Performance Evaluation of a Multichannel Transceiver System for ADSL and VHDSL Services

Peter S. Chow, *Student Member, IEEE,* Jerry C. Tu, *Student Member, IEEE,* and John M. Cioffi, *Senior Member, IEEE*

***Abstract*—We study the performance of a multichannel modulation method for two contemplated subscriber line data services known as asymmetric digital subscriber lines (ADSL) and very high-speed digital subscriber lines (VHDSL). In the ADSL case, we find that over all unloaded North American subscriber lines in our test set, an unidirectional 1.536 Mb/s data rate service from the end office to the customer premises is possible on a single twisted pair at an error rate of 10^{-7} with at least a 6 dB margin using coded multichannel modulation with sufficient transmit power. Furthermore, we find that the proposed ADSL service can co-exist with basic-rate access ISDN (or voiceband analog services) on the same twisted pair with our proposed system. In the VHDSL case, data rates in excess of 100 Mb/s can be transmitted reliably, at an error rate of 10^{-7}, using uncoded multichannel modulation on a single twisted pair over a relatively short distance (≤ 150 feet) with a sufficiently high sampling rate (≈ 24 MHz) and transmit power. In this study, the dominant line impairments in the ADSL environment include intersymbol interference (ISI), far-end crosstalk (FEXT) from other ADSL services, spill-over near-end crosstalk (NEXT) from baseband services in the same wire bundle, spill-over far-end baseband signal on the same twisted pair due to imperfect filtering, and additive white Gaussian noise (AWGN) from such sources as electronic and thermal noises. In the VHDSL environment, ISI, FEXT and NEXT from other VHDSL services in the same wire bundle, as well as AWGN, are included. Finally, we show that a cost-effective multichannel transceiver design that has been suggested for high-speed digital subscriber lines (HDSL) service will also work well for the proposed ADSL and VHDSL services with only minimal modifications.**

I. Introduction

WITH the advent of high-speed digital subscriber lines (HDSL) technology in recent years, a number of new high-speed transport concepts have been proposed in the communications industry. Among these newly proposed transport concepts, asymmetric digital subscriber lines (ADSL) and very high-speed digital subscriber lines (VHDSL) are two of the most promising future data transmission services in today's market. In this paper, our emphasis is on the description and performance evaluation of one type of HDSL transceiver technology that can be used to achieve reliable data transmission for both ADSL and VHDSL services.

The proposed ADSL service will support a 1.536 Mb/s data rate on standard twisted-pair telephone lines, unidirectionally, from the Central Office to the customer premises. The term "asymmetric" in ADSL refers to a high data rate in one direction only, as ADSL is distinguished from HDSL, the latter of which is bidirectional and will only be offered for the restricted set of loops within the so-called carrier serving area (CSA)[1]. Furthermore, HDSL is intended for conventional T1 or DS1 data rate services, while ADSL is a consumer service with the intended application being the transmission of compressed TV-quality video (see [2]) with distribution over almost the entire loop plant, including those outside of the CSA. Lastly, ADSL is also currently being considered, for economical reasons, to be superimposed on the same single twisted pair that delivers basic-rate access ISDN service [or possibly the plain-old telephone service (POTS)] while HDSL will be a substitute service for basic-rate access ISDN or POTS.

Because of the asymmetry in the transmission system for ADSL, near-end crosstalk (NEXT) from one ADSL service to another cannot occur, and we will show that in the ADSL environment, the dominant line impairments are intersymbol interference (ISI), far-end crosstalk (FEXT) from other ADSL services, spill-over near-end crosstalk (NEXT) from baseband services, spill-over far-end baseband signal on the same twisted pair, and additive white Gaussian noise (AWGN) from such sources as electronic and thermal noises. The absence of near-end crosstalk due to other ADSL services in the same wire bundle significantly improves the data transport capability of the twisted pair and allows 1.536 Mb/s transmission on all loops in our test set with adequate margin when a sufficiently powerful transceiver is used.

VHDSL, on the other hand, is a high-speed data transport concept that serves as a component of the eventual

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The authors are with the Information Systems Laboratory, Department of Electrical Engineering, Stanford University, Stanford, CA 94305.

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[1]The CSA consists mainly of loops that are less than two miles in length (see Bellcore RFI 90-03 [1]).

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migration to fiber within the loop plant. VHDSL considers full-duplex transmission of data at rates significantly higher than the HDSL rate but over only that segment of the loop plant located between the pedestal and the customer premises, and it presumes a high-speed media (such as fiber) from the end office to the pedestal. Data rates of interest include those sufficiently high to sustain computer network applications, such as 8, 10, and 16 Mb/s, the T3 rate of 44.736 Mb/s, the OC-1 rate of 51.84 Mb/s, the FDDI rate of 100 Mb/s, and the OC-3 rate of 155.52 Mb/s.

Typical VHDSL channels are short drops (≤ 150 feet) of unshielded twisted pair (26 gauge or better) from the pedestal to the customer premises. There are usually two such pairs into the average customer premises, and these two pairs can crosstalk into one another if both are used. ISI on such channels is not as severe as those encountered in HDSL or ADSL but is still significant at the data rates of interest, and we find near-end crosstalk (NEXT) to be more limiting than far-end crosstalk (FEXT) when both pairs are used.

In this paper, we focus on multichannel modulation and show that it is an excellent method for delivering reliable high data rates to the customer, both in terms of performance and cost, for ADSL and VHDSL. In Section II, we describe the loop plant characteristics for both the ADSL and the VHDSL environments. We also describe the provisions made for possible co-existence of ADSL with other analog and digital baseband data services and the inclusion of a reverse channel capability in ADSL. In Section III, we briefly review the characteristics of a multichannel modulation system that is intended for use with ADSL as well as VHDSL. In particular, we use discrete Fourier transform in the transceiver design with the goal of maximizing performance and minimizing computational requirements. In Section IV, we investigate achievable data rates both within and outside of the CSA for ADSL and find that all loops in our test set will reliably support ADSL service with at least a 6 dB margin using our specific version of coded multichannel transceiver. In Section V, we investigate achievable data rates for VHDSL with varying system parameters, and we find that with a very high signaling rate (24 MHz) and sufficient transmit power, we can reliably transmit over 100 Mb/s through VHDSL lines even without an additional trellis code. Finally, we summarize our findings in Section VI.

II. Loops and Signal Characteristics

A. ADSL Transmission Characteristics

The proposed ADSL service is closely related to HDSL; however, unlike HDSL, ADSL provides *unidirectional* high data rate service to consumers within the carrier serving area (CSA) and possibly outside of the CSA. At the present time, there are no established design rules for ADSL, though several key characteristics of the ADSL environment have been identified as follows [3]:

1) All loops are nonloaded.

2) All loops consist of 26 gauge or coarser cables, either used alone or in combination with other gauge cables.

3) The maximum allowable loop length, including bridged taps, is 18 kft.

4) A reverse channel with a maximum data rate of 9.6 kb/s is to be provided for control purposes.

In addition, it would be economically advantageous if the ADSL service can be superimposed on the same line that delivers basic-rate access ISDN or analog baseband services, e.g., POTS. For this reason, though it is not a formal design rule for ADSL, we will evaluate a multicarrier system that delivers the unidirectional 1.536 Mb/s data rate without occupying or interfering the lower 50 kHz of the frequency spectrum. Furthermore, an important design objective for the authors is that the existing HDSL transceivers should be easily modified to handle ADSL service, e.g., via simple software changes, so that minimal design impact is to be incurred while implementing ADSL as a subset of HDSL.

B. Representative Loops in the ADSL Environment

The set of loops under study is shown in Fig. 1. They are representative of "lossy" loops within and outside of the CSA [4], [5]. The impulse response and power spectral density[2] characteristics for these channels have been determined with data in [6], using a modified version of the LINEMOD program[3]. A pole-zero model is added to the response of each loop to eliminate the dc component and to simulate the effects of the transformer coupling that exists at both ends of the twisted pair. This pole-zero model consists of a double-zero at dc and a double-pole that makes the power gain of the transformer equals to -6 dB at 300 Hz.

We will find, as a general result, that channels outside of the CSA perform significantly worse than those within the CSA. This is to be expected as channels outside of the CSA suffer much larger signal attenuation because they are longer in length. Cable attenuation characteristics of various channels in the frequency domain are illustrated in Fig. 2, which compares the power spectral densities of loops outside of CSA to one of the "worst-case" CSA channels (channel 6). We should point out that these power spectral densities are calculated based on source-to-load loss. Therefore, power spectral densities of the same loops calculated based on insertion loss should be scaled 6 dB higher uniformly across the entire frequency band, assuming that the source and load impedances are matched. We will use these power spectral densities based on source-to-load loss in our computer analysis for achievable throughput of our proposed system, which implies that we have tacitly included a 6 dB noise margin

[2]We define the "power spectral density" of a channel as the magnitude squared of the Fourier transform of the channel impulse response.

[3]We thank Prof. D. G. Messerschmitt of the University of California at Berkeley for making the source code of this program available to us.

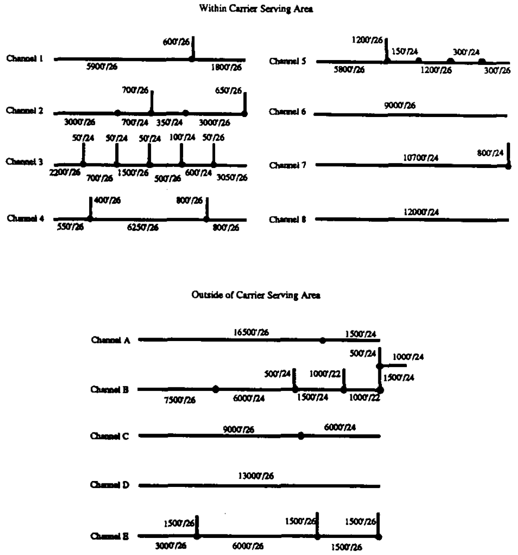

Fig. 1. ADSL loops under study within and outside of the CSA.

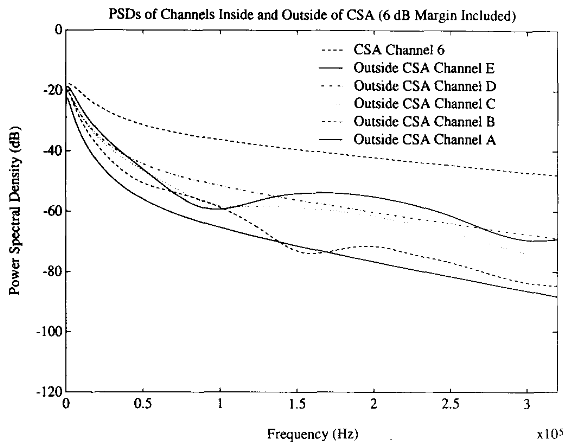

Fig. 2. Power spectral densities of representative ADSL loops under study.

(as is now required in HDSL applications) in any environment where the dominant source of noise does not depend on the channel characteristics. This is the case for HDSL, where the channel-independent near-end crosstalk is the dominant source of noise [7], and we will show in Section V-B that this is also true in the VHDSL environment. In the case of ADSL, however, the dominant source of impairment in most of our test loops turns out to be the channel-dependent far-end crosstalk, which is proportional to the channel transfer function. Thus, the factor of 6 dB margin does not apply. We will describe the various channel impairments for both ADSL and VHDSL in more detail in Sections II-C and II-F, respectively, and we will evaluate the consequences of including a 6 dB noise margin explicitly on the ADSL system in Section IV-B.

C. ADSL Line Impairments

Empirical studies indicate that crosstalk is likely to be the limiting impairment on two-way transmission at the ISDN frequencies of interest. It has been shown that the crosstalk phenomenon can be accurately modeled using only two terms, namely the near-end crosstalk (NEXT) and the far-end crosstalk (FEXT) [8]. The NEXT and FEXT terms are identified in Fig. 3. In the ADSL environment, data is transmitted unidirectionally. Therefore there will be no NEXT term in the frequencies of interest due to other ADSL services in the same wire bundle. (We have also assumed that there will be no T1 or HDSL systems in operation within the same wire bundle, for they would be extremely damaging to an ADSL system due to their large high frequency content.) However, there will be a significant amount of spill-over NEXT from bidirectional baseband services in the same wire bundle due to nonideal filtering. This NEXT term can be modeled with a coupling function of the form [8]:

$$|H_{\text{NEXT}}(f)|^2 = K_{\text{NEXT}} f^{3/2} \qquad (1)$$

where f is frequency in Hz and K_{NEXT} is determined through empirical measurement. In our study, we will assume the worst-case scenario of 49 crosstalkers all due to basic-rate access ISDN service, where $K_{\text{NEXT}} \approx 10^{-13}$ [9], [10]. The input power spectrum to this coupling function is the transmitted spectrum of the baseband service. For basic-rate access ISDN service over AT&T DSL's, square pulses are passed through a second-order lowpass filter with the 3 dB point at 40 kHz and at least a 50 dB/decade rolloff after 50 kHz [11]. The actual signal voltages used are ± 2.5 volts and $\pm\frac{5}{6}$ volt, leading to an average power of approximately 25 mW with a 135 Ω load impedance. We will use this transmit spectrum and power level for BA-ISDN in our computer evaluation, though this is not the worst-case transmit spectrum. In particular, if sinc pulses are used instead of square pulses, we would have a flat instead of sinc squared spectrum.

In addition to spill-over NEXT, FEXT also exists in the ADSL environment, and it can be modeled with a coupling transfer function of the form [12]:

$$|H_{\text{FEXT}}(d, f)|^2 = K_{\text{FEXT}} d |C(f)|^2 f^2 \qquad (2)$$

where $|C(f)|^2$ is the channel power spectral density function as plotted in Fig. 2, d is the length of the cable in kft, f is frequency in Hz, and K_{FEXT} is determined through empirical measurement. In our study, we will mostly assume that $K_{\text{FEXT}} \approx 10^{-16}$ [13]-[15], though we will also investigate the effect of varying the level of K_{FEXT} at saturation power level in Section IV-B.

In addition to crosstalk noise, there are several other impairments that will degrade the performance of any system operating over twisted copper pairs at ADSL rates. Because we have assumed that the ADSL service is superimposed on the same wire that delivers baseband BA-ISDN, the spill-over far-end signal from BA-ISDN on the same wire will be received along with the ADSL signal,

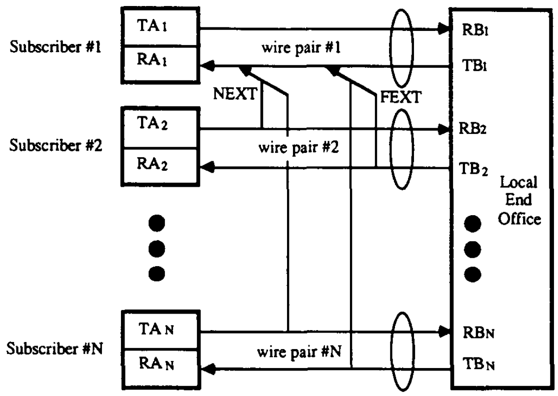

Fig. 3. Near-end and far-end crosstalk.

and this spill-over "signal" must be treated as noise by the ADSL receiver. Electronic noise, including quantization noise from the A/D converter, and thermal noise in the analog portion of the receiver can be modeled as additive white Gaussian noise (AWGN). We will assume a fixed level of AWGN at -140 dBm/Hz in our study, which represents a fairly conservative estimate [5]. Inductive noise at 60 Hz and its harmonics generally do not cause a problem at ADSL rates [1]; therefore, it will be ignored in our study. Residual echo noise is also negligible since there will be no echo due to ADSL at the receiver end as data is transmitted unidirectionally, and residual echo from baseband service is assumed to be sufficiently cancelled and filtered out. Impulse noise caused by switching transients, lightning, and electrical machinery is less well understood, and usually an operational margin of 6–12 dB is placed on a system to handle such infrequent but high-peak noise. We shall not include impulse noise in our model. However, the block processing nature of our multicarrier system is advantageous when dealing with this type of short but intense noise, since the noise energy contained in the impulse is effectively spread over the entire block and the blocklength used is typically in the order of several hundred times that of the duration of the impulse (see [7]). Impulse noise of long duration, on the other hand, is hopefully eliminated by the high-pass filter in ADSL. Lastly, intersymbol interference (ISI) is inherent in ADSL loops since the transfer characteristics of the channels are nonideal. We shall describe how ISI can be mitigated using a multicarrier approach in Section III.

D. VHDSL Transmission Characteristics

The proposed VHDSL service is an enhancement [16] of the presently developing HDSL service. VHDSL provides reliable, bidirectional data transmission at rates of 10 Mb/s or higher over only relatively short distances. Loops intended for use with VHDSL service are those that are located between the pedestal and the customer premises. These are generally no more than 150 feet in length and no smaller than 26 gauge in size. As in the case of ADSL, VHDSL is a new transport concept with no established design rules yet. Some of the target data rates that may be desirable for VHDSL applications include:

- Current Ethernet standard rate = 10 Mb/s
- DS-3 (digital signal, level 3) = T3 rate = 44.736 Mb/s
- OC-1 (optical carrier, level 1) = 51.84 Mb/s
- FDDI rate = 100 Mb/s
- OC-3 (optical carrier, level 3) = 155.52 Mb/s

E. Representative Loop in the VHDSL Environment

In this study, we only consider the worst-case VHDSL loop, i.e., 150 feet of 26 gauge wire. Furthermore, we assume that PIC cables operating at 70°F with matched source and load resistances of 110 Ω are used and that within this short 150 feet line there are no bridged taps or wire gauge changes. As in ADSL loops, a transformer is added to both ends of the cable to eliminate the dc component in the frequency response. Again, the effects of the transformer coupling is simulated by a pole-zero model that consists of a double-zero at dc and a double-pole that makes the power gain of the tranformer equal to -6 dB at 300 Hz. Using the modified version of LINEMOD with data in [6], we can determine the impulse response and power spectral density characteristics for this channel. We found that the maximum useful bandwidth of this worst-case loop is around 12 MHz (see Fig. 4). Thus, a maximum signaling rate of 24 MHz will be used in this study.

F. VHDSL Line Impairments

As in the case of ADSL, NEXT, and FEXT are two of the most serious line impairments encountered in VHDSL. The coupling function for NEXT is given by (1) in Section II-C. In [9], Lin estimated that $K_{\text{NEXT}} \approx 10^{-13}$ for the 49-crosstalker case from data in a Bellcore Technical Reference [10]. In the case of VHDSL, however, the twisted pairs are most likely unbundled. Therefore, instead of 49-crosstalkers, there will usually be a maximum of only one crosstalker, as existing customer lines often contain two twisted pairs into each customer premises. We will assume that $K_{\text{NEXT}} \approx \frac{1}{50} \times 10^{-13} = 2 \times 10^{-15}$ for our test loop. Note that if we assume there are normally only 6 dominant crosstalkers in a 50-pair bundle, then the VHDSL K_{NEXT} with one crosstalker should be closer to $\frac{1}{6} \times 10^{-13}$. We will investigate the effect of varying the level of K_{NEXT} in Section V-B. The coupling function for FEXT is given by (2) in Section II-C. As in the case of K_{NEXT}, K_{FEXT} is determined through empirical measurement [13]–[15]. In this study, we will assume that $K_{\text{FEXT}} \times d \approx \frac{1}{50} \times 0.15 \times 10^{-16} = 3 \times 10^{-19}$ for the test loop, since there will only be one far-end crosstalker instead of 49 and $d = 150$ ft $= 0.15$ kft. The performance of DMT for K_{FEXT} varying over several orders of magnitude are discussed in Section V-B. Besides NEXT and FEXT, the transmitted data may also be corrupted by interaction crosstalk and apparatus crosstalk at VHDSL fre-

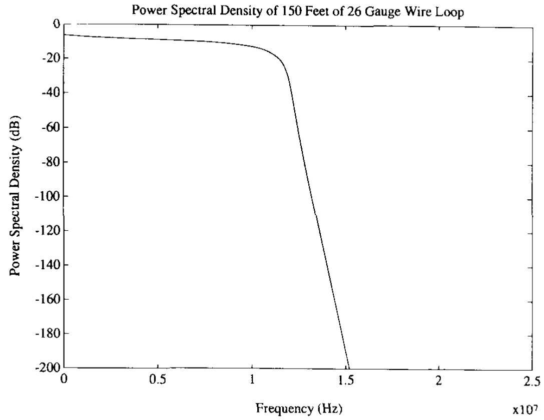

Fig. 4. Power spectral density for 150 feet of 26 gauge copper twisted pair loop.

quencies. However, the specific effects of interaction crosstalk, which is the coupling between two pairs involving other pairs, and apparatus crosstalk, which takes place at terminals, splices, and cross-connects, are not well understood [17]. Therefore, they will not be included here.

In addition to various forms of crosstalk, there are several other potential impairments that will degrade the performance of a VHDSL system. Residual echo noise does exist in VHDSL loops, and it is typically modeled as additive white Gaussian noise (AWGN). Assuming an input signal power of around 20 mW, a typical VHDSL loop will attenuate the signal by approximately 10 dB, resulting in a total received signal power of 3 dBm. With an excellent echo canceller that can reduce the echo to a level of 40 dB below the received signal, we expect a total noise power of around -37 dBm across a two-sided bandwidth of 24 MHz, yielding a noise power spectral density of approximately -110 dBm/Hz. Therefore, we will fix the level of AWGN due to residual echo at a constant level of -110 dBm/Hz throughout this study. Other sources of AWGN noise, such as electronic noise and thermal noise, are less damaging. We will assume a fixed level of electronic noise at -140 dBm/Hz in our study [5], which represents a fairly conservative estimate and can be ignored since it is much lower than the AWGN level of the residual echo noise. The effects of inductive noise and impulse noise in VHDSL are similar to those in ADSL, and lastly, ISI can be mitigated by using a multicarrier modulation approach described in Section III.

III. Multichannel Modulation

Recently, various multicarrier systems [18]–[21] have been proposed to transmit data reliably in the presence of severe intersymbol interference (ISI) for digital subscriber line applications. In this study, we focus on a specific implementation of multicarrier modulation, known as the discrete multitone (DMT) modulation [21].

A. *The Discrete Multitone System*

The fundamental concept of multicarrier modulation is the conversion of a data transmission channel with intersymbol interference (ISI), and possibly crosstalk and/or colored noise, into a set of parallel, independent, and ISI-free subchannels. In [20], Bingham gives a comprehensive tutorial on the various multicarrier modulation methods, and the specific modulation technique evaluated in this study is known as the discrete multitone (DMT) modulation. DMT modulation is based on frequency-division partitioning of the channel spectrum using the discrete Fourier transform, and it is an enhanced version of what appears in [21]. We shall not attempt a detailed description of the DMT modulation in this section, but rather we refer readers to the companion paper [7] by J. S. Chow and two of the co-authors of this paper on the performance evaluation of the DMT system for high-speed digital subscriber lines (HDSL) in this same issue.

B. *Trellis Code Concatenation*

One attractive feature of multicarrier modulation is that since the independent subchannels are memoryless, coset codes (trellis codes) can be concatenated to the modulation structure, and the coding gain of these powerful codes can be realized by the system. In [22] and [23], Forney characterized most of the known good codes for bandlimited channels as coset codes. In the DMT system studied here, we will use a method of code concatenation described in [24] known as "coding down the block," in an effort to achieve a good tradeoff between hardware complexity and decoding latency. This method was also inherent in [25] and in a recently filed patent application by Decker of Telebit [26]. In this method, a single encoder/decoder pair is used and several coded symbols from the encoder are concatenated together to form the multidimensional transmit vector that is to be processed by the multicarrier modulator. Each orthogonal dimension of the transmit vector is assigned to succeeding subchannels of the multicarrier system until all subchannels are used. The details of the coset code concatenation procedure are given in [24].

C. *Channel Identification and Protocol*

An accurate estimate of the channel response is necessary in discrete multitone modulation, and this information must be available to both the transmitter and the receiver in order for the bit allocation algorithm and the pole-cancelling filter to function properly. The channel identification process takes place during the initial startup procedure. We will now briefly summarize the general protocol used by the DMT for ADSL and VHDSL services.

After receiving a request for transmission signal from one subscriber, predefined test patterns are transmitted to verify that the channel is indeed operational. Timing and synchronization are established at this time. Then a predefined pseudorandom training sequence is sent from the transmitter for channel identification. An ARMA (autoregressive/moving average, or pole-zero) model for the channel is derived from the received training sequence via a least squares fit to minimize the prediction error between the ARMA model and the received channel outputs. The receiver can now set the coefficients for the pole-cancelling filter, compute the SNR for each frequency-indexed subchannel, and perform the bit allocation algorithm. The bit assignments are then sent back to the transmitter via a reliable feedback channel, which can be one of the subchannels operating in the reverse direction using highly redundant repetition code to ensure reliability. After frame synchronization is established, the DMT transceiver system is ready for user data transmission. In the ADSL environment, it should be stated that the information sent from the customer to the Central Office during this startup period will not be in the lower 50 kHz of the frequency band; thus, it will not interfere with existing baseband services on the loop. This requires the use of at least one subchannel in the opposite direction, furthermore, this subchannel is silent in the forward direction. For our proposed system, we will use two of our frequency bins, occupying the frequency spectrum from 50 to 52.5 kHz, for this purpose, which will also serve as a reverse channel for system control and monitoring functions during normal operation.

IV. Performance Evaluation for ADSL

In this section, we analyze the performance of the DMT transceiver for ADSL. The method of computer evaluation is given in [7], and we shall not repeat it here.

A. System Parameters

The system that we evaluate has a blocklength of 512 and a signaling rate of 640 kHz. These system parameters are only representative and can be changed easily. In particular, we note that a longer blocklength (say 1024) or a higher signaling rate (for example, 800 kHz) will yield somewhat higher data rates at the cost of longer decoding delay and higher complexity. In [27], Aslanis and Cioffi showed the capacity of loops inside the CSA, subject to a chosen signaling rate, will continue to increase until the signaling rate approaches 600–800 kHz, and we based our choice of signaling rate on this result. However, that study included NEXT from HDSL, whereas in ADSL we need not consider this impairment. Thus, higher sampling rates could yield yet better performance. Our choice of blocklength, on the other hand, reflects the current technology in terms of feasible computational complexity and cost, which will be discussed briefly in Section IV-C.

B. Performance Results

The performance of our proposed DMT transceiver in the ADSL environment is evaluated as a function of transmit power. Using the DMT's frequency-specific bit allocation algorithm, we can shape the transmitted spectrum for the ADSL environment. We show in Fig. 5 an example of bit allocation for CSA channel 6 at the maximum achievable data rate (>1.536 Mb/s) using 20 mW of transmit power with BER = 10^{-7} and explicitly allowing a 12 dB noise margin. We note that the lower 50 kHz of the frequency band has been left free of transmission for spectral compatibility with existing baseband services, such as basic-rate access ISDN or voiceband POTS. Also note that the actual numbers of bits assigned to frequency bins between 50 and 75 kHz are smaller than those for bins immediately after 75 kHz, since spill-over noise from baseband services is extremely damaging in this frequency range.

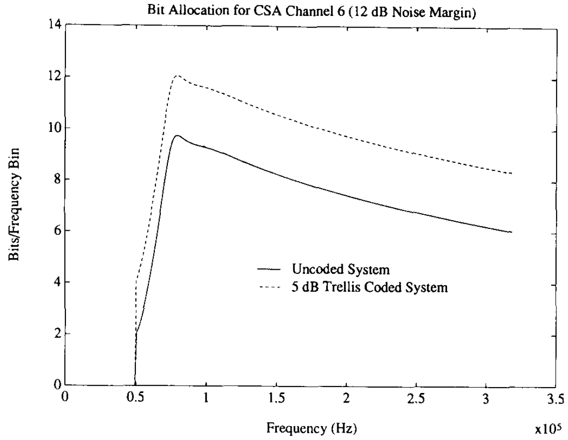

Fig. 5. Uncoded and coded bit spectral efficiency for CSA channel 6 with 20 mW of power and 12 dB noise margin.

We evaluated the system for transmit power level ranging from 1 to 200 mW (0–23 dBm) for both CSA loops and loops outside of CSA. The achievable data rates at a bit error rate (BER) of 10^{-7} are shown in Figs. 6 and 7. We next investigated the performance of the DMT system with a concatenated trellis code. We assumed the coding gain achievable using an 8-dimensional, 64-state trellis code, i.e., approximately 5 dB of coding gain. The resulting data rates are shown in Figs. 8 and 9. Since the exact transmission impairments over digital subscriber lines are not completely understood at this time, an operational margin of 6 dB or more is typically required for a system to handle any unexpected source of noise. In Figs. 10 and 11, we have plotted the achievable data rates of a coded DMT system with a 6 dB operational margin. Our evaluation shows that all channels in our test set can support 1.536 Mb/s data rate transmission in the ADSL environment using a DMT transceiver with reasonable transmit power. In fact, using 10 mW of power, we can transmit over 3.2 Mb/s over the worst CSA channels using a coded DMT system (Fig. 8). The performance level can be further improved if a steeper lowpass filter is used in the BA-ISDN service, say a lowpass filter with a 70 dB/decade rolloff instead of the assumed 50 dB/decade rolloff, to isolate the baseband signal from the passband ADSL service.

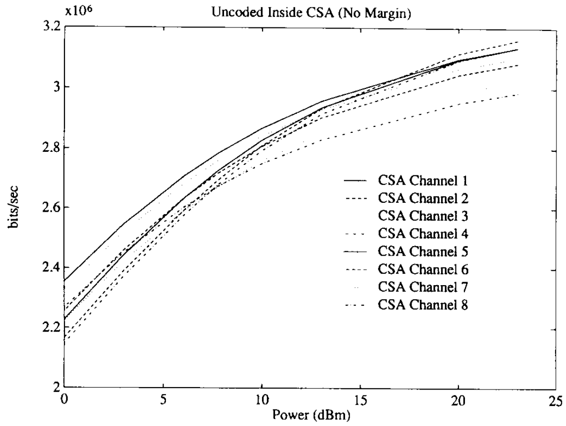

Fig. 6. Uncoded data rates for CSA channels as a function of transmit power.

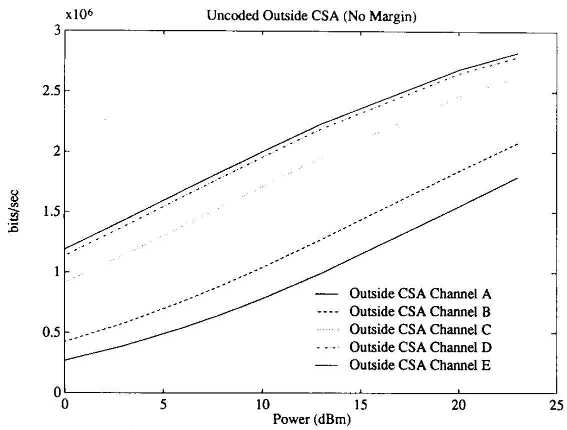

Fig. 7. Uncoded data rates for channels outside of CSA as a function of transmit power.

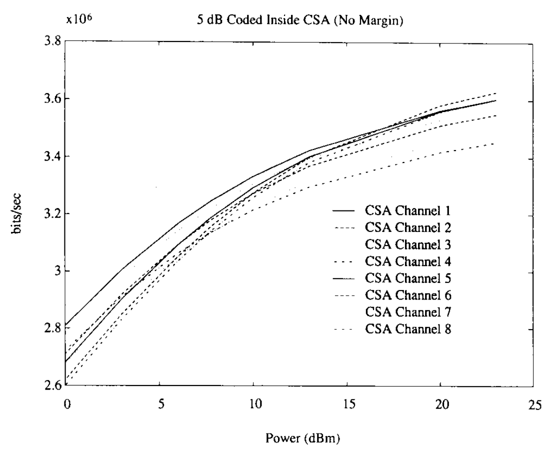

Fig. 8. Coded data rates for CSA channels as a function of transmit power.

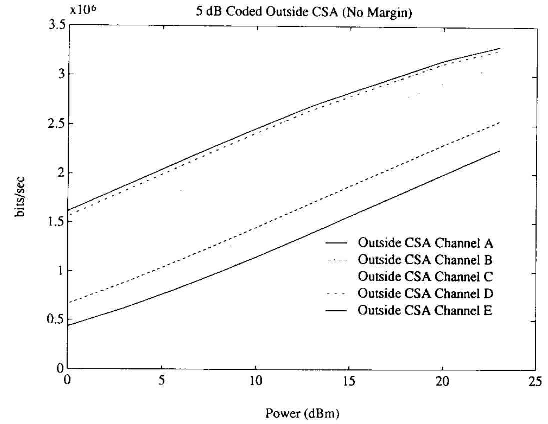

Fig. 9. Coded data rates for channels outside of CSA as a function of transmit power.

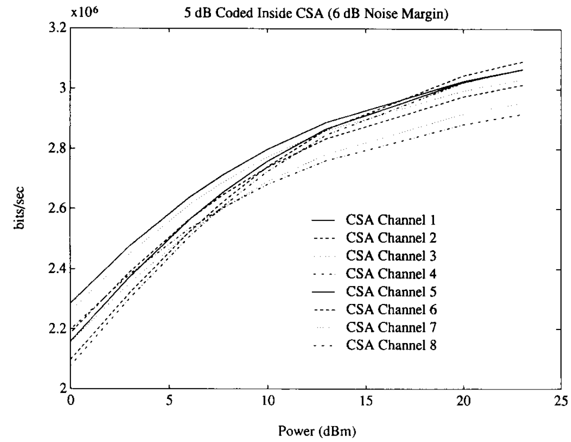

Fig. 10. Coded data rates for CSA channels with 6 dB noise margin.

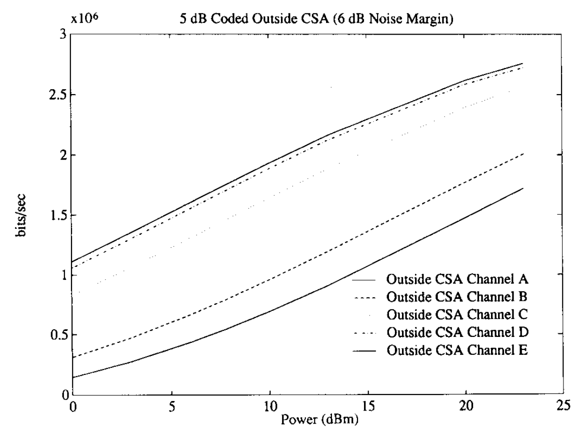

Fig. 11. Coded data rates for channels outside of CSA with 6 dB noise margin.

We note that in all cases, i.e., coded and uncoded transmissions over loops within the CSA as well as those outside of the CSA, the achievable data rates will eventually saturate, due to the presence of far-end crosstalk, no matter how much more power we use at the transmitter. Furthermore, this saturation level can be explained

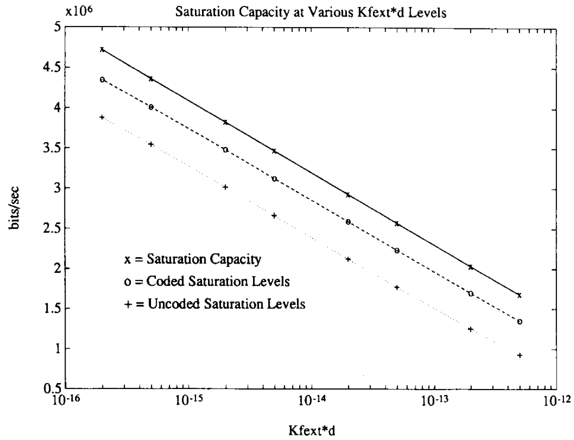

Fig. 12. Data rates at saturation power level.

and obtained from the following saturation capacity approximation, assuming that FEXT is the only dominant source of line impairment.

$$C = \sum_{k=0}^{N-1} \frac{W}{2N} \log_2 (1 + \text{SNR}_k) \tag{3}$$

$$\cong \int_{50\text{ kHz}}^{320\text{ kHz}} \log_2 (1 + \text{SNR}(f))\, df \tag{4}$$

$$= \int_{50\text{ kHz}}^{320\text{ kHz}} \log_2 \left(1 + \frac{E_x |C(f)|^2}{E_x K_{\text{FEXT}} d |C(f)|^2 f^2}\right) df \tag{5}$$

$$= \int_{50\text{ kHz}}^{320\text{ kHz}} \log_2 \left(1 + \frac{1}{(K_{\text{FEXT}} d) f^2}\right) df \tag{6}$$

We note here that this saturation level is independent of channel characteristics other than the channel length. Therefore, we can characterize the saturation level of any FEXT-dominated channel as a function of $K_{\text{FEXT}} \times d$. In Fig. 12, we have plotted the achievable data rates as a function of $K_{\text{FEXT}} \times d$ for the coded and uncoded cases as well as the theoretical channel capacity calculated using (6) above. Note that the uncoded saturation data rates differ from the coded saturation data rates, and this phenomenon is due to the fact that trellis codes effectively increase the minimum distance between signal points while not affecting the noise variance.

C. Computation Requirements

One of the advantages of our proposed DMT system over DFE-based ADSL systems is the savings in computation. A detailed computation estimate of the proposed DMT transceiver can be found in [7]. Excluding a separate programmable Viterbi decoder chip [28] and the synchronization/timing recovery circuits, approximately 20 MIPS (million instructions per second) of computing power is required to implement the receiver at a signaling rate of 640 kHz with a blocklength of 512, which represents only one (custom-designed) programmable DSP chip. Furthermore, the transmitter requires less than 10 MIPS of computing power at a signaling rate of 640 kHz with a blocklength of 512. Therefore, a custom DSP with a FFT peripheral could easily implement all the signal processing functions of both the transmitter and receiver within a single chip.

V. Performance Evaluation for VHDSL

In this section, we analyze the performance of the DMT transceiver for VHDSL in a similar manner as was performed for ADSL. The default system parameters of the transceiver under study are: blocklength of 512, transmit power of 10 dBm or 23 dBm, and signaling rate of 24 MHz. Throughout our evaluation, we study the effect of different system parameters on the achievable data rate. We varied the signaling rate from 640 kHz to 24 MHz, transmit power from 0 dBm to 30 dBm, and system blocklength from 128 to 1024. In addition, we examine the relative importance of near-end and far-end crosstalk in the VHDSL environment by varying these crosstalk effects over three orders of magnitude.

A. Performance Results

1) Signaling Rate: It is important to choose the best signaling rate to fully utilize the available bandwidth of VHDSL loops. In Section II-E, we showed that the worst-case VHDSL loop has a usable bandwidth of 12 MHz (24 MHz signaling rate), so we investigate here the effect of varying signaling rate up to 24 MHz. Fig. 13 shows the achievable throughputs for signaling rates of 640 kHz to 24 MHz. The DMT system configuration is blocklength 512 with 10 dBm transmit power. Crosstalk coupling constants are $K_{\text{NEXT}} = 2 \times 10^{-15}$ and $K_{\text{FEXT}} \times d = 3 \times 10^{-19}$. The achievable throughput ranges from about 4 Mb/s at 640 kHz signaling rate to 90 Mb/s at 24 MHz signaling rate. It is clear that, with all other system parameters constant, the choice of signaling rate is crucial to achieve the highest possible throughput. We conclude that 24 MHz signaling rate, or some convenient number close to 24 MHz, should be used to fully utilize all the available VHDSL loop bandwidth.

2) Transmit Power: We studied the effect of increasing the transmit power of the DMT transceiver, holding constant the system blocklength at 512 and VHDSL crosstalk coupling at $K_{\text{NEXT}} = 2 \times 10^{-15}$ and $K_{\text{FEXT}} \times d = 3 \times 10^{-19}$. Fig. 14 shows the achievable throughputs as a function of transmit power for various signaling rates. We see that the data rate improves with increasing power at low power levels and that the increase is negligible at levels above 20 dBm. The data rate "saturates" above this level because of crosstalk noise that depends on transmit power. Assuming that the crosstalker uses the same signaling strategy and power level as the DMT transceiver, as the transmit power increases the crosstalk noise level increases by the same proportion and the overall signal-to-noise ratio remains constant. The achievable data rate therefore saturates as the crosstalk noise level domi-

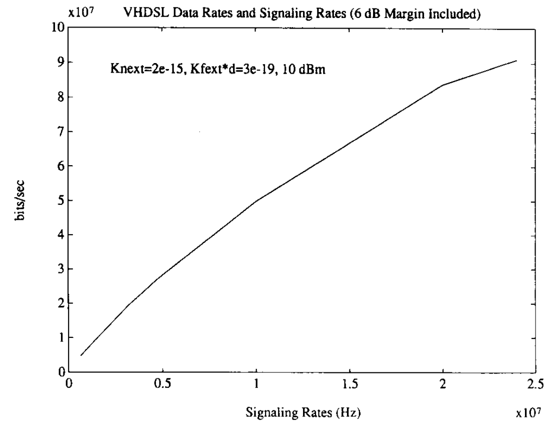

Fig. 13. DMT performance at different signaling rates.

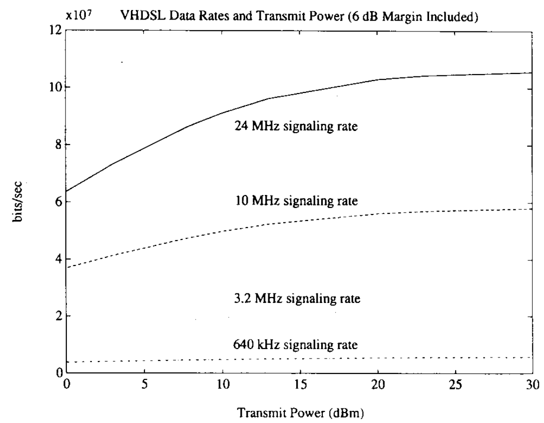

Fig. 14. DMT performance with different transmit power levels.

nates over background additive white Gaussian noise (AWGN). We see from Fig. 14 that 10–15 dBm of transmit power is sufficient to achieve the best throughputs at most signaling rates.

3) Blocklength: We studied the effect of system blocklength on achievable throughput. We found that varying the blocklength from 128 to 1024 (not including cyclic extension) had minimal effect on the data rate. Using 24 MHz signaling rate, 10 dBm transmit power, $K_{\mathrm{NEXT}} = 2 \times 10^{-15}$, and $K_{\mathrm{FEXT}} \times d = 3 \times 10^{-19}$, we found that the DMT can transmit approximately 90 Mb/s for all blocklengths from 128 to 1024. There is negligible performance improvement in going from blocklength 512 to blocklength 1024 (see Fig. 15). Using a longer blocklength incurs longer decoding delay, but it results in easier implementation as computations are distributed over a longer block interval. Also, a longer block allows the DMT system to have additional immunity to short impulsive noise spikes. Considering all of these issues, we conclude that a blocklength of 512 offers the best balance between implementation ease, immunity to impulse noise, and decoding delay.

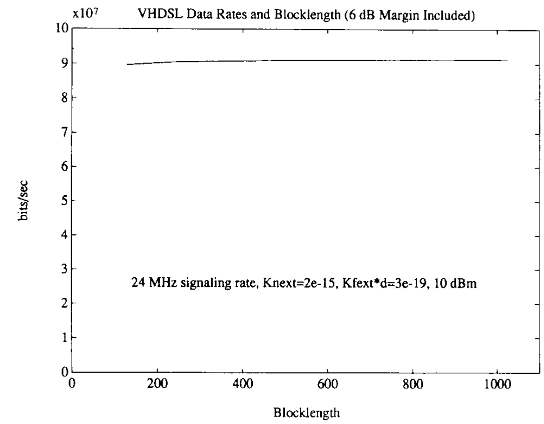

Fig. 15. DMT performance as a function system blocklength.

B. Crosstalk Coupling

In Section V-A, we suggested that crosstalk interference limits the system performance at high power levels. In this section, we examine the relative importance of NEXT and FEXT for VHDSL applications.

1) NEXT Noise: We varied the NEXT coupling coefficient over four orders of magnitude to investigate whether NEXT is the dominant noise source. Using 24 MHz signaling rate and blocklength of 512, we evaluated DMT performance as a function of K_{NEXT} for two different transmit powers. We maintained $K_{\mathrm{FEXT}} \times d$ at 3×10^{-19}. The results are shown in Fig. 16. Our results show that the achievable data rates depend significantly on the value of K_{NEXT}. With 23 dBm of transmit power, as K_{NEXT} is increased over three orders of magnitude, the achievable VHDSL data rate dropped from 140 Mb/s to 50 Mb/s. This suggests that NEXT is a significant, if not dominant, noise impairment in VHDSL at 24 MHz signaling rate.

2) FEXT Noise: We varied the factor $K_{\mathrm{FEXT}} \times d$ over four orders of magnitude, keeping K_{NEXT} constant at 2×10^{-15}. We used a DMT transceiver with 24 MHz signaling rate and blocklength 512. We studied two different transmit powers, and the results are shown in Fig. 17. We see that the achievable data rate does not decrease very much with increasing K_{FEXT}. The insensitivity of VHDSL performance as a function of $K_{\mathrm{FEXT}} \times d$ indicates that FEXT is *not* the dominating noise impairment in VHDSL applications with high signaling rate.

3) No Crosstalk: A final evaluation assumed the absence of all crosstalk noise and considered only AWGN at -110 dBm/Hz. This scenario is present if we silence one of the two twisted pairs in the VHDSL loop, or if we use coordinated transmission on the two pairs with crosstalk cancellation/equalization across the two pairs in the loop. If there is no crosstalk present, then the achievable data rates for VHDSL do not saturate with increasing transmit power, as shown in Fig. 18. Very high data rates are achievable in the complete absence of crosstalk noise (> 100 Mb/s). However, physical limitations on transmit

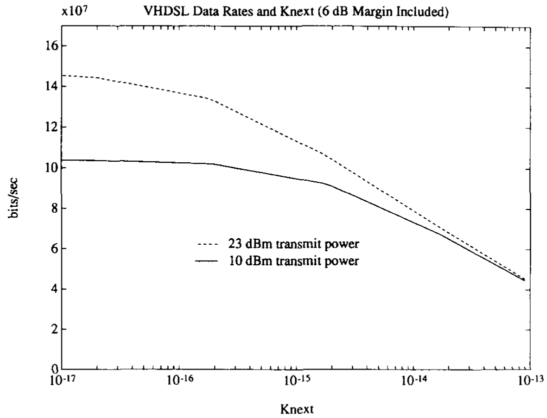

Fig. 16. Effect of K_{NEXT} on DMT performance.

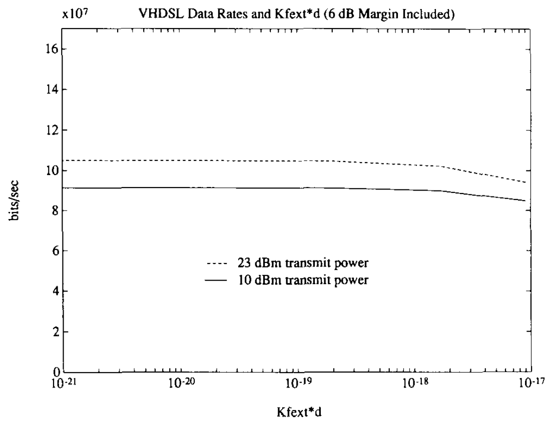

Fig. 17. Effect of K_{FEXT} on DMT performance.

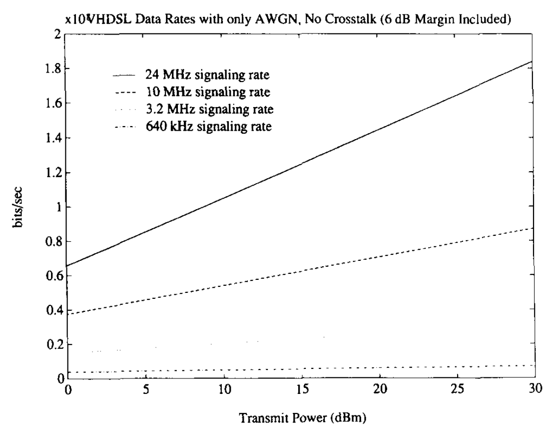

Fig. 18. DMT performance in the absence of crosstalk.

power, including power dissipation within the twisted pairs, may prevent implementing very high power transmitters.

C. Discussion of VHDSL Results

Our performance evaluation showed that high data rates (>50 MHz) are possible for the DMT transceiver in the VHDSL environment, provided that the system parameters are properly chosen. The analysis identified the most crucial design parameter to be the signaling rate, with the parameters of blocklength and transmit power level having a less significant impact on achievable throughput. Our evaluation also showed that NEXT is the dominant noise of impairment when high signaling rates are used. NEXT depends on transmit power level and causes the achievable throughput to saturate when the transmit power exceeds 20 dBm. FEXT is of secondary importance in VHDSL since the cable lengths are very short in VHDSL.

When neither NEXT or FEXT is present, our analysis predicts that very high data rates may be achieved provided that a sufficiently powerful transmitter is used. However, other noise impairments may limit system performance at these high transmit powers. We note in Fig. 18 that we can achieve over 100 Mb/s data rates while signaling at 24 MHz with 10 dBm of transmit power in the absence of crosstalk. This zero-crosstalk scenario may be achieved in two ways: silencing one pair and transmitting solely on the other twisted pair (single duplex), or using coordinated transmission on two pairs (dual duplex) along with crosstalk cancellation and/or equalization. We currently believe that the single duplex option is the more desirable alternative. Silencing one of the two pairs in a bundle eliminates the interference posed by NEXT and FEXT and allows much higher throughputs that do not saturate with increasing transmit power. Achieving over 100 Mb/s in a single duplex mode rather than dual duplex is desirable for hardware complexity/cost issues as well, since only one transceiver is required. Crosstalk cancellation or equalization across two coordinated pairs is a less desirable option. It requires crosstalk equalization that is analogous to the complex multidimensional equalization methods for adjacent channel interference. It also requires twice as much hardware—one transceiver for each of the two coordinated pairs. Finally, the VHDSL environment may not always have *two* pairs of twisted pair available from the pedestal to the customer premises for dedicated VHDSL service, making dual duplex operation less flexible and less applicable.

VI. Conclusion

In this study, we evaluated the performance of a DMT transceiver system for the newly proposed ADSL and VHDSL services. In both instances, we found that our proposed DMT system performs extremely well and is able to meet all of the recommended performance requirements at this early stage of development of these two new transport concepts.

In the case of ADSL, with the addition of a powerful 5 dB trellis code and sufficient transmit power, we found that all channels in our test set can support the 1.536 Mb/s unidirectional ADSL data rate with an error rate of 10^{-7} and a 6 dB operational margin, while maintaining complete spectral compatibility with existing baseband services such as BA–ISDN and POTS. In fact, all subscriber lines within the CSA can support 3.088 Mb/s data rate, i.e., twice the T1 rate, on a single twisted pair with an error rate of 10^{-7} using coded multichannel modulation with only 10 mW of transmit power. However, the 3.088 Mb/s data rate cannot be sustained over a single twisted pair for worst-case subscriber lines outside of the CSA.

In the case of VHDSL, our primary goals are to identify the key system parameters and study the effects of crosstalk. We found that the signaling rate is the single most important system parameter, while blocklength and transmit power level have a less significant impact on the achievable data rate. On the issue of crosstalk, we found that NEXT is more damaging than FEXT, due to the short length of the typical VHDSL line. However, if we silence one of the two twisted pairs and operate in a single duplex mode to effectively eliminate crosstalk, data rates over 100 Mb/s can be achieved with a signaling rate of 24 MHz. Of course, such a high signaling rate may be difficult to implement in practice, and other sources of noise not identified in this study may limit system performance at the high transmit power level and sampling frequency required.

Based on these performance results, the computational efficiency of the DMT system, and the additional benefit that a DMT-based HDSL transceiver [7] can be easily modified to accomodate either ADSL or VHDSL with only minor changes, we conclude that the discrete multitone system represents an excellent design choice for both the ADSL and VHDSL services.

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References

[1] Bell Communications Research, "High bit-rate digital subscriber line (HDSL) systems that support DS1 rate access in the copper loop plant," *Bellcore Request for Information*, RFI 90-03, Mar. 1990.

[2] A. Wong, C. T. Chen, D. J. Le Gall, F. C. Jeng, and K. M. Uz, "MCPIC: A video coding algorithm for transmission and storage applications," *IEEE Commun. Mag.*, vol. 28, no. 11, pp. 24–32, Nov. 1990.

[3] E. Manchester, "ADSL—A technical analysis," *U.S. West Advanced Technol. Tech. Anal.*, Jan. 1991.

[4] K. Sistanizadeh, "A tentative CSA loop population for preliminary studies of HDSL transmission schemes," T1E1.4 Tech. Subcom. Working Group Contrib., Sept. 24, 1990.

[5] J. T. Aslanis Jr, "Coding for communication channels with memory," Ph.D. dissertation, Stanford University, Stanford, CA, Nov. 1989.

[6] AT&T, "Digital data system: Channel interface specification," *Bell Syst. Tech. Ref.*, 62310, Sept. 1983.

[7] J. S. Chow, J. C. Tu, and J. M. Cioffi, "A discrete multitone transceiver system for HDSL applications," *IEEE J. Select. Areas Commun.*, this issue, pp. 895–908.

[8] D. G. Messerschmitt, "Design issues in the ISDN U-interface transceiver," *IEEE J. Select. Areas Commun.*, vol. SAC-4, no. 8, pp. 1281–1293, Nov. 1986.

[9] D. W. Lin, "Wide-band digital subscriber access with multidimensional block modulation and decision-feedback equalization," *IEEE J. Select. Areas Commun.*, vol. 7, no. 6, pp. 996–1005, Aug. 1989.

[10] Bell Communications Research, "ISDN basic access digital subscriber lines," *Bellcore Tech. Ref.*, TR-TSY-000393, May 1988.

[11] K. Kerpez, Bell Communications Research, private communication, Mar. 1991.

[12] S. V. Ahamed, P. P. Bohn, and N. L. Gottfried, "A tutorial on two-wire digital transmission in the loop plant," *IEEE Trans. Commun.*, vol. COM-29, no. 11, pp. 1554–1564, Nov. 1981.

[13] G. J. Pottie, "Transmission impairments for asymmetric digital subscriber lines," T1E1.4 Tech. Subcom. Working Group Contrib., Feb. 18, 1991.

[14] S. H. Lin, "Statistical behavior of multipair crosstalk," *Bell Syst. Tech. J.*, vol. 59, pp. 956–974, July/August 1980.

[15] D. H. Morgen, "Expected crosstalk performance of analog multichannel subscriber carrier systems," *IEEE Trans. Commun.*, vol. COM-23, no. 2, pp. 240–245, Feb. 1975.

[16] R. Hsing, Bell Communications Research, private communication, June 1990.

[17] J. W. Lechleider, "High bit rate digital subscriber lines: A review of HDSL progress," *IEEE J. Select. Areas Commun.*, this issue, pp. 769–784.

[18] I. Kalet, "The multitone channel," *IEEE Trans. Commun.*, vol. 37, no. 2, pp. 119–124, Feb. 1989.

[19] N. Zervos and I. Kalet, "Optimized decision feedback equalization versus optimized orthogonal frequency division multiplexing for high speed data transmission over the local cable network," presented at the 1989 Int. Conf. Commun., Boston, MA, June 1989.

[20] J. Bingham, "Multicarrier modulation for data transmission: An idea whose time has come," *IEEE Commun. Mag.*, vol. 28, no. 5, pp. 5–14, May 1990.

[21] A. Ruiz, J. M. Cioffi, and S. Kasturia, "Discrete multiple tone modulation with coset coding for the spectrally shaped channel," to appear in *IEEE Trans. Commun.*

[22] G. D. Forney Jr., "Coset codes I: Introduction and geometrical classification," *IEEE Trans. Inform. Theory*, vol. 34, no. 5, pp. 1123–1151, Sept. 1988.

[23] —, "Coset codes II: Binary lattices and related codes," *IEEE Trans. Inform. Theory*, vol. 34, no. 5, pp. 1152–1187, Sept. 1988.

[24] J. C. Tu and J. M. Cioffi, "On the concatenation of coset codes with multichannel modulation methods," presented at Proc. GLOBECOM '90, San Diego, CA, Dec. 1990.

[25] A. Ruiz and J. M. Cioffi, "A frequency-domain approach to combined spectral shaping and coding," presented at the 1987 Int. Conf. Commun., Seattle, WA, June 1987.

[26] J. Bingham, Telebit, private communication, Oct. 1990.

[27] J. T. Aslanis and J. M. Cioffi, "Achievable information rates on digital subscriber loops: Limiting information rates with crosstalk noise," to appear in *IEEE Trans. Commun.*

[28] H. Lou and J. M. Cioffi, "A programmable Viterbi decoder architecture," presented at Proc. GLOBECOM '90, San Diego, CA, Dec. 1990.

Peter S. Chow (S'87) was born in Taiwan on November 26, 1964. He received the B.S.E. degree in electrical engineering and computer science from Princeton University, Princeton, NJ, and the M.S.E.E. degree in electrical engineering from Stanford University, Stanford, CA, in 1988 and 1989, respectively.

Presently, he is a Ph.D. candidate in the Electrical Engineering Department at Stanford University, and his current research interests are in the areas of digital communications, multiuser information theory, and computer networks.

Mr. Chow is a recipient of the National Science Foundation Graduate Fellowship and the Tau Beta Pi Graduate Fellowship, and is a member of Tau Beta Pi and Phi Beta Kappa.

Jerry C. Tu (S'86), for a photograph and biography, see this issue, p. 908.

John M. Cioffi (S'77–M'78–SM'90), for a photograph and biography, see this issue, p. 764.

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