L (μ H/km)	G (nMho/km)	C (μ •F/km)
0.3	353.22	0.579	94	49.55
0.5	408.00	0.573	159	50.71
0.7	492.54	0.567	224	50.96
1	613.31	0.554	318	50.70
1.5	728.78	0.540	476	50.55
2	848.40	0.530	636	50.54
3	1019.26	0.518	953	50.64
5	1288.27	0.505	1580	50.87
7	1520.23	0.497	2235	51.04
10	1809.56	0.489	3238	51.25
15	2176.92	0.482	4808	51.50
20	2516.51	0.477	6474	51.64
30	3093.75	0.471	9737	51.87
40	3606.84	0.68	13077	52.03

Table 15.17 Two-Pair Twisted Drop Cable by Bellcore

Frequency (MHz)	R (ohm/kft)	L (mH/kft)	C (nF/kft)
0.772	112.64	0.1436	14.16
0.978	128.30	0.1400	13.89
1.481	163.02	0.1377	13.79
1.991	194.34	0.1384	13.94
3.013	250.16	0.1374	13.94
5.135	349.72	0.1364	13.93
6.905	426.53	0.1359	13.92
9.850	539.99	0.1353	13.90
14.91	720.14	0.1345	13.86
20.05	890.36	0.1338	13.82
30.34	1191.67	0.1328	13.74
40	1488.43	0.1321	13.68

Table 15.18 Flat-Pair Drop Cable by Bellcore

Frequency (MHz)	R (ohm/kft)	L (mH/kft)	C (nF/kft)
0.772	170.5	0.129	10.5
0.999	190.1	0.134	10.34
1.497	253.6	0.144	10.37
1.993	324.2	0.151	10.43
2.987	462	0.159	10.51
4.987	702.9	0.167	10.55
6.98	900.9	0.172	10.56
10.05	1153	0.175	10.56
15.06	1478	0.179	10.55
20.05	1738	0.181	10.54
30.06	2141	0.183	10.53
40	2455	0.184	10.5

End Notes

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Glossary

Numbers/Symbols

2B1Q line code A baseband line code with two bits per quaternary, four levels, symbols.

3B2T line code A baseband line code with three bits for every two ternary, three levels, symbols.

4B3T line code A baseband line code with four bits for every three ternary, three levels, symbols.

A

ABCD parameters A two-by-two matrix describing input voltage and current-to-output voltage and the current relationship of a two-port network.

ADC (Analog to Digital Converter) An analog integrated circuit that samples, quantizes, and holds analog signals for digital signal processing.

AFE (analog front end) The part that connects digital signal processing circuits to an analog signal. AFE mainly consists of an ADC and a DAC.

AGC (Automatic Gain Control) A part of the AFE that adjusts the received signal level to that suitable for the input of the ADC.

AMI (Alternate Mark Inversion) line code A simple baseband line code that alternates the parity of data symbols for adjacent 1s. For example, a sequence of 1011000 is encoded as +A 0 -A +A 0 0 0 for the AMI line code for which A is the magnitude of the line code.

Amplifier nonlinearity The nonlinear amplifier input-to-output relationship.

ASIC (Application Specific Integrated Circuit) An integrated circuit (IC) that is designed for a specific application in contrast to some general purpose ICs that are available from an IC catalog.

ASSP (Algorithm Specific Signal Processor) An integrated circuit (IC) that is designed for a specific algorithm for a category of general applications. An example is an adaptive filter IC.

Asymptotic coding gain A coding gain that can only be obtained when the SNR value approaches infinity.

Attenuation The signal loss caused by a transmission channel at a particular frequency.

AWG (American Wire Gauge) A measuring system designed to distinguish twisted-pair telephone cables with different diameters of conductors.

B

B8ZS (bipolar with 8-zero substitution) line code A baseband line code in which ones and zeros are encoded with opposite parities and every consecutive of eight zeros is replaced with a special non-all-zero sequence.

Background noise A random noise in front of, and external to, a receiver.

Balance condition The condition in which a hybrid circuit is most effective at separating transmitted and received signals.

Bandwidth The frequency band size of a signal.

Baseband signals Data symbols that are sent to the channel without modulation.

Baud rate equalizers Adaptive channel equalizers operating at the symbol of baud rate.

BCH (Bose, Chaudhuri, and Hocquenghem) code A class of cyclic codes that is easy to define and implement.

BER (bit error rate) A performance measurement, usually in terms of the number of errors per second, of a transmission system.

Bessel filter A filter whose frequency response is described by Bessel functions.

Biphase line code A baseband line code with different phases for zeros and ones. *See also* Manchester line code.

Bit stuffing Inserting an extra number of bits to adjust the frame size.

Brickwall filter An ideal filter frequency response with 0 loss passband, infinite loss stop-band, and no transition region in between.

Bridged taps Unused twisted-pair cables attached to a telephone subscriber loop.

Byte A group of eight bits.

C

CAP (Carrierless AM/PM) system A variation of the passband QAM line code.

Capacitance The capacity of electrical energy storage in positive and negative charges.

Channel capacity The maximum transmission throughput of a channel with a correctable number of transmission errors.

Channel intersymbol interference Interference between adjacent data symbols caused by channel frequency and phase distortion.

Clipping noise Noise caused by the clipping of signal peaks by DAC or ADC.

CMOS (Complementary Metal Oxide Semiconductor) A semiconductor technology whose element circuit is based on the complementary combination of *p* and *n* types of semiconductor material.

CO (Central Office) A telephone company office for terminating telephone subscriber loops and for housing telephone switching equipment.

Code book A mapping between input information sequences and output code words.

Companding A method to compress a large amplitude signal to increase the dynamic range of a voice encoder through non-uniform quantization.

Conductance The capacity of electrical energy storage in the strength electromagnetic fields.

Convolution code A coding mechanism whose output codewords depend on input information sequences as well as the state of the encoder.

Crosstalk noise Noise caused by a signal carried over adjacent telephone subscriber loops.

CSA (Carrier Serving Area) A design guideline for the constraining of the length of telephone subscriber loops to accommodate digital telephone transmission equipment.

Cyclic code An error correction code. The cyclic shift of a codeword from a particular cyclic code is also a codeword of the same cyclic code.

D

DA (Distribution Area) The area served by a remote terminal.

DAC (Digital to Analog Converter) An analog integrated circuit that takes in a group of binary digits and generates an analog signal of corresponding magnitude.

dBm The Log format power ratio with respect to 1 nano-watt including a C-message filter.

DC resistance The resistance at direct current or zero frequency.

DFE (Decision Feedback Equalizer) A channel equalizer consisting of a feedforward filter (FFF) whose input is the received signal and a feedback filter whose input is the sequence of estimated data symbols.

DFT (Discrete Fourier Transform) The discrete version of the Fourier transform. It is usually implemented with the FFT algorithm using digital hardware.

Differential channel capacity The channel capacity within a unit frequency. The differential channel capacity is derivative of the conventional channel capacity. The differential channel capacity can be used to examine the frequency characteristic aspect part of a channel.

Differential hybrid circuit A hybrid circuit that connects to ADC and DAC with differential input and output, respectively.

Discrete time domain Signal values that are defined at sampling instances of periodic intervals.

Distribution cable A part of the telephone subscriber loop that connects feeder cables to drop wires.

Disturbers Signal sources of crosstalk noises.

DLC (Digital Loop Carrier) system A high-throughput digital transmission system developed to connect a Central Office to a remote terminal. A DLC system can be implemented using existing twisted-pair telephone subscriber loops or optical fiber cables.

DMT (Discrete MultiTone) systems A multi-carrier line code whose modulation and demodulation are implemented using Inverse Fast Fourier Transform (IFFT) and Fast Fourier Transform (FFT), respectively. The twisted-pair channel inter-symbol interference of a DMT system is mitigated by dividing the transmission spectrum into many sub-channels.

double-Barker code A synchronization word whose contents are the double bit assignment of a Barker code. The double-Barker code is defined for the HDSL frame structure.

Drop wire A telephone cable (twisted or flat) that connects a distribution cable to a subscriber premises.

DS1 (Digital Signal 1) circuit The first level of the North American digital hierarchy with a nominal rate of 1.544 Mbps. A specific implementation of DS1 is further referred to as DSX-1.

DSL (Digital Subscriber Lines) The original name for the physical layer of the Basic Rate Access ISDN channel. DSL uses a 2B1Q baseband line code with a transmission throughput of 160 kbps. Most recently, the term DSL is used as a generic name for any Digital Subscriber Loop system (ADSL, HDSL, and so forth).

DSP (Digital Signal Processor) A programmable semiconductor circuit capable of implementing multiple signal processing algorithms with software

DTMF (dual tone multifrequency) A method used for telephone touch dialing.

E

EC (echo cancellation) Separation of the transmit signal from the received signal through the use of an adaptive filter.

Echo canceller The digital signal processing circuit implementing the echo cancellation algorithm.

Echo path impulse response The time-domain impulse response of the echo path (from transmit to receive).

Echo path transfer function The frequency domain description of the echo path.

Echo return loss The average signal loss of the echo path.

Eigenvalues The characteristic vector of a matrix. The eigenvalues are obtained by solving a set of linear equations derived from the matrix.

EOC (Embedded Operations Channel) A virtual communication channel embedded in the physical layer for inter-transceiver and network maintenance purposes.

Euclidean distance The geometric distance in a multidimensional signal space.

exclusive-or operation The binary operation that returns a 1 only when one of its two operands is 1.

F

FBF (FeedBack Filter) The feedback filter of a DFE. An FBF cancels the tail portion of the channel impulse response using recovered data symbols. An FBF can be implemented only at the baud rate.

FDD (Frequency Division Duplex) systems Transmission channels in opposite directions are separated by frequency division.

FDI (Feeder Distribution Interface) A twisted-pair telephone loop plant interface point between feeder and distribution cables.

FDM (Frequency Division Multiplex) transmissions Transmission channels separated by frequency division.

Feeder cables Twisted-pair telephone cables coming out of COs.

FEXT (Far End Crosstalk) noise The crosstalk noise from the other end of the twisted-pair telephone loop.

FFF (FeedForward Filter) The feedforward filter of a DFE. An FFF compensates for the channel distortion with linear equalization. An FFF can be implemented at a multiple of the baud rate.

FFT (Fast Fourier Transform) An efficient digital implementation of the Fourier transform, which converts a time-domain signal to its frequency-domain equivalence.

FIR (Finite Impulse Response) filter A digital filter with a finite number of delay and multiplication elements. The output of an FIR filter is the convolution of the input and filter coefficients.

Fourier transform A linear transform based on the sinusoid expansion of a signal.

Fractionally spaced equalizer A channel equalizer operating on multiples of the signaling rate.

G

GAFF (General Adaptive FIR Filter) chip A semi-programmable semiconductor circuit for the implementation of DSL transceivers.

Galois field A closed algebra field for describing encoding and decoding operations.

Gaussian distribution A bell-shaped statistical distribution function. Also called the *normal distribution*.

Guard period A period used in a DMT system to elevate the effect of channel distortion.

H-I

Hamming distance The minimum number of bits that are different between every pair of codewords.

IDFT (Inverse Discrete Fourier Transform) The discrete version of the inverse Fourier transform. It is usually implemented with the IFFT algorithm using digital hardware.

IFFT (Inverse Fast Fourier Transform) An efficient digital implementation of the Inverse Fourier transform that converts a frequency-domain signal to its time-domain equivalence.

Impedance The relationship between the applied voltage and the induced current of an electrical device.

Impulse noise Short-duration, high-magnitude noise whose occurrence does not follow the normal distribution.

Impulse response The time-domain response at the output of a network corresponding to an impulse applied at the input of a network.

INA (Intelligent Network Architecture) A general telephone networking concept that coordinates trunk, network, service, and management capabilities.

Inductance The measurement of the amount of energy stored in the magnetic field format.

IPC (ISDN to POTS Converter) A device that converts an ISDN interface to a POTS interface.

ISDN (Integrated Service Digital Network) A digital network with integrated services, switch, and maintenance functions.

ISI (Inter Symbol Interference) Interference between adjacent data symbols caused by channel distortion.

L

LC ladder An analog filter structure where inductance and capacitance are connected in an alternate cascading fashion.

LEC (Linear Echo Canceller) An echo canceller dealing only with a linear echo path.

Linear channel equalizer A channel equalizer with only received signals as its input.

LMS (Least Mean Square) algorithm Algorithm commonly used for the adaptive filters for its simplicity and robustness.

Load coils Inductance coils used in telephone loops to improve the transmission performance of the voice band channel.

Look-up table A map of the input-output logic relationship stored in memory or described by a table instead of a logic expression.

Loopback A direct connection between the transmit path and the receive path. Loopback can be used for testing purposes. Loopback is also used to describe the direct connection of the transmit clock to the clock recovered from the received signal.

LSB (Least Significant Bit) The bit that determines the presence of the zero power of two terms in a binary polynomial representation.

LT (line termination) The Central Office-side DSL transceiver defined for ISDN Basic Rate Access channel.

M

Magnitude The amplitude of a signal.

Manchester line code A line code that maps a data bit 1 into a pulse containing both a positive and a negative level and a data bit 0 into the same pulse with inverted polarity.

MDB (Modified Duobinary) line code A baseband line code in which each data symbol is followed by its negative replica after a two-symbol delay.

MMSE (Minimum Mean Square Error) An optimization objective to minimize the average power of error.

MSB (Most Significant Bit) The bit that determines the presence of the highest power of two terms in a binary polynomial representation.

N

NEC (Nonlinear Echo Canceller) An echo canceller capable of dealing with an echo path with a certain degree of nonlinearity.

NEXT (Near End Crosstalk) noise Crosstalk noise coming from near-end transmitters.

NID (Network Interface Device) A telephone company supplied module for interconnection between the telephone subscriber loop and in-house wiring at the loop entry point.

NML (Near Maximum Likelihood) algorithm A receiver algorithm whose performance is close to that of a maximum likelihood algorithm, but with a simplified implementation.

Noise Random signal with no detectable information.

Noise floor The background noise level across a frequency band.

Nonlinearity The input/output value deviations for Digital to Analog Conversions (DAC) and Analog to Digital Conversions (ADC).

NT (network termination) The subscriber-side DSL transceiver defined for an ISDN Basic Rate Access channel.

Nyquist shaping filter A digital FIR filter that introduces no intersymbol interferences at particular periodic sampling points.

O-P

OAM (Operation, Administration, Maintenance, and Provisioning) channels An embedded utility channel for carrying out auxiliary functions.

PAR (peak-to-average voltage ratio) The measure for the dynamic range of a line signal.

Parity check code An encoding method for which the coding redundancy is the parity check bits of message bits.

Passband signal A line code that is modulated with a carrier of cosine and/or sine waves.

PBX (Private Branch Exchanges) A telephone switch for private business applications.

PCM (Pulse Code Modulation) A binary code word representation of a baseband multi-level signal.

PDM (Pulse Duration Modulation) A pulse width representation of a baseband multi-level signal. In other words, the magnitude of a signal is translated into the width of a fixed-height pulse.

Peak-to-peak voltage The voltage level measured from the negative peak to the positive peak.

Phase The starting angle of a sinusoidal signal.

Phase delay The starting time shift of a sinusoidal signal.

Phase invariant Trellis codes Trellis codes that preserve the angular information of data symbols.

PLL (Phase Locked Loop) A feedback loop with a very tight bandwidth that can be used to track the clock signal.

Post-cursor (channel impulse response) The portion after the peak of an impulse response.

POTS (Plain Old Telephone Service) The traditional general-purpose telephone set.

Pre-cursor (channel impulse response) The portion before the peak of an impulse response.

Propagation constant The attenuation and phase shift of a cable with ideal terminations.

PSD (power spectrum density) The power density of a signal at different frequencies.

PSTN (Public Switched Telephone Network) The general telephone switched network connecting COs.

Q

QAM (Quadrature Amplitude Modulation) line code A passband line code generated through modulating in-phase and quadrature data symbols with cosine and sine phases of a carrier frequency, respectively.

QPSK (Quadrature Phase Shift Keying) signals A passband signal generated through modulating complex data symbols with four different angles of a carrier frequency.

Quantization noise The noise generated when converting an analog signal to a digital signal by rounding to the nearest signal level.

Quaternary symbol A four-level symbol.

R

Rate conversion filters A digital filter that has different input and output data rates.

RC balance network A network of resistors and capacitors for the purpose of balancing a hybrid circuit.

Receiver front-end noise A combination of background and electronics noise at the input of a transceiver.

Reed-Solomon code A class of non-binary BCH codes with features for easy code definition and implementation.

Resolution The accuracy, in terms of number of bits, of a digital circuit.

Ring trip The disconnection of ringing voltage from the loop when the called party picks up the phone.

RJ45 plugs A compact telephone style plug designed for connecting up to four twisted pairs.

Roll off A point at which the energy of a spectrum or the magnitude of the transfer function is reduced by a certain amount, usually 3 dB.

S

Sigma-delta modulator An analog-to-digital conversion technique that trades high processing speed for high resolution.

Single-ended hybrid circuit A hybrid circuit that connects to a single-ended line driver.

SNR (signal-to-noise ratio) A measure of the quality of the received signal.

Superframe format A data/synchronization structure build upon data/synchronization frames of a lower hierarchy.

SW (Synchronization Word) A particular bit/symbol pattern designed for the synchronization of data frames.

Syndrome vector A decoder-derived vector whose pattern is uniquely related to a particular error sequence.

T

T1 line A specially equipped telephone line capable of a transmission throughput of T1 rate.

TCM (Time Compression Multiplex) transmissions Transmission in opposite directions are achieved through time division multiplex. The transmission throughput over the channel is faster than the data rate.

TR (Timing Recovery) An algorithm to derive the clock from the received signal.

Transfer function The frequency response of a linear system or a linear channel.

Transmit power density The transmit power level at a particular frequency.

Transmit symbol pulse shape Defined for a baseband line code as a time-domain mask to ensure inter-operability between different equipment vendors.

Trellis code modulation A combination of a specially selected convolution code and a bit-to-symbol mapping algorithm for achieving significant coding gain with no bandwidth expansion.

U-V

Unger NEXT model A simplified piece-wise linear NEXT model derived on extensive computer simulation results.

Variable frame structure A flexible frame structure whose length can be altered to accommodate asymmetry transmission throughputs in opposite directions.

VCCO (Voltage Controlled Crystal Oscillator) A high-accuracy crystal oscillator whose frequency can be slightly adjusted through applying different amounts of voltage.

Viterbi algorithm A received data symbol recovery procedure whose decision making is based on a sequence of received data symbols.

VLSI (Very Large Scale Integrated) circuit A high-density semiconductor circuit that can include multiple signal processing capabilities in a single chip.

W-Z

White noise A noise whose power spectrum is flat.

Wiener Filter A filter whose frequency response is based on the signal-to-noise ratio of the received signal for the optimal detection of data symbols.

Zero forcing channel equalizer A linear channel equalizer whose filter coefficients are designed to produce zero inter-symbol interference at every sampling point.

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DSL

Simulation Techniques and Standards Development for Digital Subscriber Line Systems

Dr. Walter Y. Chen was born in Shanghai, China, on September 8, 1956. He received a B.S. (summa cum laude) degree in electrical engineering from the Polytechnic Institute of New York, Brooklyn, in 1982, an M.S. degree in electrical engineering from California Institute of Technology, Pasadena, in 1983, and a Ph.D. in electrical engineering from Polytechnic University of New York in 1989.

From 1982 to 1987, he was a Member of Technical Staff at AT&T Bell Laboratories, where he worked on PBX systems engineering and voice band modem VLSI chip design. From 1987 to 1989, he was a Member of Technical Staff at NYNEX Science and Technology, where he did computer network traffic and adaptive signal processing simulation studies. From 1989 to 1995, he was a research scientist at Bellcore, where he was involved with HDSL, ADSL, and HFC modem simulation studies and standards development. From 1995 to 1997, he was a research branch manager at the DSPS R&D Center of Texas Instruments, Inc., for the development of DSL technologies and High-throughput Digital Home Networks. He is currently with the Information Systems Group of Motorola working on next-generation DSL and Digital Home Network technologies. Since 1989, he has made more than 40 technical contributions at T1E1.4, ADSL Forum, and IEEE 802.14 standards committees. His main research interest is to apply circuit and system theories to the area of advanced digital communication. He holds four U.S. patents and has several others pending.

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