Optimization of Discrete Multitone to Maintain Spectrum Compatibility with Other Transmission Systems on Twisted Copper Pairs

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***Abstract*—The growing demand to transmit high-speed digital data in many local area networks (LAN's) and digital subscriber lines (DSL's) has resulted in a wide variety of transmission systems that have to co-exist on twisted wire copper pairs. In this paper, we address the problem of maintaining spectrum compatibility between various services that may use different transmission technologies, by shaping in an optimal manner, the power spectral density (PSD) of the transmit signal. A multitone modulation scheme such as discrete multitone (DMT) has the flexibility of optimizing the power spectrum over more than one (disjoint) frequency band, and is suitable for twisted pair subscriber loops, and other transmission media, where the optimized transmit spectrum is likely to occupy more than one frequency bands. DMT has been selected by the American National Standards Institute (ANSI) T1E1.4 Standards Committee as the standard modulation scheme for asymmetric DSL (ADSL). The results presented in this paper are for the specific application of DMT to transport ADSL payloads of over 6 Mb/s from the network to the customer. We consider spectral compatibility between ADSL, the T1 repeater system, high bit-rate DSL (HDSL), and integrated services digital networks (ISDN) basic rate access (BRA) systems.**

The simulation results show that: 1) One can customize the transmit PSD to achieve optimum ADSL performance in a specified noise environment; 2) this optimum performance can result in as much as approximately 6 dB improvement in signal-to-noise ratio (SNR) when compared to the nonoptimized PSD chosen by the T1E1.4 committee; 3) in achieving the above improvements, the total maximum transmit power is still consistent with the limit set by the T1E1.4 committee. Further work is required to support the simulation results with measured data. The mathematical analysis is based on the use of Lagrange multipliers to solve the constrained optimization problem, and is easily extended to other asymmetric and full-duplex wireline transmission systems operating at much higher data rates. The practicality of implementing the proposed optimization routine requires further investigation.

I. Introduction

THERE IS A growing demand to transmit high-speed digital data over twisted wire copper pairs in many LAN's and DSL's. As such, a wide variety of public and private network transmission systems and technologies that are application specific, have to co-exist within the same, or in adjacent binder groups. It is desirable that the introduction of new services such as ADSL does not degrade the performance of existing in-binder or adjacent-binder services such as ISDN BRA, HDSL, and the T1 repeater system, and *vice versa.*

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In this paper, we address the problem of maintaining spectrum compatibility between the various services, by shaping in an optimal manner, the PSD of the transmit signal that is being introduced into the network. In addition to maintaining spectrum compatibility, optimal spectral shaping also compensates for cable attenuation, limits impairments arising from discontinuities caused by artifacts such as wire gauge changes, splices, connectors, jacks, and punch-down blocks. It can also be used to control electromagnetic emissions from the cable, to insure that emissions and susceptibility requirements are met. Optimal spectral shaping has the potential to further increase the reach of twisted wire pair transmission systems.

When a new service is introduced into the network, an attempt is usually made to place it in an unused segment of the frequency band that is not occupied by existing services. This may not be always feasible because of the desire to fully exploit the low frequency portion of the twisted pair cable that has very good loss characteristics. In general, the optimized transmit spectrum is likely to occupy two or more (disjoint) frequency bands because of the different frequency characteristics of the crosstalk noise from other services.

A single-tone modulation scheme such as quadrature amplitude modulation (QAM) is restricted to concentrating transmitted power in one frequency band. On the other hand, a multitone modulation scheme such as DMT [1], [2] has the flexibility of optimizing the power spectrum over more than one (disjoint) frequency band. As such, the performance (achievable data rate for a given error probability) of QAM, even with a minimum mean square error (MMSE) decision feedback equalizer (DFE) [3], is likely to be worse than the performance of DMT, for twisted pair subscriber loops. In this paper, we investigate the performance of the DMT technology, which has been selected by the ANSI T1E1.4 Standards Committee as the standard modulation scheme for ADSL. The results presented are for a specific application of DMT to transport ADSL payloads of over 6 Mb/s from the network to the customer. In this case, we consider spectral compatibility between ADSL, the T1 repeater system, HDSL, and ISDN BRA systems. However, the mathematical analysis, which is based on the use of Lagrange multipliers to solve the constrained optimization problem, is easily extended to other

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asymmetric and full-duplex systems that utilize much higher data rates.

The rest of the paper is organized as follows. The optimal spectrum shaping algorithm that was developed for DMT is explained in Section II. The ADSL simulation parameters are described in Section III, simulation results are presented in Section IV, and conclusions are presented in Section V.

II. Optimal Shaping of DMT

In this section we present a streamlined derivation of the spectral shaping algorithm for the DMT transmit signal. Lagrange multipliers are used to solve the optimization problem. We will not attempt to give a detailed description of the DMT modulation in this section, but rather we refer readers to papers that present the theoretical foundation [1], [2], and performance evaluation of the DMT system for HDSL [4]. Where appropriate there may be some overlap with the analysis published elsewhere in order to preserve continuity in the development.

A. The DMT System

In the specific implementation of the system that we studied, data from the bit stream arrives at the input to the transmitter at the rate of R b/s. Over a baud interval of T (the DMT block length), the bit stream is buffered into blocks of $b = RT$ bits. The DMT channel spectrum is divided into N independent subchannels, an N-point Fast Fourier Transform (FFT) is used to approximate the subchannel center frequencies, and b_i bits are assigned to each positive frequency subchannel. Hence $b = \Sigma_{i=1}^{N/2} b_i$, where some b_i's may be zero. Each set of b_i bits are mapped in the encoder into a complex subsymbol, which forms the QAM constellation for that subchannel.

If the DMT sampling rate is denoted by f_s, then the positive subchannel center frequency $f_k = kf_s/N$ Hz, for $k = 1, \cdots, N/2$. The corresponding subchannel center frequency transfer function is denoted by $H_A(f_k)$, where $H_A(f)$ is the overall channel transfer function (including any transmitter and receiver filters). The N-point inverse FFT (IFFT) that follows the encoder, combines the $N/2$ complex subsymbols into a set of N real-valued time domain samples. Strictly speaking, the subchannels are not independent for finite N, so a cyclic prefix [5] consisting of the last v ($\leq N$) samples from the output of the IFFT is prefixed to the IFFT output samples to remove intersymbol interference (ISI) between the subchannels. $1/T \neq f_s/N$ in general, because of the nonzero cyclic prefix. The $N + v$ samples are converted to serial format by the parallel-to-serial converter (PSC), and applied to a digital-to-analog converter (DAC), with signaling rate $f_s = (N + v)/T$. The lowpass filtered signal from the DAC is the continuous-time modulated waveform. This process is repeated for the block of data in each baud period.

At the receiver, the noise-corrupted channel output is lowpass filtered and converted to digital format by the analog-to-digital converter (ADC). The cyclic prefix reduces the achievable data rate by a factor of v/N. Since v may not be negligible when compared to N, a linear time domain equalizer (TEQ) follows the ADC to reduce the effective constraint length of the channel. The cyclic prefix is stripped from the TEQ output, and the residual converted into a parallel format by a serial-to-parallel converter (SPC), and is processed by the N-point FFT. The $N/2$ output frequency samples from the FFT is decoded and buffered to generate data at the rate of R b/s.

B. Aggregate Number of Bits Per DMT Block

In this section we derive an expression for the number of bits that are transmitted in each DMT block that spans T_s. Most of the material in this section is taken from [1]. It is summarized and presented here in a concise manner to aid in the development that follows in the next section. The interested reader is referred to [1] for more details.

The output signal from each subchannel can be considered to be a QAM signal. Let K represent the used, positive frequency subchannels, and assume an L_k-QAM two-dimensional constellation on the kth subchannel in the set. Then the aggregate number of bits per DMT block is given by

$$b = \sum_{k \in K} \log_2 L_k. \tag{1}$$

The symbol error probability on the kth subchannel is approximated by [1]

$$P_e \approx 4\Phi\left(\frac{d_{\min,k}}{2\sigma_k}\right) \tag{2}$$

where $\Phi(\cdot)$ denotes the normal probability integral, $d_{\min,k}$ is the minimum distance between subchannel constellation points at the channel output, and σ_k^2 is the noise variance per dimension on the kth subchannel. It is assumed that the noise PSD $S_N(f_k)$ is approximately flat over the kth subchannel. In this case, $\sigma_k^2 = S_N(f_k)f_s/N$.

The minimum distance between subchannel constellation points at the channel output is given by

$$d_{\min,k}^2 = d_k^2 |H_A(f_k)|^2 \tag{3}$$

where d_k is the distance between constellation points at the transmitter. In order to maintain the same probability of symbol error per dimension ($P_e/2$) on each subchannel, we require that

$$\left(\frac{d_{\min,k}}{2\sigma_k}\right)^2 = \gamma \tag{4}$$

be constant. γ is usually referred to as the required SNR. For example, if $P_e/2 = 10^{-7}$, then we require that $\gamma = 14.5$ dB. Note that a margin Δ_M may be added to γ, and/or a coding gain Δ_C may be subtracted from γ, to account for unforeseen channel impairments, and/or the coding gain of any applied code (trellis coding, for example).

The average subsymbol energy (or normalized power) P_k in an L_k-QAM constellation is given by

$$P_k = \frac{L_k - 1}{6} d_k^2. \tag{5}$$

The total transmit power

$$P_T = \sum_{k \in K} P_k / R_L \tag{6}$$

where R_L is the ADSL load termination resistance.[1] Substituting (3) and (4) into (5), gives the following expression for the number of points in the QAM constellation on the kth subchannel

$$L_k = 1 + \frac{3P_k|H_A(f_k)|^2}{2\sigma_k^2\gamma}. \tag{7}$$

Let us define the SNR gap Γ, which relates the performance of the DMT scheme considered to the Shannon capacity of the channel, such that $3\Gamma = \gamma$. If we define the SNR of the kth subchannel to be $\mathrm{SNR}_k = P_k|H_A(f_k)|^2/2\sigma_k^2$, then it can be shown that the maximum number of bits that can be supported on this subchannel is

$$b_k = \log_2\left(1 + \frac{\mathrm{SNR}_k}{\Gamma}\right) \tag{8}$$

where $\Gamma = 9.8 + \Delta_M - \Delta_C$ dB. This is the capacity of the subchannel with a factor of Γ less SNR than that which achieves the Shannon capacity. Δ_M is frequently the quantity of interest in subscriber loop applications. From (1) and (7), the aggregate number of bits per symbol period can be expressed as

$$b = \sum_{k\in K} \log_2\left(1 + \frac{P_k|H_A(f_k)|^2}{2\sigma_k^2\Gamma}\right). \tag{9}$$

The optimization problem now involves selecting the normalized power distribution $\{P_k\}$ to maximize the achievable number of bits per DMT block, assuming a given probability of symbol error, as specified by (2), and subject to the total transmitted power constraint in (6), and spectrum compatibility constraints to be derived in the next section.

C. T1 and HDSL SNR Constraints

In this specific application of DMT to transport ADSL payloads of over 6 Mb/s from the network to the customer premises, it is expected that T1, HDSL, and ISDN BRA systems could also share the same, or adjacent binder group with ADSL. It is therefore necessary to consider spectral compatibility between these systems. Because of the frequency response characteristics of T1 and HDSL crosstalk noise, the optimized transmit spectrum will occupy at least two (disjoint) frequency bands. On the other hand, the ISDN BRA bandwidth is approximately twenty percent of the HDSL bandwidth, while the transmitted power is approximately the same. Based on computer simulations using a wide variety of background and crosstalk noise impairments, it has been observed that the SNR performance of ISDN BRA, taking into account interference from ADSL, is quite often close to an order of magnitude better than HDSL. This assumes that the ISDN BRA SNR constraint is not explicitly included in solving the spectral optimization problem. As a result, the ISDN BRA SNR constraint will be omitted from what follows. Instead we will first derive the T1 SNR constraint then the HDSL SNR constraints.

The T1 repeater configuration used in this work is described in [6]. The repeater line section is assumed to be equivalent to that of a 6 kft, 22 AWG cable. Lesser cable lengths are accommodated by a selected quantized line-build-out (LBO) network that makes any cable length appear as approximately 6 kft. The equalized preamplifier transfer function is approximated by the asymptotic frequency response given in [6], and the other T1 system parameters are as specified in [7].

It is assumed that near-end crosstalk (NEXT) from DMT interferes with the T1 repeater system, and *vice versa.* Crosstalk noise from other T1 transmitters, and from HDSL and ISDN BRA could also be present. In the presence of all these impairments, we wish to maintain a minimum SNR at the input to the T1 detection circuit. Let P_Y denote the integrated signal power at the input to the T1 repeater detection circuit, and P_N the additional noise power at that point (excluding any interference from DMT NEXT). If the magnitude squared response of the T1 receiver (up to the input to the detection circuit) to the DMT NEXT transfer function is denoted by $R(f)$, then the noise power at the input to the T1 repeater detection circuit, due to interference from DMT NEXT, can be approximated by

$$P_W \approx \sum_{k\in K} P_k r_k \tag{10}$$

where $r_k = R(f_k)$. Hence, the SNR at the input to the T1 detection circuit is given by

$$\mathrm{SNR}_{T1} = \frac{P_Y}{\sum_{k\in K} P_k r_k + P_N}. \tag{11}$$

An ideal MMSE DFE receiver is assumed for the HDSL system. Let $H_H(f)$ denote the overall frequency response of the HDSL transceiver (up to the input to the DFE), $S_E(f)$ the spectral density of the noise at the to the DFE), σ_h^2 the variance of the input symbol sequence, and $1/T_h$ the baud rate. If in the HDSL frequency band of interest ($|f| < 1/2T_h$), contribution of the aliased components of $H_H(f)$ and $S_E(f)$ are negligible, then the maximum achievable pre-detection SNR is given by [3]:

$$\mathrm{SNR}_H \approx \exp\left\{2T_h \int_0^{1/2T_h} \ln\left(1 + \frac{\sigma_h^2}{T_h}\frac{|H_H(f)|^2}{S_E(f)}\right) df\right\} \tag{12}$$

where $\sigma_h^2 = 5A_h^2/9$ for 2B1Q signaling, and A_h denotes the peak HDSL voltage at the input to the loop. Assume that NEXT from DMT is present at the input to the HDSL receiver, and *vice versa.* Crosstalk noise from other HDSL transmitters, and from T1 and ISDN BRA could also be present. In the presence of all these impairments, we also wish to maintain a minimum SNR at the input to the DFE. Let $S_O(f)$ denote the PSD of the noise (excluding ADSL NEXT) at the input to the HDSL detector, and $H_R(f)$ the frequency response of the HDSL receiver filter. Also, $D(f)$ represents the magnitude squared response of the HDSL receiver (up to the input to the DFE) to the DMT NEXT transfer function. If $|H_H(f)|^2$ and $S_E(f)$ are approximately flat over each ADSL subchannel,

[1] Note that P_T and P_k are measured in W and V^2, respectively.

then (12) reduces to

$$\mathrm{SNR}_H \approx \exp\left\{c_h \sum_{k\in K_h} \ln\left(1+\frac{g_k}{P_k d_k + s_k}\right)\right\} \tag{13}$$

where $c_h = 2T_h f_s/N, g_k = \sigma_h^2|H_H(f_k)|^2/T_h, s_k = S_O(f_k)|H_R(f_k)|^2, d_k = D(f_k)N/f_s$, and K_h represents the set of used subchannels in the frequency band $\{0, 1/2T_h\}$.

D. Optimization of DMT Transmit PSD

In order to insure spectrum compatibility between the DMT-based ADSL system and T1, it is required that crosstalk noise from the optimized DMT signal (and also from other T1's, HDSL, and ISDN BRA, if present), should not cause the T1 pre-detection SNR to fall below some minimum value denoted by $T1_{\min}$. From (11), this idea can be expressed as follows:

$$\frac{P_Y}{\sum_{k\in K} P_k r_k + P_N} \ge T1_{\min}. \tag{14}$$

For HDSL, let the minimum pre-detection SNR be denoted by $H_{\min}$. Then, the optimized DMT crosstalk noise (and also crosstalk noise from other HDSL's, T1, and ISDN BRA, if present), into HDSL, should insure that $\mathrm{SNR}_H \ge H_{\min}$. It follows from (13), that

$$\exp\left\{c_h \sum_{k\in K_h} \ln\left(1+\frac{g_k}{P_k d_k + s_k}\right)\right\} \ge H_{\min}. \tag{15}$$

For ease of notation, let $h_k = |H_A(f_k)|^2, \beta_k = \sigma_k^2$, and $\alpha = 2\Gamma$. We wish to choose the set K of all subchannels which support at least one bit per symbol, and find the set of normalized powers $\{P_k\}$ to maximize the aggregate number of bits per symbol as defined in (9), i.e.,

$$\max_{\{P_k\}}\left\{\sum_{k\in K} \log_2\left(1+\frac{P_k h_k}{\alpha\beta_k}\right)\right\} \tag{16}$$

subject to the constraints in (6), (14), and (15). We solved this problem using Lagrange multipliers. If the values for $T1_{\min}$ and $H_{\min}$ are absorbed into the Lagrange multipliers λ_2 and λ_3, respectively, then the Lagrangian is

$$\begin{aligned} L(\{P_k\}) = &\sum_{k\in K} \log_2\left(1+\frac{P_k h_k}{\alpha\beta_k}\right) \\ &- \lambda_1 \sum_{k\in K} P_k - \lambda_2 \sum_{k\in K} P_k r_k \\ &- \lambda_3 \exp\left\{c_h \sum_{k\in K_h} \ln\left(1+\frac{g_k}{P_k d_k + s_k}\right)\right\}. \end{aligned} \tag{17}$$

Setting the derivative with respect to P_k equal to zero gives[2]

$$\begin{aligned} &\frac{h_k}{P_k h_k + \alpha\beta_k} - \lambda_1 - \lambda_2 r_k \\ &+ \lambda_3 \frac{g_k d_k}{(P_k d_k + s_k + g_k)(P_k d_k + s_k)} = 0. \end{aligned} \tag{18}$$

[2]Using the relationship, $(d/dx)e^{f(x)} = e^{f(x)} f'(x)$.

Equation (18) is the desired design equation for obtaining the optimized spectrum of the DMT-based ADSL transmit signal, when subject to a total transmitted power constraint, and spectrum compatibility constraints from T1, HDSL, ISDN BRA, and other ADSL signals. An appropriate search algorithm can be used to obtain the optimal P_k's from (18).

III. ADSL Simulation Parameters

The following base parameters were used for simulation of the downstream segment of a DMT transceiver for the ADSL system:

Data rate (R)	6.536 Mb/s
FFT size (N)	512
Sampling rate (f_s)	2.208 MHz
Carrier separation (f_s/N)	4.3125 kHz
Symbol Rate $(1/T)$	4 kHz
Cyclic prefix length (v)	40
Target bit error rate (BER)	10^{-7}.

The ADSL downstream data rate of 6.536 Mb/s represents one DS2 (6.312 Mb/s) plus an ISDN BRA (160 kb/s) plus a control channel (64 kb/s). The computer simulation results are presented in terms of SNR performance margins. This represents the amount by which the measured SNR exceeds the theoretical SNR that is required for a BER of 10^{-7}. For some LAN applications, the target BER is 10^{-11}. The performance margins for this BER are obtained by subtracting 2.0 dB from the SNR margins presented in the next section.

In addition to the above parameters, we assume ideal echo cancellation, no guardband, and no coding. The use of a practical echo canceler to separate the downstream and upstream channels would reduce the performance margins, while the use of coding would increase the margins. We also assume that $R_L = 100\ \Omega$. A maximum of 16 bits were allocated to the subchannels. This gives virtually the same performance as the bit allocation strategy that assigns the theoretical maximum number of bits that are based on the measured channel output SNR. Bits are rounded to the nearest integer, not exceeding the maximum bit allocation, and then assigned to each subchannel.

Adjacent binder grouping is assumed for crosstalk noise to, and from T1, otherwise, we assume the same binder grouping. We have also included the 5.5 dB averaging loss [8] when considering T1 NEXT interference into ADSL.

IV. Computer Simulation Results

For data rates over 6 Mb/s, the CSA is the target coverage area for the DMT-based ADSL system. Eight CSA loops are used in the simulation study. Listed in Table I are some rudimentary information about the CSA loops. The list includes calculated resistance, loop length, and insertion loss between $R_L = 100\ \Omega$ termination, at a temperature of 70°F.

Eleven (11) sets of noise environments were simulated on each loop, in order to capture the potential impairments that may affect the performance of ADSL, T1, HDSL, and ISDN BRA, from a spectrum compatibility perspective. We observed variations of up to approximately 4, 1, and 2 dB, in the SNR performance margins for ADSL, T1, and HDSL (and ISDN BRA), respectively, for the range of simulated noise. Unless

TABLE I
RESISTANCE, LOOP LENGTH, AND INSERTION LOSS VALUES FOR CSA TEST LOOPS AT 70°F, WITH $R_L = 100\ \Omega$ TERMINATION

CSA Loop Number	*Insertion Loss (dB)* 100 kHz	200 kHz	400 kHz	600 kHz	800 kHz	1000 kHz	*Resistance* (Ω)	*Length* (ft)
1	27.6	36.5	40.8	48.7	60.0	62.1	643	8300
2	26.0	41.4	37.7	48.8	54.6	58.3	556	8400
3	28.4	33.3	43.4	52.9	61.4	70.9	696	8850
4	29.6	39.6	49.2	53.8	56.5	67.3	635	8800
5	31.2	32.4	43.2	50.2	56.2	62.9	634	8950
6	30.0	35.2	45.1	54.4	62.8	70.2	752	9000
7	26.8	39.3	43.1	57.9	60.9	72.0	562	11500
8	27.7	34.3	46.9	57.4	66.3	74.2	630	12000

TABLE II
SIMULATED NOISE ENVIRONMENTS FOR ADSL, T1, HDSL, AND ISDN BRA SYSTEMS. NEXT NOISE FROM T1 INTO OTHER SYSTEMS IS ASSUMED TO BE IN ADJACENT BINDERS. ALL OTHERS ARE IN THE SAME BINDER. BACKGROUND NOISE AT −140 dBm/Hz IS ALSO ADDED TO EACH SYSTEM IMPAIRMENT LISTED IN THE TABLE

System	*Crosstalk Noise*
ADSL	24 T1 NEXT + 24 ISDN BRA NEXT
T1	24 ADSL NEXT + 10 T1 NEXT + 10 HDSL NEXT
HDSL	24 ADSL NEXT + 10 T1 NEXT + 10 HDSL NEXT
ISDN BRA	24 ADSL NEXT + 10 T1 NEXT + 10 ISDN BRA NEXT

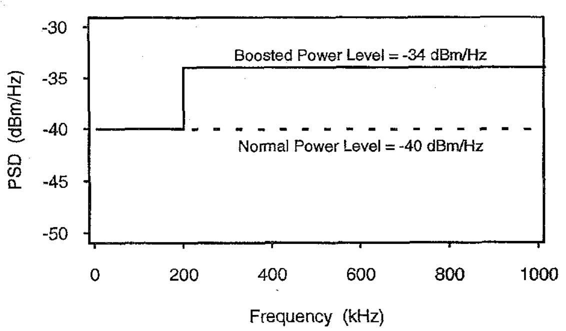

Fig. 1. The proposed ADSL transmit power PSD mask taken from [9].

otherwise specified, the results presented in this paper are based on the set of noise impairments listed in Table II.

Fig. 1 shows the ADSL transmit signal PSD mask that was proposed for consideration by the T1E1.4 Standards Committee [9]. The nominal level is −40 dBm/Hz up to 200 kHz, and increasing up to maximum of −34 dBm/Hz above 200 kHz, depending on line conditions. Shown in Fig. 2 are the optimized PSD's on CSA loops #4 and #8. The optimized PSD's for the other 6 CSA loops fall within these two bounds.

Fig. 3 presents some performance results for ADSL and T1, when the proposed PSD mask in Fig. 1 is used as the ADSL transmit signal. The SNR margins are shown as a function of the amount of power boosting (in 2 dB increments) in the high frequency region of the ADSL frequency band. In all cases, the nominal HDSL and ISDN BRA SNR margins are approximately 12 dB or more. Shown in Fig. 4 are the ADSL performance margins for the optimized ADSL transmit signal. Note that the results are presented as a function of the four T1 SNR margins that correspond to the power boosted levels for the proposed PSD mask in Fig. 3. Furthermore, in order to make a direct comparison between the results in Figs. 3 and 4, the HDSL and ISDN BRA performance margins are also maintained at the same levels in both cases.

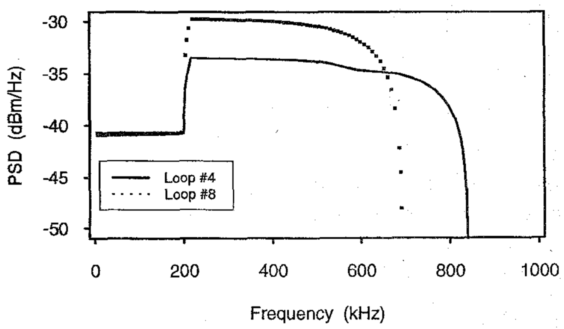

Fig. 2. Example of the line dependent, optimized ADSL transmit power PSD's for CSA loops #4 and #8. Optimized PSD's for the other 6 CSA loops fall within these two bounds.

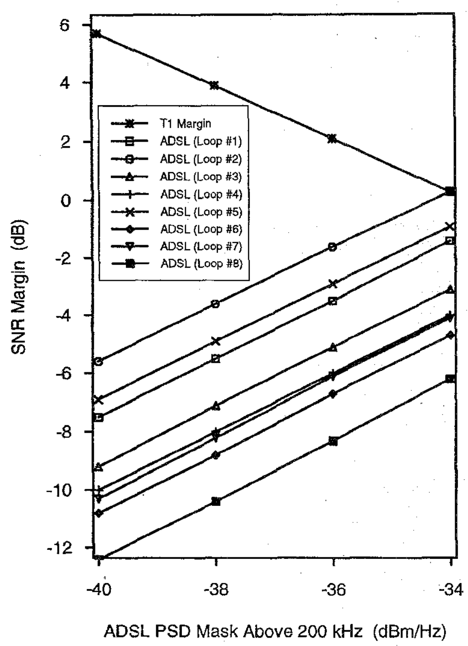

Fig. 3. T1 and ADSL performance margins as a function of the proposed ADSL PSD mask in [9]. Below 200 kHz the PSD mask is fixed at −40 dBm/Hz. In all cases, the nominal HDSL and ISDN BRA performance margins are approximately 12 dB or more, on all 8 CSA loops. Data rate is 6.536 Mb/s.

An example of the improvements (over the proposed PSD mask), in terms of ADSL SNR margins, due to optimal spectral shaping is shown in Fig. 5. It is assumed that the high frequency power of the proposed mask is boosted to −34 dBm/Hz. The T1 margin (0.3 dB), and HDSL and ISDN BRA margins (approximately 12 dB or more), are the same for the proposed and optimized ADSL transmit signals.

V. CONCLUSION

In this paper, we presented the results of a study to determine the extent to which optimal spectral shaping of a DMT-based transmission scheme can be used to maintain spectrum compatibility between various services on twisted wire pairs, that may use different transmission technologies. The results presented are for a specific application of DMT to transport downstream ADSL payloads of over 6 Mb/s. The desired objective is to maintain spectral compatibility between ADSL, the T1 repeater system, HDSL, and ISDN

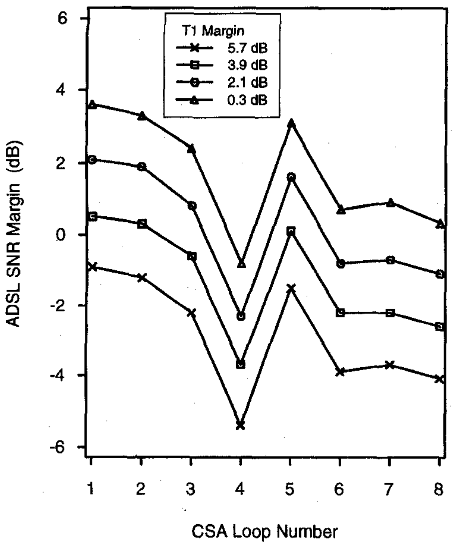

Fig. 4. ADSL performance margins on all 8 CSA loops with the optimized ADSL transmit spectrum. The T1 margins of 5.7, 3.9, 2.1, and 0.3 dB, correspond to those obtained when the proposed PSD mask (above 200 kHz) in [9] is set at −40, −38, −36, and −34 dBm/Hz, respectively. The corresponding HDSL and ISDN BRA performance margins are the same as for Fig. 3. Data rate is 6.536 Mb/s.

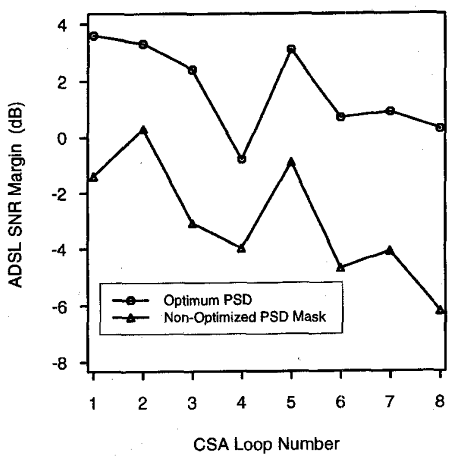

Fig. 5. Comparison of ADSL performance on all 8 CSA loops, with the proposed ADSL PSD mask (at −34 dBm/Hz above 200 kHz), and the optimized ADSL transmit spectrum. The T1 margin (0.3 dB), and HDSL and ISDN BRA margins (approximately 12 dB or more), are the same for both transmit signals. Data rate is 6.536 Mb/s.

BRA systems. DMT is also suitable for other transmission media where the optimized power spectrum is likely to occupy more than one frequency bands.

It is shown through mathematical analysis and computer simulation studies that shaping of the transmit signal spectrum in an optimal manner can improve spectrum compatibility between services on twisted wire pairs, and enhance the performance of these services. We obtained up to approximately 6 dB improvement in signal-to-noise ratio (SNR) using the optimized ADSL power spectral density (PSD) when compared with the non-optimized PSD chosen for ADSL by the T1E1.4 committee. The transmit power was maintained within the limit set by that committee. Further work is required to support the simulation results with measured data. The mathematical analysis presented in this paper is quite general. It is easily extended to other asymmetric and full-duplex wireline transmission systems operating at much higher data rates. The practicality of implementing the optimization routine needs to be investigated.

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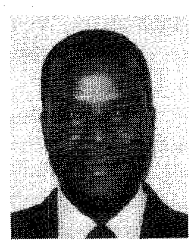

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