

High Bit Rate Digital Subscriber Lines: A Review of HDSL Progress

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Abstract—This paper gives an overview of the progress that has been made in the evolution of technology to provide DS1 rate telephone access in a restricted segment of the loop plant without intermediate repeaters, loop conditioning, or pair selection in assignment. This technology has come to be called the high bit rate digital subscriber line (HDSL). Discussed are background information on electronics in the loop plant and characterization of the transmission environment in the relevant frequency band. The progress of the HDSL study project in the T1 subcommittee of the American National Standards Institute (ANSI), accredited Committee T1, Working Group T1E1.4 is outlined. Analytical and theoretical studies to determine the limits on the transmission capabilities of the loop plant, motivated by the need to determine the feasibility limits of HDSL's, are reviewed. Also reported on is progress in technical work on suitable transmission formats, including several innovative transmission architectures and work in the areas of timing, equalization, echo cancellation, and sundry other issues. Finally, the possibility of an asymmetrical digital subscriber line (ADSL), transmitting at the DS1 rate from the central office to a remote distribution point, through the entire nonloaded loop plant is discussed, as is duplex transmission at rates in excess of 10 Mb/s over shorter distances.

I. INTRODUCTION

A SUBSCRIBER line is the basic access connection of a telephone user to the telephone network. It is the means by which the customer transmits information to a local switch to be distributed to other customers or subscribers of the same network. Consequently, the subscriber line must be relatively inexpensive if network access is to be widespread. For this reason, telephone subscriber lines have been voice frequency analog links until very recently. The principal reason for the analog nature of access was its low cost and bandwidth. With the advent of inexpensive computers, the demand for data transmission grew. Economics lead to the choice of voice frequency modems to provide data transmission at the low rates of 300 and 1200 b/s that were initially required by most users. However, transmission for these modems still used the ubiquitous voice frequency network as their transport channel. In the 1960's and 1970's, a number of technological advances led to the increase of the speeds to 2.4 kb/s, and more sophisticated signal processing and echo cancellation brought this to 4.8 kb/s in the early 1980's. More recently, trellis coded modulation (TCMod)

has been used to increase the data speed to 14.4 kb/s [1] and combined equalization and coding has produced a 19.2 kb/s rate. However, all of these techniques used the voice band to transmit data, and there was no attempt to use any of them as a substitute for the basic analog access link to the network. In order to do this, it would be necessary to replace the analog representation of voice signals by an encoded digital stream. The existing standard for doing this was embodied in digital carrier systems which, in the United States, used 56 kb/s bit streams to encode voice signals [2]. It was not economically feasible to provide information rates as high as this over an entire telephone loop plant, so the idea of a digital subscriber line was not seriously considered for ubiquitous application until recently. However, a 56 kb/s data system that required loop pair conditioning and used four-wire transmission was widely deployed [3] and, in the 1970's, long standing problems of loop transmission began to be treated by electronic means [4].

At this time, planning was begun for the evolution of the loop plant from a basically passive voice-frequency system to an access system for digital services [5]. The need for this planning was brought about by the introduction of digital loop carrier [6] and digital subscriber lines [7]. Early digital subscriber lines used time compression multiplexing (TCM) at 64 kb/s for a voice channel in Japan [8] and 56 kb/s in the United States [9]. TCM was used to avoid echoes and near-end crosstalk (NEXT). In TCM or, as it is frequently called, ping-pong, bursts of data are alternately transmitted in opposite directions on the channel. To avoid interference between the two directions of transmission, there is a short quiescent period between the end of one burst and the beginning of the next. Consequently, TCM requires a bandwidth of about 2.25 times the data rate because of time compression of the data to be transmitted and the interleaving of the quiescent period between bursts.

The advent of ISDN basic access [10] increased the data rate to 144 kb/s and required an additional 16 kb/s of overhead for a total of 160 kb/s. Even with the 2B1Q (four-level PAM) line code adopted in the North American standard [11], the resulting line rate is 80 kbaud. At these high line rates, increased pulse height loss in transmission would make TCM incapable of satisfactorily covering the nonloaded loop plant. Putting the two directions of transmission in separate frequency bands to avoid

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NEXT and echoes also incurs high loss for the direction transmitting in the upper frequency band, where losses are higher. Consequently, single-pair full duplex transmission is used. Transceivers used to achieve this employ echo cancellation (EC) [12] and decision feedback equalization (DFE) [13]. Using this technology, it was found possible to provide 160 kb/s full duplex channels on essentially all of the nonloaded wire-pair loop plant with a bit error ratio (BER) of less than 10^{-7} . Furthermore, it appears that this can be done economically enough to eventually replace voice frequency service with ISDN basic access, which consists of two "B" for "bearer" channels, transmitting at 64 kb/s, and one "D" or "data" channel operating at 16 kb/s.

ISDN basic access studies established some basic principles that should be satisfied by any digital subscriber line (DSL):

- **Ubiquity:** Any DSL should be applicable to essentially all loops that fall within some broadly defined segment of the loop plant. The basic access DSL can be satisfactorily used on essentially all nonloaded loops. Ubiquity permits relatively easy administration of the DSL.
- **Repeaterless:** Repeaters, regenerators, or any other intermediate equipment should not be part of the standard application of a DSL. Repeaters are expensive to install and maintain, difficult to administer, and require special craft training.
- **No Loop Conditioning:** It should not be necessary to alter a loop's structure to make a DSL perform satisfactorily. For example, it should not be necessary to remove gauge changes or bridged taps. Loop conditioning is expensive and causes considerable delay in providing service.
- **No Pair Selection:** The DSL should perform satisfactorily on any pair selected by any assignment system. Any restrictions should be extremely simple; restriction to nonloaded pairs is an example of a permissible restriction.
- **Compatibility:** The DSL should be spectrally compatible with voice frequency services, all ISDN services, and wire-pair digital loop carriers.

Voiceband modems and ISDN basic access transceivers achieve high performance at data rates that are four times the available channel bandwidth. This suggests that transmission at rates in the neighborhood of 800 kb/s is possible with a bandwidth of 200 kHz in a significant portion of the loop plant [14]. Furthermore, this could be done in a way that would be consistent with the principles enunciated above. This possibility suggested that DS1 rate (1.544 Mb/s) transmission in the loop plant could be achieved by employing two duplex loops operating in the neighborhood of 800 kb/s as illustrated in Fig. 1.

A combination of the loss characteristics of the loop plant and the capabilities of modern transmission technology suggested that a carrier serving area (CSA) might be an appropriate plant segment for such a system. A CSA is a loop plant administration area, as illustrated in Fig.

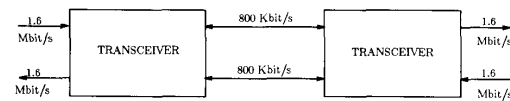


Fig. 1. A 1.6 mb/s dual duplex access system.

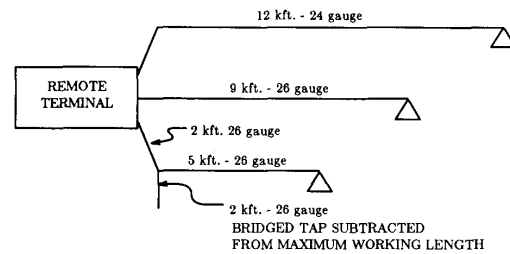


Fig. 2. A carrier serving area around a remote terminal.

2. A CSA surrounds a remote terminal or a wire center. In a CSA, a maximum working loop length of 12 kft is permitted only if there are not bridged taps and only if 24 gauge wire is used exclusively. A 26 gauge pair is restricted to a maximum length of 9 kft, and bridged tap length is subtracted from maximum working length. In a CSA that surrounds a wire center, the maximum permissible working length is reduced by a kilofoot. The definition of a CSA is sometimes modified to suit local needs and the concept of a CSA is still evolving.

Preliminary feasibility studies of the dual-duplex transmission approach to network access at the DS1 rate led to a study project of American National Standards Institute (ANSI) accredited T1, Working Group T1E1.4 [15]. Thus, the study of DS1 rate DSL's or high bit rate DSL's (HDSL's) became formalized. Since that time, considerable work has been done in T1E1.4, industry, and the academic world in an effort to bring HDSL to reality. Some of the more recent contributions to this work appear as companion papers in this issue of J-SAC. This paper attempts to summarize some of the remarkable contributions to technology that have emerged from HDSL studies.

Section II summarizes the proceedings, until October 1, 1990, of Working Group T1E1.4 of ANSI that focused on the HDSL. The discussion is cursory, but the author attempts to add structure to the presentation by parenthetical remarks and by correlation of the standards study work with work that is reported in the published professional literature.

Various attempts have been made to establish limits on the transmission capability of the loop plant. There have been many theoretical and simulation studies of HDSL transmission in an attempt to establish the general viability of HDSL's in the loop plant. Section III attempts to summarize some of these and to relate the work on capabilities to work on modeling of the loop plant.

As in the study of ISDN basic access, line codes and line signal formats have received considerable attention in HDSL studies. Section IV is an annotated tour of this

work. At this point, it should be noted that the problems and issues of reliable wire-pair digital transmission are similar to those of magnetic recording [16], [17]. The work on line codes and transmission format, in general, has benefitted considerably from the magnetic recording work. Section V discusses work that has been done on various aspects of HDSL transceivers, including timing, echo cancellation, filtering, and decision feedback equalization (DFE). Section VI attempts to provide an overall summary of the HDSL work and suggests some possible areas for future developments in wire-pair transmission technology. It is suggested there that it might be possible to provide an asymmetrical channel at the DS1 rate over most of the nonloaded loop plant without repeaters and without pair conditioning. It is also pointed out that there is interest in transmission at multimegabit rates over shorter loops for LAN applications.

Because the bandwidth for an HDSL is greater than that of the basic access DSL, it can be anticipated that NEXT noise mitigation is a more difficult problem. This fact is the principal reason for restricting the range of an HDSL to a CSA. Although an adequate model of NEXT is believed to exist [18], [19], work has continued on the measurement and modeling of crosstalk. For example, crosstalk and impulse noise data on European cables was reported by Szechenyi [20]. Of some concern with respect to crosstalk characteristics is premises wiring. The paper by S. A. Cox, L. M. Clarke, A. J. Slater, and M. A. Wright reported on some measurements in the United Kingdom in this area [21]. S. B. Pierce reported on measurements in North American cables [22]. A. Fung, L. S. Lee, and D. D. Falconer described preparations for characterizing crosstalk in [23]. An attempt was made by J. W. Lechleider to find a simple representation of NEXT to use in design to suppress NEXT [24]–[26]. A long-standing conjecture concerning NEXT is that since it is not Gaussian, the capacity of any NEXT-impaired channel is different than it would be for a Gaussian channel. In fact, it has been suggested that advantage could be taken of the fact that NEXT must, of necessity, be bounded above and below. In a theoretical and simulation paper, K. Kerpez showed that, in fact, NEXT is Gaussian except in the low probability density tails of the distribution. He then calculated the entropy power of NEXT and found that it was essentially the same as Gaussian noise with the same power spectral density [27].

Although there is scant data on crosstalk in the frequency band of interest to the HDSL, it is believed that crosstalk is reasonably well represented by the extant theoretical models. Impulse noise, on the other hand, is not as well understood, even though it has been studied for some time [28]–[32]. A more recent model was proposed by J. J. Werner [33]. Since the maximum loop loss that must be accommodated by ISDN basic access receivers is much greater than that which will be required of HDSL transceivers, impulse noise may not be as serious a problem in the HDSL. However, the impulse noise environment is changing in the loop plant because of the increase

in the variety of equipment that is being installed in recent years and because of the changes in customer premises wiring. Because of the relative paucity of information on impulse noise, there have been several recent attempts to measure it [34]–[38]. Further work was done on characterizing, by measurement, the performance of systems in an impulse noise environment [39]–[41]. J. W. Modestino and M. V. Eyuboglu simulated receiver performance in an impulse noise environment [42]. J. W. Modestino, D. Sargard, and R. E. Bollen proposed coding to mitigate against impulse noise [43].

Finally, in addition to impulse noise and crosstalk, induction noise is also relevant to HDSL performance. One reason for this is that induction currents can be substantial enough to cause nonlinearities in system performance. C. Chrysanthou reported on plans for testing of inductive noise on DSL circuits in [44].

II. STANDARDS COMMITTEE ACTIVITY

In 1988, the American National Standards Institute published a standard for the ISDN basic access interface. Many of the issues that were considered in the deliberations that led to this standard are also involved in formulating the requirements for an HDSL. In fact, many of the participants in the Working Group (now designated T1E1.4) that formulated the basic access standard are also involved, in their respective organizations, in studies of the HDSL. Working Group T1E1.4 initiated a study project for the HDSL in 1988 [45]. This section attempts to give a running account of the consequent deliberations of the Working Group up until the time of writing this paper. References to contributions (to ANSI) of Working Group T1E1.4 are given in the contribution number, which takes the form "T1E1.4/YY-NNN" where YY is the year of the contribution and NNN is the sequential number of the contribution in the indicated year.

Early in the study, Lechleider suggested that there are a number of issues that needed study before an HDSL could be implemented [48]. These included A/D conversion, timing, signal processing (including exploiting NEXT cyclostationarity), echo cancellation, Viterbi receivers, adaptive transmitters, wire center and customer premises wiring characteristics, impulse noise, other services sharing the same cable sheath as the HDSL, and architecture issues. R. A. McDonald indicated some other issues that must be addressed by the study project [47]. He pointed out that technical feasibility and the type of service to be offered would have a bearing on the desirability of establishing a new standard for the HDSL. For example, if the HDSL is used for ISDN primary access transport, no new standard would be required. He and C. F. Valenti also outlined the assumptions used in the Bellcore studies of HDSL's [48]. The critical issues that they discussed are that the same data used in basic access studies would again be used, that the required performance would include a bit error ratio of 10^{-7} , and that the margin used in theoretical and simulation studies would be 12

dB, while the margin on a measured piece of equipment need only be 6 dB. The crosstalk model has a NEXT loss of about 57 dB at 80 kHz and decreases at about 15 dB per decade for frequencies above about 20 kHz. Plant models were taken from the 1983 loop survey, and impulse noise data were taken from a survey of impulse noise by the NYNEX Corporation. These data were used in basic access studies. Valenti went into further detail on the margin determination in a later contribution [49]. P. Adams and J. Cook gave the margin issue further debate in a later contribution [50]. They argued that a 6 dB margin on an experimental HDSL prototype and a 12 dB margin on a theoretical model of an HDSL were too severe. Rather, more detailed accounting of the various impairments that HDSL's are likely to encounter in practice is desirable.

J. W. Lechleider pointed out that DS1 rate, single-pair HDSL's using modified basic access technology would have a range of about a mile [51]. He also pointed out that 800 kb/s DSL's using modified basic access technology, but operating at 800 kb/s, would perform satisfactorily on most CSA loops but would have a theoretical margin of about 6 dB on the lossiest loop in a population of CSA loops [52]. A study of using trellis coded modulation (TCMod) to make up the needed extra 4 dB was reported by him [53]. He found that a simple trellis code would provide the performance improvement required in a 2B1Q line code, but he assumed that a DFE operating without error was embedded in a Viterbi receiver in the study. Since the DFE would use detection estimates formed prior to maximum-likelihood processing, the DFE would be operating on data with a higher bit error ratio than the required line performance. He also introduced structured channel signaling (SCS), which is discussed later in this paper, as a possible method of achieving the required performance [54].

V. B. Lawrence, J. J. Werner, and N. Zervos studied the Shannon capacity of subscriber loops using a simple crosstalk model and identical signal spectra in the two directions of transmission [55]. They found that the capacity of subscriber loops far exceeds the DS1 rate. A guide for an economic study of HDSL's was given by C. I. Cook [56]. Some possible enhancements to basic access technology, discussed by E. Arnon [57], achieve performance that is acceptable on a large percentage of loops. This contention was later supported by data on the distribution of losses in a CSA [58]. K. Sistanizadeh calculated the performance of a DFE with an infinitely long feedback filter on a population of CSA loops [59]. He found that performance is not enhanced by increasing the length of the feedforward filter in the DFE to more than nine taps, with baud-rate sampling. G. E. Smith proposed coded QAM as the line code for the HDSL [60]. He suggested that two-wire transmission at DS1 rates was feasible in a CSA. He also pointed out that, since QAM is a passband code, it could be used above voice. In another contribution [61], he gave calculations of the theoretical performance of 64 QAM with a bandwidth of 257.33 kHz and

a TCMod gain of 4 dB. These theoretical results showed that performance was adequate for the crosstalk model used, which was less conservative than the Bellcore model. In addition, there was no margin. C. F. Valenti and R. A. McDonald compared four-wire simplex and duplex transmission performance [62]. They pointed out the range penalty suffered by the simplex approach. J. A. C. Bingham pursued further analysis of the simplex-duplex issue, noting that the best line signal used in these two cases might be different [63].

Spectral compatibility of HDSL systems with themselves and other systems in the same cable sheath was the topic of a contribution by T. Starr and M. Bono [64]. They proposed various standard mixes of HDSL systems and other systems such as ISDN basic access and primary access, and 56 Kb DDS. They later revised their proposed mix of systems [65]. In another contribution along this line, N. Cole and N. Lynch-Aird suggested that a wider assortment of interfering systems be considered in HDSL feasibility studies [66]. They later revised their estimates of range [67]. In this same vein, G. E. Smith suggested that many parameters, not only self-NEXT, should be used in comparing HDSL systems [68]. G. J. Pottie made a comparison of the performance of baseband and passband systems on a 9 kft, 26 gauge loop and on a 12 kft 24 gauge loop [69]. He found that in self-NEXT, baseband and passband systems have roughly the same performance, and that signaling at two bits per dimension leads to better performance than signaling at three. Another result was that baseband systems perform better in the presence of interference from ISDN basic access systems in the same cable. The reason for this is that interference from basic access systems is very low at frequencies below 20 kHz so that the low-frequency portion of the HDSL band is a good place to put power to mitigate the effects of basic access interference. K. Sistanizadeh also compared the performance of QAM and PAM on a large population of CSA loops [70]. His results corroborated those of Pottie.

A set of reference impulse waveforms derived from a survey of impulse noise by the NYNEX Corporation was presented by E. Arnon [71]. He pointed out that, because the maximum loss in HDSL's is lower than it is in basic access DSL's, impulse noise should be a less severe problem in the HDSL. K. J. Kerpez reported on a comparative study of impulse noise performance of 2B1Q PAM systems and QAM systems [72].

In three separate contributions [73]–[75], Lechleider pointed out that crosstalk can be cancelled, signal-to-noise ratios can be averaged, and pair-to-pair correlation of noise can be exploited in two-pair digital subscriber lines. He also showed how to implement transceivers to realize these benefits in a later contribution [76]. More detail is given on the content of this contribution later in this paper and in a companion paper by Lechleider [77].

G. Forney, Jr., gave what was essentially a tutorial on coded modulation as applied to HDSL transmission [78]. This was coupled with a discussion of generalized Tom-

linson precoding by M. V. Eyuboglu, G. Pottie, and L. Brown as it might be applied to HDSL's [79]. N. J. Lynch-Aird made an analytical comparison of recursive and nonrecursive equalizers with Tomlinson precoding and a DFE and a PAM line code [80]. Tomlinson precoding is discussed in more detail later in this paper. A tutorial was also given on carrierless AM/PM by M. Sorbara, J. J. Werner, and N. Zervos [81].

In an effort to narrow the possibilities for consideration for HDSL implementation, D. L. Waring and K. Sistanizadeh suggested that HDSL configurations to be considered in further study be limited to repeaterless DS1 applications in CSA [82].

R. C. McConnell noted that there might be complications with HDSL's that are connected to DS1 lines that have slight frequency or baud offsets [83]. Another implementation issue, the frame format, was discussed by E. Arnon and E. Ehrlich [84]. Other framing format proposals were made by W. R. Murphy for a 2B1Q implementation of the HDSL [85] and by H. Nicholas and H. Flagg [86].

The first report on measurements of experimental HDSL's was made by Arnon and Ehrlich [87]. This contribution reported a margin of 3.7 dB on a 12 kft, 24 gauge loop for a full duplex 772 kb/s system that uses the 2B1Q line code. R. Cupo and M. Sorbara reported on measurements of a carrierless AM/PM HDSL system [89]. They showed margins of greater than 6 dB on some CSA loops for a DS1/2 two-pair simplex system with simulated crosstalk noise. K. T. Foster reported on measurements of the performance of a 3B1O system [89]. He reported an operating margin of 3.8 dB in a 49-disturber simulated NEXT environment. Although the number of PAM levels that achieve optimal performance in a self-NEXT environment is between five and six, the eight level code was used for simplicity of implementation.

A plan for the laboratory testing of proposed HDSL systems was given by B. A. Blake [90]. The proposed plan is similar to that used for the testing of ISDN basic access transceivers. Blake also described a loop emulator testbed that permits the rapid emulation of a large number of loop configurations for the testing of proposed HDSL architectures [91]. A collection of eight CSA loop configurations that are representative of the population and may be used for short studies of HDSL architectures was presented by K. Sistanizadeh [92]. J. Bingham proposed the use of the transmission matrix as a frequency domain representation of a loop in response to perceived disparities in calculated loop transmission characteristics [93].

D. A. Fisher and D. R. Donovan reported on a 6B + D HDSL using the 2B1P line code that is being studied for use in Europe [94]. They also proposed a framing structure for the system [95].

Finally, T. Starr and M. Bono pointed out that two loops, selected at random between a central office and a customer, might follow substantially different routes [96]. Consequently, if routine provisioning practices are to be used, a two-pair HDSL system must be capable of han-

dling some difference in the transmission characteristics of the constituent pairs.

III. THE TRANSMISSION LIMITS OF THE LOOP PLANT

The fact that serious consideration is currently being given to duplex wire-pair transmission at 800 kb/s for distances up to 12 kft would have been a surprise to most engineers a short time ago. In fact, in 1979, R. K. Even, R. A. McDonald, and H. Seidel estimated that a limit point for loop transmission was 200 kb/s over 20 kft of 22 gauge cable for duplex transmission [97]. The advances in VLSI technology and signal processing that have made 800 kb/s duplex transmission possible have also stimulated considerable interest in the transmission capabilities of wire-pair telephone loops. This work has, in turn, stimulated new work in high-speed wire-pair transmission. This section summarizes some of this work.

Perhaps the earliest published calculations of loop capacity were by V. Joshi and D. D. Falconer [98], with later elaboration by Joshi in his thesis [99]. These papers, and others [55], [100], calculate the capacity of (self) crosstalk-limited digital channels by using the basic formula for the capacity of a continuous channel impaired by colored Gaussian noise:

$$C = \int_0^W \log_2 \left[1 + \frac{|H(f)|^2 S_1(f)}{|X(f)|^2 S_2(f)} \right] df \quad (3.1)$$

where C is the channel capacity in bits per symbol, W is the single-sided bandwidth of the digital channel, H is the transmission (loop) transfer function, S_1 is the transmitted pulse spectral energy density, X is the NEXT transfer function, and S_2 is the disturbing (crosstalking) pulse's spectral energy density. Related work is given in [101], where it is found that a peak-power constraint lowers channel capacity below that of an average power-constrained channel. Also relevant is [102], which obtains a lower bound on capacity with peak power limits on band-limited channels. Peak power constraints are relevant to structured channel signaling, discussed later in this paper. If the assumption is made that NEXT is Gaussian, another relevant paper is that of W. Hirt and J. Massey [103]. As mentioned earlier, K. Kerpez found that the assumption of Gaussian NEXT leads to substantially the same results for channel capacity using the actual distribution of NEXT.

In all of the above studies, it was assumed that the transmitting and disturbing pulse shapes were the same so that (3.1) becomes

$$C = \int_{f \in A} \log_2 \left[1 + \frac{H^2}{X^2} \right] df \quad (3.2)$$

where A is the support of the pulse shape function, so that the capacity depends only on the region of the frequency band that is occupied by the pulse and not on the details of its shape. Using (3.2), the above authors obtained varying results depending on the detailed assumptions that

they used, but Shannon capacities well in excess of the DS1 rate for 9 kft of 26 gauge pair are typical.

Although Shannon capacity calculations are indicators of the best that is physically achievable, they must, in the current case, be considered only as indicators. One reason for this is that it is difficult to approach the channel capacity with any signaling scheme except under special conditions, and then with TCM. For example, it is known that for Gaussian channels there is a gap of about 9 dB between the Shannon limit and uncoded QAM [104]. Furthermore, the Shannon capacity is the limit on error-free transmission rates using possibly infinitely complex coding and decoding procedures. The "cutoff rate" [105] is often considered to be a more realistic bound on achievable information rates because the complexity of sequential decoders increases rapidly as the code rate is increased beyond the cutoff rate [106]. The cutoff rate was calculated by J. J. Aslanis and J. M. Cioffi [107] for a collection of CSA loops with results similar to those obtained in the Shannon capacity calculations, i.e., CSA loops are capable of transmitting the rates in excess of DS1 rate with bit error ratios less than 10^{-7} .

As noted above, assuming that the pulse shapes transmitted in the two directions are the same eliminates the effect of the transmitted pulse shape on the Shannon capacity when self-NEXT is the only impairment. Lechleider noted that capacity might be improved if different pulse shapes that have the same support on the frequency line were used in the two directions of transmission [108], but found that the improvement in achievable capacity is small when optimal pulse shapes are used.

If NEXT is the only impairment, the above arguments on capacity are, to some extent, only theoretical guides because NEXT can be completely eliminated if different frequency bands are used for the two directions of transmission. Time compression multiplexing achieves the same result. Another technique that would eliminate NEXT is crosstalk cancellation, as discussed by K. Steiglitz, M. L. Honig, and B. Gopinath [109]. The possibility of eliminating NEXT led Lechleider to propose an adaptive block-code transmitter that would achieve the same results [110]. In another new departure in NEXT suppression, P. Crespo, M. L. Honig, and K. Steiglitz found a combination of linear prefiltering and post-filtering that will eliminate NEXT [111]. This technique, like NEXT cancellation, requires, in effect, the knowledge of NEXT transfer functions between the pairs involved. For this reason, it is impractical because the identity of the disturbing transmitters is not known in most applications. An exception to this is in coordinated transmission [77], where both NEXT cancellation and post-filtering to eliminate NEXT between the constituent pairs of a two-pair HDSL are employed.

In the absence of knowledge of NEXT transfer functions, NEXT can still be eliminated but substantial excess bandwidth is required. In order to see this, it is only necessary to note that a receiver must be able to detect the distinction between a received signal and NEXT in each

baud interval, so that the signal and NEXT must be linearly independent. This requires that the function space occupied by the received signal and crosstalk interference in a given baud interval must be at least two-dimensional if there is only one crosstalk. This implies that the system bandwidth must be twice as great as that required for the signal alone if there is one crosstalk present. Of course, as the number of distinct, i.e., linearly independent, NEXT signals increases, so does the bandwidth requirement. Since signal loss increases and crosstalk loss decreases as bandwidth increases, this bandwidth expansion implies greater crosstalk energy to be suppressed.

The possibility of eliminating crosstalk has led to consideration of other limitations on the transmission capabilities of subscriber loop pairs. One such approach is to consider limited precision in receivers. One looks for transmitted signal sets that will satisfy some minimal separation criterion at the receiver as M. L. Honig, S. Boyd, and B. Gopinath did [112], and as later pursued by M. L. Honig, K. Steiglitz, and B. Gopinath [113], [114] and in other papers [115], [116]. This work naturally led to the problem of finding transmitted signals that are far apart at the receiver [117], [118]. A problem connected with finding transmitted signals with maximal separation is that of transmitting faster than the Nyquist rate to achieve greater channel throughput. Doing this naturally leads to excess loss and intersymbol interference [119], [120]. The problem of finding maximally separated signals when signaling faster than the Nyquist rate [121] has received considerable attention by D. J. Hajela [122], [123] and D. J. Hajela and M. L. Honig [124].

Another notion that has received renewed theoretical interest because of the HDSL effort is that of the ϵ capacity of a channel which was discussed by W. L. Root [125] and subsequently used in other studies [126], [127]. The ϵ capacity of a channel is the logarithm of the maximum number of output signal points that have minimum separation between any two of them of at least ϵ when the transmitted signal has energy less than or equal to one. This concept is illustrated in Fig. 3 for a two-dimensional channel. More typically, the input and output spaces are of higher dimension and the problem of estimating the ϵ capacity is similar to the sphere-packing problem [128]. D. J. Hajela and M. L. Honig applied the concept of ϵ capacity to multiinput/multioutput systems in a way that is approximately applicable to telephone cable transmission in [129].

The general conclusion of all the work on channel capabilities appears to be that there are no fundamental physical limits to prevent provision of DS1 service without regeneration, pair conditioning, or pair selection within a CSA. However, this work has led to some new innovations related to performance theory. For example, in what has come to be called structured channel signaling (SCS) [130], Lechleider used a discrete analog of the orthogonal function set used by R. G. Gallager [131] and J. L. Holsinger [132] to determine the capacity of time-invariant channels. Lechleider also showed that these

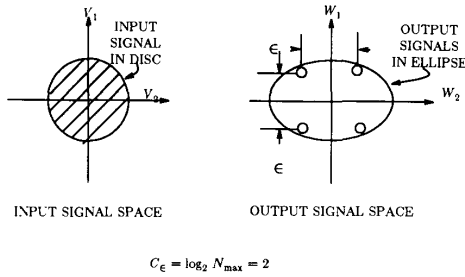


Fig. 3. ϵ capacity for a particular two-dimensional channel.

functions form a sequence of functions that are the signal shapes that, for a given transmitted energy, produce maximum signal-to-noise ratio (SNR) at the detector. In fact, consider a channel operator that is completely continuous so that it maps bounded sequences of signals into convergent sequences. Consider the eigenvalue problem

$$\mathbf{H}^* \mathbf{H} \phi_n = \kappa_n \mathbf{N} \phi_n \quad (3.4)$$

where $\mathbf{H}$ is the channel operator and $\mathbf{N}$ is the noise covariance operator, i.e.,

$$\mathbf{N} = \langle \mathbf{n} \mathbf{n}^* \rangle. \quad (3.5)$$

Since the channel operator $\mathbf{H}$ is nonsingular, the eigenvalues $\{\kappa_n\}$ in (3.4) are bounded below and must obviously be positive. But zero is a limit point of the spectrum [133]. If the noise covariance operator is nonsingular, the κ_n are bound above. With a singular noise covariance operator, the noisy channel is singular and it has unbounded capacity [134]. The eigenvalues in (3.4) are clearly the ratio of the output signal energy to the expected noise energy when the transmitted signal is an eigenvector and the noise is projected onto the corresponding eigensubspace. Thus, the ϕ_n are the signals that maximize SNR at the detector. Studies of the capacity of subscriber loops using the channel structure elucidated in [130] were reported on in [135] and [136]. It was found in these studies that reliable information transmission rates are increased by using SCS, but the required added transmitter complexity appeared formidable. SCS is discussed further in Section IV.

IV. TRANSMISSION FORMAT

The studies related to ISDN basic access considered many line codes [14], [137], [138]. Since the 2B1Q line code was adopted for the North American standard for ISDN basic access [139], [140], it is quite natural that it should be considered for application to the HDSL. Since straightforward application of basic access technology to HDSL applications does not provide adequate performance, as indicated in Section II, some enhancement of this technology is required before the 2B1Q code can be considered as a viable contender for HDSL use. Some of these enhancements are discussed in Section V. One

promising technique, which moves ISI cancellation from the DFE in the receiver to the output of the transmitter, is Tomlinson precoding [141]. Tomlinson precoding avoids the degradation of the performance of TCM that is brought about by spectral nulls such as those that exist in partial response channels [142], [143] and by intersymbol interference in general [144]–[147]. Tomlinson precoding requires an adaptive transmitter, however. Coding techniques for employing TCM on dispersive channels have also been proposed [148], [149]. These techniques effectively code to produce a spectrum that is least sensitive to known channel frequency variability. Another approach is that used by A. Ruiz and J. M. Cioffi [150].

Quadrature amplitude modulation (QAM) is a second line code, which has emerged from extensive application in radio and voiceband modems. The applicability of QAM to HDSL application is, at first, somewhat surprising because QAM puts the transmission band at frequencies higher than baseband systems do, with consequent greater signal loss and lower crosstalk loss. However, if the effective carrier frequency is low enough, this is apparently a relatively small disadvantage [151], [152]. One reason for this is that most subscriber loops have severe amplitude and phase distortion in the baseband below 20 kHz, which is avoided by a passband system with its lower band-edge above 20 kHz. This makes QAM systems more attractive candidates for TCM [153]. There has also been comparison of QAM performance with faster-than-Nyquist signaling [154].

There are some fundamental similarities between 2B1Q and 16 QAM that come about because both systems must have their frequency bands severely limited to suppress noise and NEXT. For example, if a 100% duty cycle square pulse is used in 2B1Q, two consecutive equal phases will, after bandlimiting, approximate a cycle of a cosine wave, as shown in Fig. 4. Contiguous opposite-polarity pulses lead to a cycle of sine wave, as shown in the figure. These two shapes approximate the only two functions that have essentially all of their energy confined to a time interval of length $2T$ and a frequency band of width $1/2T$.¹ In fact, these two shapes are the same ones that are used in baseband QAM. In addition, contiguous 2B1Q pulses of unequal magnitude would lead to a linear combination of the pulses of Fig. 4 after bandlimiting, completing the analogy to QAM. Because of the analogy between these two, any performance difference between them depends on the location of the transmission band for QAM. As noted earlier, impulse noise tends to have its spectrum concentrated at frequencies below 70 kHz. Consequently, bandpass systems such as QAM offers a relatively improved performance, relative to impulse noise, over baseband systems. Of course, the relative difference in performance depends on the location of the lower band-edge of the passband system.

¹More precisely, the pulses in Fig. 4 are linear combinations of the two prolate spheroidal wave functions [155] with greatest energy in the given time interval and frequency band.

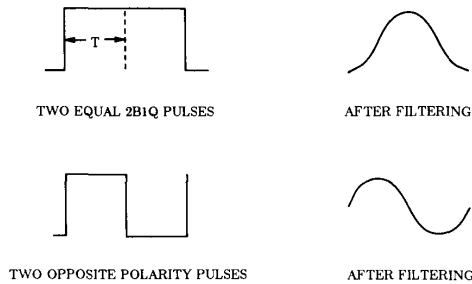


Fig. 4. Two contiguous 2B1Q pulses before and after band limiting.

QAM uses two different signal shapes in the same time interval, that are separated at the receiver by appropriate signal processing. From the above discussion, it is apparent that two-length blocks of 2B1Q pulses could be used as information carriers that are separated out at the receiver. This notion can be further extended to block codes that are simultaneously transmitted in the same time interval to be separated out at the receiver by using appropriate signal processing techniques. This technique was optimized by Lechleider in what has come to be known as structured channel signaling (SCS) [130]. This technique, illustrated in Fig. 5, forms blocks of data by using finite impulse response filters that are added together to form a transmitted block. The constituent blocks are separated out at the receiver by using a bank of FIR filters with tap-weight profiles that are biorthogonal to the received constituent blocks. Of course, implementation of SCS requires block signal processing such as block echo cancellation and block-decision-feedback equalization, which are discussed later in this paper. SCS performance was studied by K. Sistanizadeh and J. W. Lechleider. SCS requires, for optimum performance, that the number of levels transmitted on each subchannel be adaptive. This adds a level of complexity that is probably too great for economical HDSL implementation.

There have been several variations of the SCS idea that have been published. The earliest appears to be Kasturia *et al.* [156], which examined the concept of block coding. D. W. Lin investigated block modulation schemes of one to four dimensions with time division multiplexing [157]–[159]. In these papers, Lin uses prewhitening filters with an added noise predictor that encompasses the predictor studied by C. A. Belfiore and J. H. Park [160]. S. Kasturia and J. M. Cioffi exploit the idea of using vector coding to treat intersymbol interference within blocks in [161] and J. T. Aslanis joined them in another paper along the same lines [162]. In [163], Dudevoir *et al.* investigate least-squares block-decision-feedback equalization with four-dimensional trellis codes. A suboptimal variant of SCS is multitone signaling, which is given an excellent review by J. A. C. Bingham [164]. The performance capabilities of multitone were studied by N. Zervos and I. Kalet [165], who conclude that multitone is of marginal value compared to PAM and QAM, and by Kalet [166]

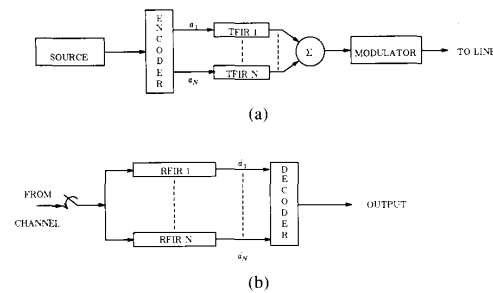


Fig. 5. Structured channel signaling (SCS). (a) Structured channel transmitter. (b) Structured channel receiver.

who shows that multitone QAM is about 9 dB worse than channel capacity, which is the same as the result for PAM and not as good as that for a DFE [167]. Tu *et al.* apply some of the ideas of vector transmission to multitone [168]. One of the advantages of multitone transmission is that impulse noise energy is distributed across the subchannels of the multitone channel, thus possibly reducing the sensitivity of the system to impulses. In addition, it was pointed out by J. Tu and J. M. Cioffi [169] that multitone is relatively robust to channel estimation errors compared to a DFE receiver.

One line coding idea that, at first, appears peculiar to two-pair DSL's is coordinated transmission [77]. Coordinated transmission relates the height of a pulse transmission on one pair of a two-pair DSL to that transmitted on the other. The relative heights of the pulses on the two pairs are determined by the pair channel characteristics. Coordinated transmission has channel signal processing gains due to:

- averaging of signal-to-noise ratios received on each channel (in dB);
- cancelling mutual NEXT between the constituent pairs;
- exploiting the correlation between the interference and NEXT on one pair with that on the other [170].

In effect, coordinated transmission is a method of increasing the channel capacity. Effective coding is a means of approaching the capacity that the channel has. In fact, one of the effects in coordinated transmission, averaging of the SNR's on the constituent pairs, is really not even increasing channel capacity. Averaging the SNR's narrows the spread of SNR's that can be anticipated in engineering for the DSL.

In addition to the gains inherent from improving channel capacity by using coordinated transmission, any line code can be concatenated with it to achieve compounded gains. K. J. Kerpez has discovered a way that a coordinated channel may be coded to achieve the channel capacity gains [171]. All of the signal formatting or line-coding techniques mentioned above may be combined with trellis-coded modulation (TCMod) [172]. The literature on TCMod is vast; to wit, recent issues of this JOURNAL [173]. One direct application of TCMod was given

by Falconer *et al.* [174], which uses reduced state sequence estimation because of the possible complexity inherent in TCM and Viterbi decoders on the dispersive channels that HDSL's will be implemented on. V. Joshi and D. D. Falconer studied the reduction of the complexity of receivers for TCM further in [175] where they show, by simulation, that RSSE is better than a conventional DFE. This paper also investigates the M-L algorithm, which searches through M states where $M \ll K^L$. This line of work was continued by Joshi in his thesis [176]. M. V. Eyuboglu, S. U. Qureshi, and M. P. Chen also studied RSSE for this purpose [177]. RSSE was also discussed by Eyuboglu and Qureshi in [178]. Other work on the reduction of the complexity of Viterbi receivers abounds [179]–[182]. J. W. Bergmans, S. A. Rajput, and F. A. M. van de Laar [183] used decision feedback equalization to simplify Viterbi receivers [184] used for detection on dispersive subscriber loops [185].

J. Tu and J. M. Cioffi focus on the concatenation of coset codes and SCS in [186]. Of particular interest are multidimensional codes [187], [188] because of their potential natural application to SCS or coordinated transmission.

Because NEXT loss decreases at about 15 dB per decade and signal loss increases at about 5 dB per decade, signaling techniques that decrease bandwidth are particularly interesting to HDSL implementers [189]. One such technique is to signal faster than the Nyquist rate [120] and accept the consequent intersymbol interference but apply the appropriate signal processing [119], [190], [191]. This work naturally suggests the problem of determining the minimum distance between signals with faster-than-Nyquist signaling. J. E. Mazo and H. J. Landau [192] showed that for rates up to about 25% above the Nyquist rate, the minimum distance between the output sequences of an ideal bandlimiting channel does not drop below the value that it would have in the ideal case. The work was based on earlier work by D. J. Hajela [122], which was discussed in Section III.

Another technique that has received some attention that affects the line code is Tomlinson precoding [141]. As mentioned earlier, Tomlinson precoding effectively replaces the feedback filter in a DFE with an adaptive preconditioning filter at the transmitter output, thus avoiding error propagation when a detection error is made. Tomlinson precoding was proposed for HDSL application by S. Kasturia and J. M. Cioffi [193]. In this paper, the authors extend Tomlinson precoding to multidimensional channels. The idea of Tomlinson precoding was further extended by M. V. Eyuboglu and G. D. Forney in [194].

Since any digital transmitter produces a cyclostationary random process, crosstalk must also be cyclostationary with a period equal to the baud interval of the transmitted signal [195]. B. R. Petersen and D. D. Falconer have shown that cyclostationarity can be exploited to advantage if all of the transmitters in the same cable sheath are synchronized [196]. Synchronization of the transmitters is probably impractical in wire centers and it is certainly

so in most remote cable-ends. However, it was recognized by M. Abduhrahman and D. D. Falconer that crosstalk is frequently dominated by a small number of disturbers so that cyclostationarity might still be exploited on a random basis [197].

HDSL work has led to other proposals for line formats [198]. There was also a proposal for transmission at a rate that is determined by the loop that an HDSL is implemented on [199], [200]. However, rate adaptation would require reconsideration of many network issues in ISDN that have already been resolved for DS1 service and ISDN primary access, and would probably require substantial network modification.

V. HDSL TECHNOLOGY

The HDSL must be relatively inexpensive to be viable. At the same time, it must perform to exacting standards at a high bit rate for the distribution plant. These general requirements have fostered research into technologies that might otherwise not have been considered or the modification of technologies originally used in other contexts. Not all of these modifications are suitable. For example, switched capacitor receivers, suitable at lower speeds [201], would be too slow for the HDSL. The need for great speed led to investigations of digital transceivers that are computationally efficient [202]–[204], and the required low cost dictates efficient VLSI structures [205].

In addition to these general considerations, there were many specific technical issues that were given more focused attention because of the HDSL. The design of the best combination of transmitter and receiver filters to maximize SNR at the detector (or minimize mean square error) was found in studies of structured channel signaling mentioned earlier. The combination of linear transmitting and receiving filters that minimizes mean square error was studied by J. W. Lechleider and K. Sistanizadeh in [206]. P. Crespo and M. L. Honig also started a study of signal processing in [207]. A few years ago, N. Amitay and J. Salz studied the use of linear equalization to dually polarized radio channels [208]. This work was extended by M. Kavrehad and J. Salz in [209]. Related work in this area dates back to Tufts' study of joint transmitter–receiver optimization in PAM [210] and to Lechleider's thesis [211]. Recent related work was done by H. S. Malvar and D. H. Staelin [212]. Of course, the recent work of Steiglitz, Gopinath, Crespo, and Honig in the area of linear filtering that was discussed earlier in this paper is also germane. Related work on linear receivers was performed by P. A. Gallo and S. Pasupathy [213].

Timing recovery at HDSL speeds can be a critical problem because of exacting performance requirements and because of the adverse effect of mistiming on echo canceller performance, as pointed out by D. D. Falconer [214]. There are many approaches to timing a remote receiver, but these were sharpened in HDSL-related studies. For example, A. Gottlieb, P. Crespo, J. Dixon, and T. R. Hsing [215] expanded on the basic idea of K. H.

Mueller and M. Muller [216]. Channel estimation in synchronization was also used by A. Aghamohammadi, H. Meyr, and G. Ascheid [217]. Related work, that was actually done in support of ISDN basic access, was reported by C.-P. Tzeng, D. A. Hodges, and D. G. Messerschmitt and by S. G. Brophy and D. D. Falconer [218]. Just as timing can severely affect echo canceller performance, the anticipated high echo levels relative to the incoming signal levels will make timing a difficult problem. This problem was addressed by M. Moeneclaey [219]. From the above, it can be concluded that timing and echo cancellation are intimately related.

The high rate of HDSL transmission and stringent requirements for echo suppression have led to continuing investigations of echo cancellation. For excellent reviews of echo cancellation, see the papers by D. G. Messerschmitt [220] and K. Murano, S. Unagami, and F. Amano [221]. Another frequently referenced paper that discusses compensation for incoming signals in echo canceller adaptation and is a well-written tutorial in echo cancellation is the contribution on adaptive reference echo cancellation by D. D. Falconer [222].

As mentioned earlier, proper timing is essential to good echo canceller performance because of anticipated high echo levels relative to incoming signals. In this regard, digital phase locked loops (DPLL) used in timing pose the possibility of a particularly severe impairment. This is due to temporary residual or uncanceled echoes occurring while an echo canceller is adapting to changes in an echo pattern brought about by a sudden timing phase change from the DPLL. A clever method of dealing with this problem was proposed by D. G. Messerschmitt [223]. Another approach to this problem was described by A. Kanemasa, S. Koike, A. Sugiyama, and F. Nakagama [224].

Another problem that arises in echo cancellation, at levels of 60–70 dB of cancellation, is the effect of small nonlinearities such as those arising from line coupling transformers. O. Agazzi, D. G. Messerschmitt, and D. A. Hodges proposed a clever means of dealing with these nonlinearities [225]. They made the observation that the echo at (say) time zero is a function of the recently transmitted bits, i. e., $f(B_0, B_1, \dots, B_{N-1})$, where B_i is the bipolar value of the i th bit. This function can be expanded in a series:

$$\begin{aligned} f(B_0, B_1, \dots, B_{N-1}) &= \sum_k f_1(k)B_k + \sum_{k_1} \sum_{k_2} f_2(k_1, k_2)B_{k_1}B_{k_2} + \dots \\ &+ \sum_{k_1} \dots \sum_{k_j} f_{k_1 \dots k_j} B_{k_1} \dots B_{k_j} \\ &+ f_N(k)B_0B_1 \dots B_{N-1} \end{aligned} \quad (5.1)$$

where it is recognized that the series must truncate because the bits are bipolar unit binary variables. Because of the truncation of the series in (5.1), a nonlinear echo canceller can be built with only a finite number of elements that will handle any echo. In fact, there are 2^N coef-

ficients in the expansion in (5.1). It should also be noted that (5.1) is multilinear in the bits.

The idea of representing any nonlinear function of a finite number of bits by a finite expansion was exploited in an interesting study of trellis codes by R. Calderbank and J. E. Mazo [226].

To speed up the adaptation process in nonlinear cancellation, J. F. Michaelides and P. Kabal investigated lookup structures, Volterra series, and nonlinear compensators for specific channel models [227]. N. Holte and S. Studeflotten considered lookup structures [228].

Because of the need for rapid adaptation and the presence of long tails on echoes, D. W. Lin investigated mean square, rather than zero forcing, echo cancellation [229]. For similar reasons, J. M. Cioffi and T. Kailath investigated recursive structures [230]. A similar approach was studied earlier in connection with voiceband modems by M. L. Honig [231].

Connected with the issue of the high levels of echo cancellation required is the issue of conversion of the incoming analog data signal to digital form, i.e., quantization. Conventional means of A/D conversion cannot yield the high dynamic range required with a structure that is economical enough for HDSL applications. Fortunately, oversampled delta modulation techniques [232], [233] appear to be capable of achieving the required fineness of quantization in an economical way. Although this technology is still evolving, it appears to meet all the requirements of the HDSL's.

As mentioned earlier, the technology that was employed for the implementation of ISDN basic access will not be quite adequate for the HDSL's. One way to improve the performance of the receiver, at a cost of increased complexity, is to use fractionally spaced equalization (FSE), as pointed out by R. D. Gitlin and S. B. Weinstein [234]. The timing jitter insensitivity of FSE was expounded on further by D. Saha and S. Pasupathy [235]. The theory of FSE continues to evolve, as indicated by the application of the Zak transform and, consequently, wavelet representation, of the transmitted signal to FSE [236]. One of the issues in FSE implementation is stable adaptation of the tap weights in the equalizer. An optimizing algorithm for stable tap weight adaptation was given by T. Ugematsu, T. Iida, and K. Sakinawa [237].

Structured channel signaling, vector coding, and multitone all benefit from block DFE. Other techniques would also benefit from block processing [238], [239]. Related work is reported in [240].

VI. CONCLUDING DISCUSSION

At the time of writing this review, the HDSL architecture that would best serve the requirements of highly reliable DS1 rate access had not been determined. However, it is clear that an HDSL can be produced that is economically and technically viable. This is a substantial advance from 1200-baud voiceband modems that are commonplace today. However, as this paper indicates,

recent advances in VLSI technology and signal processing techniques add substantial and economical capability to the wire-pair telephone loop plant. In fact, it has been shown that the plant is not limited to two-pair duplex transmission at the DS1 rate. This suggests that other transmission configurations should be explored.

There are many possible services that might be offered that would require greater information rates from the central office to the remote distribution point than toward the CO. For example, it is frequently true that database users require great amounts of data to be transmitted to them while they transmit relatively little information, representing database changes and updates, in the reverse direction. Catalog video sales would require the transmission of dynamic video information toward the consumer with only selection and charging information transmitted in the reverse direction. Of course, entertainment video services are of this nature. Because of these potential applications, it is natural to inquire into the asymmetrical transmission capabilities of the wire-pair loop plant.

If digital transmission took place in only one direction in a frequency band, there would be no NEXT in that band. In this event, the controlling impairments to reliable transmission would be NEXT from other types of systems in the same cable sheath, impulse noise, Johnson noise in the cable pair and receiver, and FEXT between the asymmetrical channels.

If the frequency band for the asymmetrical channel is a passband with its lower band-edge substantially above the ISDN basic access band, crosstalk from basic access and voice-frequency channels would be avoided. So also would most impulse noise, which tends to have most of its energy concentrated below 100 kHz. Equal-level FEXT losses are far greater than NEXT losses, so that FEXT would not be a serious problem. Crosstalk from HDSL's would be mitigated by the fact that there would only be partial overlap of the HDSL and asymmetrical channel frequency bands. Similarly, if the upper edge of the asymmetrical channel band were below, (say) 400 kHz, interference from conventional DS1 service and digital loop carrier systems would be minimized. Thus, a 64QAM system could probably operate in the band from 100 to 367 kHz with minimal impairments (other than intersymbol interference and system loss). The greatest resistance from all the reference loops for ISDN basic access studies was 1100 Ω . The midband ($100 + 367/2 = 233$ kHz) loss of a 26 gauge loop with a resistance of 1100 Ω is 63 dB. This range could probably be achieved with a satisfactory error ratio with the transmission of 6 V pulses and by using modern coding and signal processing techniques. Consequently, an asymmetrical digital subscriber line (ADSL), operating at the DS1 rate from the central office to the customer, appears quite feasible in the entire nonloaded loop plant. Such an ADSL could operate on the same loop that provides ISDN basic access, with the "D" channel of the basic access system used for signaling and control of the ADSL. Alternatively, a 16 kb/s duplex signaling and control channel for

the ADSL could be inserted between the voiceband and the ADSL when the ADSL is provided on a POTS loop. When the ADSL was not in use, the 16 kb/s signaling and control channel could be used for data transmission. Thus, the ADSL appears an attractive possibility in the near future.

There is also interest in duplex transmission in the 10 Mb/s region over relatively short distances for LAN applications. Although these applications, call them very high bit rate DSL's (VDSL), are reasonable, little is known about the performance of the distribution cable plant, central office wiring, and customer premises wiring in this frequency band. Of particular concern is the increased possibility of interaction crosstalk, which is negligible at lower frequencies. Interaction crosstalk is coupling between two pairs that involves other tertiary pairs in the coupling mechanism. Interaction crosstalk loss typically drops off faster as frequency increases than NEXT loss does. Also of concern at these high frequencies is apparatus crosstalk, which takes place at terminals, splices, and cross-connects.

The possibilities described above are exciting. However, they are applicable only in the period of transition to a preponderantly fiber distribution plant. Although fiber will eventually replace copper, the latter provides a synergy that stimulates the demand for higher-rate services by making their provision more economical in the segment of the access network that is contiguous to the customer.

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