A Simulation Study of 2.048Mbit/s Systems for Digital Subscriber Loop Transmission

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Abstract: This paper reports the simulation study of 2.048Mbit/s data transmission systems for digital subscriber loops. A trellis-coded system with the reduced state sequence estimation or the sequential sequence estimation decoder and an uncoded system with a decision feedback equalizer are considered. A double side fractionally–spaced ($T/2$) forward filter (FSFF) is used to suppress the NEXT noise. The results show that the double side FSFF (D-FSFF) achieves a considerable performance margin over the single side FSFF (S-FSFF) in both coded and uncoded system. The trellis-coded system offers further extra system margin over the uncoded system.

1 Introduction

High rate digital subscriber loop (DSL) transmission for the future expansion of ISDN has received considerable attention in recent years. The conventional decision feedback equalizer (DFE) receiver for 4PAM offers a simple quick realization for high rate DSL transmission, but it fails in reach because of the large crosstalk noise and attenuation. More sophisticated modulation schemes should be considered in order to achieve large reach of high rate DSL transmission.

Among several types of modulation the trellis-coded modulation [1] offers better performance for high rate DSL transmission. The optimum decoding strategy of a trellis-code is the maximum-likelihood sequence estimation (MLSE) [2] implemented by Viterbi Algorithm (VA). However, the complexity of MLSE increases proportional to the number $S \times 2^{mL}$ where S is the number of the encoder states, m is the number of transmitted information bits per symbol and L is the memory length of the channel. It is well known that the DSL channel has a long impulse response. The number of the combined states of the encoder and channel is so large that the MLSE decoder is not realizable. Two alternative suboptimal decoding methods are the reduced state sequence estimation (RSSE) [3] and the sequential sequence estimation (SSE) [4], which reduce the complexity of MLSE but offer at least a partial performance gain of MLSE.

The dominant NEXT noise is colored and cyclostationary. In this paper we use instead of a noise predictor [5,6], a D- FSFF to suppress the NEXT noise. The advantage of the D-FSFF over the noise predictor is that it needs not to be added at each survivor path in the trellis-decoder. Furthermore it can exploit the cyclostationary character of the NEXT noise and it makes the synchronization easy. We will also discuss the effect of a high pass filter which mitigates the impulse noise. This high pass filter can shorten the impulse response of the channel but it results in a loss of the signal to noise power ratio.

In this paper a trellis-coded 8PAM and an uncoded 4PAM data transmission system are studied. Section 2 gives the system configurations and the details of the receiver architecture. In Section 3 we explain the use of a D-FSFF for suppression of the NEXT noise. Section 4 presents the simulation results and the conclusion.

2 System Structure

The system structure proposed is shown in Fig.1. Two information bits are sent per signaling interval. The trellis encoder produces 3 encoded bits which are assigned to the 8PAM expanded signal constellation. The uncoded system can be seen as a trellis-coded system with only one code state. The trellis encoder delivers 2 bits assigned to the 4PAM signal constellation. The modulated PAM signal is fed to a pulse shaping filter which generates an overlapping raised cosine PAM signal. This signal is filtered by a square root raised–cosine transmitter filter. The channel consists of a subscriber line and transformers. The NEXT noise is modelled as having Gaussian distribution with a spectral shaping defined in [7]. We generate the NEXT noise in a similar fashion as the received signal. This noise is added at the receiver. In addition to a square root raised–cosine receiver filter a high pass filter is used to suppress the impulse noise. The received signal is filtered by an adaptive fractionally spaced $T/2$ forward filter before the decision is made in the decoder.

2.1 Uncoded system with DFE

The first system considered uses 4 level pulse amplitude modulation (4PAM). In this system a DFE is employed to eliminate the postcursor intersymbol interference (ISI). The precursor ISI will be mitigated by an adaptive FSFF (double side or single side). The DFE has 25 taps and the FSFF has 22 taps in our study.

2.2 Trellis-coded system

The second system we considered uses a trellis-coded 8PAM. The trellis-codes used are the Ungerboeck codes [1] with 4, 8 and 16 code states. Theoretically these codes have an asymptotic gain of 3.31 dB for 4 states, 3.77 dB for 8 states and 4.18 dB for 16 states over an uncoded 4PAM system under additive white Gaussian noise channel without ISI. Unfortunately the subscriber lines belong to channels with severe ISI. The minimum Euclidean distance of the combined code and channel trellis is obviously smaller than that of the Ungerboeck code. However, the Ungerboeck code for PAM needs only one information bit for the convolutional encoder. Therefore the complexity of the RSSE decoder can be reduced by set partitioning [3].

In this system we use the RSSE decoder or the SSE decoder. We will explain them shortly. From Fig.1 the received signal at the input of the decoder can be expressed as

$$r_n = a_n + \sum_{i=1}^{L} a_{n-i} h_i + v_n \qquad (1)$$

where a_n is the transmitted symbol taken from the 8PAM signal constellation. h_i is a sample of the overall impulse response including subscriber line, transmitter, receiver and FSF filters. v_n denotes the sample of the additive noise caused by NEXT interferers.

The metric of the RSSE decoding method suggested by Chevillat and Eleftheriou [6] can be expressed as

$$\begin{aligned} M_n(\lambda_{n+1}^K) &= M_{n-1}(\lambda_n^K) \\ &\quad + |r_n - \sum_{i=1}^{L} h_i \hat{a}_{n-i}(\lambda_n^K) - a_n|^2 \end{aligned} \qquad (2)$$

where the λ_n^K is the reduced combined state given by

$$\lambda_n^K = (\sigma_{n-K}; X_{n-K}(m_K), \cdots, X_{n-1}(m_1)) \qquad (3)$$

$$X_{n-i}(m_i) = (x_{n-i}^1, \cdots, x_{n-i}^{m_i}), \qquad (4)$$

where the σ_{n-K} denotes the encoder state. X_{n-i} and x_{n-i} represent the information sequence and information bit respectively. K is the number of ISI terms represented by the combined channel and code states. The residual $K - L$ ISI terms are canceled by an internal DFE in each survivor path. m_i represents the depth of set partitioning which must satisfy the following constraint

$$m' \leq m_K \leq m_{K-1} \leq \cdots \leq m_1 \leq m \qquad (5)$$

where m' denotes the number of input bits of the convolutional encoder. With the set partitioning the reduced combined trellis has $S \times 2^{m_1+m_2+\cdots+m_K}$ states. This number of states is much smaller than $S \times 2^{mL}$ in the MLSE decoder.

If we choose $K = L$ and $m_i = m$, then the metric given by equation (2) becomes the branch metric of the VA. The SSE decoding (M-Algorithm) uses practically the same branch metric as the VA, but they differ in survivor paths. Instead of retaining all $S \times 2^{mL}$ paths in VA, the M-Algorithm selects the M survivor paths which have the minimum path metrics at each time. Therefore, we need only to calculate $M \times 2^m$ branch metrics rather than $S \times 2^{m(L+1)}$ branch metrics. Usually some extra computations are required for the selection operations in the M-algorithm.

As in the uncoded system an adaptive FSFF with 22 taps is used to eliminate the precursor ISI. The channel estimator has 25 taps.

3 Suppression of NEXT Noise

It is well known that the dominant NEXT noise is colored and cyclostationary. The noise predictor has often been used [5,6] to whiten the noise and to decrease its power. But the noise predictor can not exploit the cyclostationary character of the NEXT noise. Otherwise by decoding trellis-codes over loops with ISI the noise predictor should be added on each survivor path to avoid an error extension. This increases the complexity of the decoder because the noise predictor operates on error signals instead on symbols in the decoder.

The motivation of utilizing a D-FSFF to suppress NEXT noise is as follows. The precursor terms of FSFF will be used to reduce the precursor ISI. The degrees of freedom of the FSFF postcursor terms will be used to combat the colored NEXT noise. The postcursor ISI produced by channel and FSFF will be eliminated by DFE or estimated in the decoder. Although the large postcursor ISI may result in an error propagation, the extra gain of the SNR from D-FSFF offers a very low bit error rate which is far more from the error propagation region than in the case of S-FSFF. The large postcursor ISI will not be a great problem in the decoder, for the RSSE or SSE decoder utilizes some of the postcursor ISI energy. Another advantage of the D-FSFF is that it allows the use of a relative shorter DFE or estimator with less loss of performance. The reason is that the postcursor terms of D-FSFF can also be used to reduce partially the ISI which can not be eliminated by the short DFE. This can further reduce the complexity of the decoder. By calculating the branch metrics of the RSSE- and M-decoder for each retained path, instead of L terms of ISI, we are concerned with only L' terms. The coefficients of the FSFF and DFE (or estimator) will be updated adaptively by a LMS algorithm, which minimizes the mean square error and offers a good compromise between the suppression of colored NEXT noise and elimination of ISI. The details of the D-FSFF will be discussed in another paper [8].

4 Results and Conclusion

The simulation has been performed on a DSL with a diameter 0.4 mm. As stated above we have considered a transmission rate of 2.048Mbit/s, or 1.024MBaud with uncoded 4PAM and trellis-coded 8PAM (2bit/symbol). 6 NEXT interferers have been considered in our simulation [9]. The phase shift between NEXT clock and the signal clock was chosen for the worst case.

Fig.2 shows the increase of the system margin by using the D-FSFF. The cutoff frequency of the high pass filter used was 80kHz. For uncoded system with DFE the system performance gain achieved using D-FSFF over S-FSFF is about 430 meters. For trellis-coded system with 8 encoder states and M-decoder ($M = 16$, abbreviated S8M16) a gain given by D-FSFF over S-FSFF is 410 meters.

To demonstrate the effect of the high pass filter, we have simulated the uncoded system and coded system with different cutoff frequencies (f_c) of the high pass filter. In both systems we have used the D-FSFF. As before in the uncoded system a conventional DFE was used. In the trellis-coded system an encoder with 8 encoder states and a M-decoder with $M = 16$ were used. As seen from Fig.3 both coded and uncoded system have the same tendency. The systems with the lowest f_c have the maximum reach. It is interesting to see that the systems with f_c =80kHz have only about 10 meters performance loss over systems with f_c = 40kHz, while the systems with f_c = 160kHz [5] and f_c = 256kHz (simple 1-D digital filter see [10]) suffer a performance degradation over the systems with f_c = 40kHz by approximately 70 meters and 160 meters respectively.

Fig.4 illustrates the performance of trellis-coded systems with different encoder states. We have chosen f_c = 80kHz and have used the D-FSFF. From Fig.4 one can see that the trellis-coded system with 4 encoder states using M-decoder $M = 8$ offers only 190 meters performance gain over the uncoded system with DFE. The system with 8 encoder states using M-decoder $M = 16$ offers further 230 meters system margin. The system with 16 encoder states using M-decoder $M = 32$ does not give an obvious performance gain over the system with 8 encoder states.

Fig.5 shows the performance of the trellis-coded system when the RSSE decoder is used. For the system with 4 encoder states, we have chosen 32 combined states in the decoder ($K = 3, m_1 = m_2 = m_3 = 1$, abbreviated S4R32). For the system with 8 encoder states, 64 combined states ($K = 3, m_1 = m_2 = m_3 = 1$, abbreviated S8R64) have been used. It is clear that the RSSE decoder in our systems suffers little performance degradation compared to the SSE decoder (M-decoder) for 8 encoder states.

From Fig.2 to Fig.5 one can see that the D-FSFF has suppressed the NEXT noise and brought a considerable performance gain. In our systems a high pass filter with f_c = 80kHz offers a good compromise between the reach and the rejection of the impulse noise. The trellis-coded system can offer extra system margin needed by 2.048Mbit/s digital subscriber loop transmission, if the encoder with 8 encoder states is used. The M-decoder is slightly better than the RSSE decoder.

Acknowlegements

This work was partly supported by German Academic Exchange Service (DAAD).

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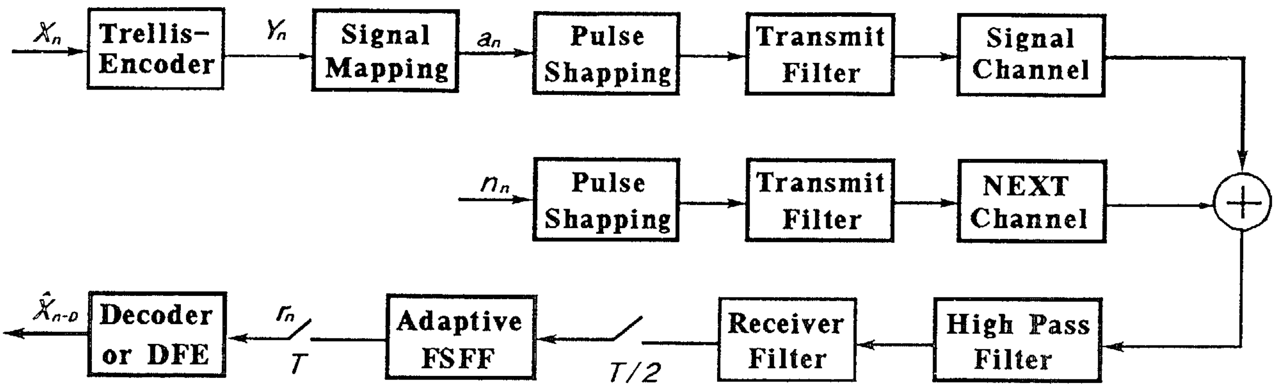
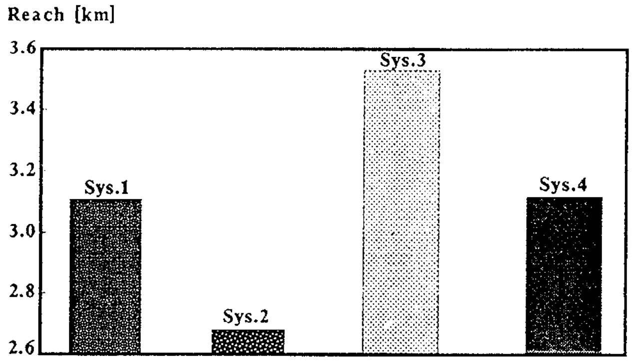
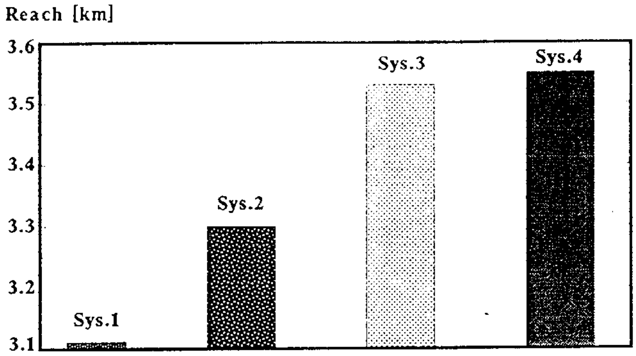


Fig.1: System structure.



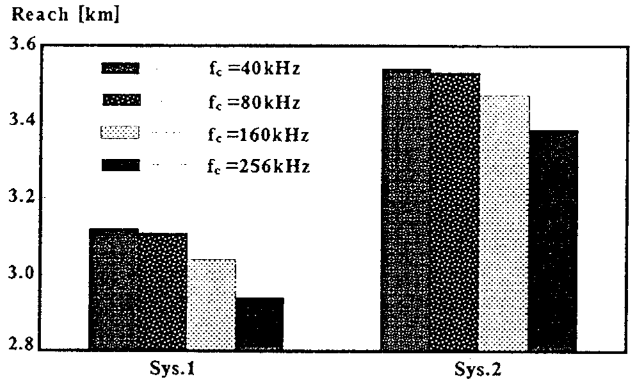
Sys.1: The uncoded system with DFE and D-FSFF
Sys.2: The uncoded system with DFE and S-FSFF
Sys.3: The trellis-coded system with M-decoder and D-FSFF
Sys.4: The trellis-coded system with M-decoder and S-FSFF

Fig.2: Comparesion of D-FSFF and S-FSFF (f_c =80kHz)



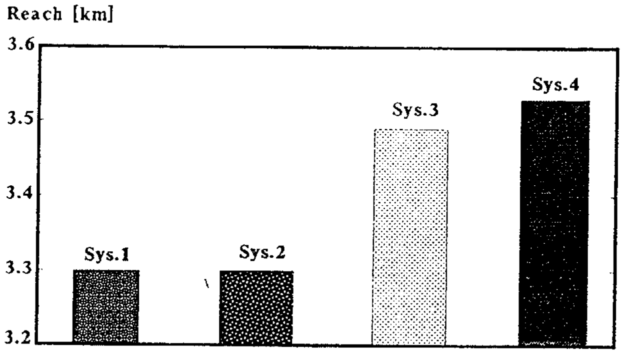
Sys.1: Uncoded 4PAM with DFE
Sys.2: Trellis coded 8PAM with S=4 and M=8
Sys.3: Trellis Coded 8PAM with S=8 and M=16.
Sys.4: Trellis Coded 8PAM with S=16 and M=32.

Fig.4: The performance of trellis coded systems with D-FSFF, f_c =80kHz and different encoder states S.



Sys.1: The uncoded system with DFE and D-FSFF
Sys.2: The trellis coded system with D-FSFF and M-decoder (S8M16)

Fig.3 The effect of the high pass filter.



Sys.1: Trellis-coded system with RSSE decoder and S=4 (S4R32).
Sys.2: Trellis-coded system with M-decoder and S=4 (S4M8).
Sys.3: Trellis-coded system with RSSE decoder and S=8 (S8R64).
Sys.4: Trellis-coded system with M-decoder and S=8 (S8M16).

Fig.5: The performance of the trellis-coded system using RSSE decoder (f_c=80kHz and D-FSFF is used)