A low complexity approach to Turbo Equalizer

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Abstract: - The turbo equalizers (TEQ) proposed in literature utilize equalizers based on trellis, soft Wiener filters. The resulting complexity of these equalizers is exponential and cubic in terms of the sampled channel impulse response (CIR). The interference cancellation based decision feedback filter based equalizers requires adaptation of two filters simultaneously. In this paper, a low complexity equalizer is proposed that neither uses a trellis nor a Wiener filter. The proposed equalizer utilizes a soft interference cancellation (SIC) technique that uses the log likelihood ratio (LLR) available at the matched filter (MF) using all the coded bits in a given block of data. The MF output is justified as Gaussian distributed and the LLRs are computed accordingly. This is fed as the *apriori* to the decoder after suitable deinterleaving. The soft estimates for the bits are used to form an estimate of the interference with the help of perfect channel tap knowledge at the decoder output. This estimate of interference is subtracted from the MF output giving the SIC framework. We call it a soft decision feedback equalizer (SDFE).The SDFE bypasses the filters completely resulting in a linear complexity in CIR. Simulation results over four different channels show that the receiver performance improves with iterations and a gap of 1-3 dB is observed from the coded AWGN bound depending on the channel type. Two different TEQs based on namely soft output Viterbi algorithm (SOVA) and the Wiener filter respectively are compared with the SDFE.

Key-Words: - SIC, Wiener Filter, LLR, SDFE, SOVA

1 Introduction

TEQs [1-13] have found application in mitigating intersymbol interference (ISI) for considerably low signal to noise ratio (SNR) channels of Proakis-B type. However, the impact of estimate of channel tap gains is likely to cause serious problems in convergence of soft log likelihood ratio (LLR) values at the equalizer output. In most of the discussions, the channel tap gains are considered known to the receiver [3-12]. In a similar vein, in this paper also, the known channel estimates are used to reduce the magnitude of error value that is fed to the equalizer algorithm.

The literature on TEQ seems to be divided in three distinct approaches. The finite state machine (FSM) nature of the ISI channel is exploited in [2-4] in order to apply the maximum a posteriori probability (MAP) [14] algorithm or the soft output Viterbi algorithm (SOVA) [15]. The second approach uses a SIC using two filters at the output of the WMF based on the work of [1]. This receiver may be viewed as an *apriori* information aided soft version of the IC proposed by Gersho and Lim [16]. The third distinct approach uses a Wiener filter [6-12] as the equalizer that updates its taps by solving the Yule-Walker equations as per the minimum mean square error (MMSE) criterion. The third approach may be considered as *apriori* aided conventional MMSE. This approach, though computationally less demanding than a trellis based equalizer, nevertheless, the inherent matrix operation needed makes it still computationally intensive. The second approach is the configuration requiring the least number of computations for a given channel and modulation alphabet at a particular value of SNR. All of these receiver configurations are considered from computational point of view in the waterfall or the error floor region. The common feature in all three types of receivers is the use of a whitened matched filter (WMF) as the front-end low pass filter as per Forney's [17] structure that uses a whitening filter.

This paper attempts to provide a simplified equalizer from likelihood point of view, as compared to the SIC proposed in [1]. This gives a flavor of a structure similar to the SIC proposed by [1]-[13],[18-20], however, without any apparent use of feedforward transversal filters [21].

2 Problem Formulation

In Wiener filter based TEQ, the observation z_k is passed through a filter which is estimated as the inverse of the known channel tap gain vector assuming the MMSE criterion. In the SIC, the observation z_k passes through a feedforward filter and the feedback filter is responsible for making the interference estimate which is further subtracted from z_k. In other approaches, where the MAP and maximum likelihood (ML) criteria are used, these LLR values are derived as the a posteriori probability (APP) before and after interleaving conditioned on the entire received sequence. The solution to filter tap gain convergence is expected to get reduced if the relative magnitude of the estimation error is reduced. As soft decisions are subtracted from the matched filter output in a feedback manner, we term the proposed receiver as soft decision feedback equalizer (SDFE).

In the next section, the LLR value computation is derived as a function of measurement noise variance component and also as ISI noise variance. A technique to reduce the noise variance component due to ISI is proposed and analyzed.

2.1 LLR EVALUATION AT THE MF OUTPUT

The WMF output z_k for the $k-$th sampling epoch, assuming baud-rate sampling is

$$z_k = \sum_{l=0}^{L-1} h_l x_{k-l} + w_k = h_0 x_k + \sum_{\forall l, l\neq 0} h_l x_{k-l} + w_k \tag{1}$$

where $w_k \sim N\left(0, \sigma_w^2\right)$ is an i.i.d. noise sample, the ISI channel taps are $h_i, i = 0, \ldots L-1$. The LLR at the WMF output may be approximated as

$$\Lambda_{II}\left(x_k\right) = \log \frac{p\left(z_k \mid x_k = 1\right)}{p\left(z_k \mid x_k = -1\right)} = \log \frac{\exp\left(-\frac{\left(z_k - h_0\right)^2}{2\sigma_{new}^2}\right)}{\exp\left(-\frac{\left(z_k + h_0\right)^2}{2\sigma_{new}^2}\right)} = \frac{2 z_k h_0}{\sigma_{new}^2} \tag{2}$$

where $\sigma_{new}^2 = \sigma_w^2 + \sigma_{ISI}^2$ (3)

where σ_{ISI}^2 is the power contained in the interfering symbols and σ_w^2 is the AWGN variance. This formulation follows by assuming z_k to be Gaussian distributed with mean $\sum_{l=0}^{L-1} h_l x_{k-l}$ and variance $\sigma_w^2 + \sigma_{ISI}^2$. The LLR in (2) is computed for all the samples in a given received data frame and after suitable deinterleaving, fed as *apriori* to the channel decoder, which in turn, needs to compute the updated LLE for each sample.

$$\Lambda_I\left(x_n\right) = \log \frac{p\left(x_n = 1 \mid \Lambda_{II}\left(x_0\right), \Lambda_{II}\left(x_1\right) \ldots \Lambda_{II}\left(x_{N-1}\right)\right)}{p\left(x_n = 0 \mid \Lambda_{II}\left(x_0\right), \Lambda_{II}\left(x_1\right) \ldots \Lambda_{II}\left(x_{N-1}\right)\right)} \tag{4}$$

The equivalent of (4) can be represented as

$$\Lambda_I\left(x_n\right) = \log \frac{\sum_{X^+ : x_n = +1} p\left(\Lambda_{II}\left(x_0\right), \Lambda_{II}\left(x_1\right) \ldots \Lambda_{II}\left(x_{N-1}\right) \mid x_n = 1\right) p\left(x_n = 1\right)}{\sum_{X^- : x_n = 0} p\left(\Lambda_{II}\left(x_0\right), \Lambda_{II}\left(x_1\right) \ldots \Lambda_{II}\left(x_{N-1}\right) \mid x_n = 0\right) p\left(x_n = 0\right)} \tag{5}$$

where X^+ is the set of all combinations of binary data that contain a +1 in the $n-$th position and similarly X^- follows. A binary zero is taken as -1. Some algebraic manipulation of (5) results in

$$\Lambda_I(x_n) = \log \frac{\sum_{X^+:x_n=+1} p(\Lambda_{II}(x_0), \Lambda_{II}(x_1) \ldots \Lambda_{II}(x_{N-1}) \mid x_n = 1)}{\sum_{X^-:x_n=0} p(\Lambda_{II}(x_0), \Lambda_{II}(x_1) \ldots \Lambda_{II}(x_{N-1}) \mid x_n = 0)} + \log \frac{p(x_n = 1)}{p(x_n = 0)} \tag{6}$$

The first term in the RHS of (6) is computed recursively by applying the MAP or SOVA algorithm. The basic computing module is the computation of the transition probability from a given state to another that is expressed as

$$\gamma_n = \prod_{i=1}^{n_0} \exp\left(\Lambda_{II}(x_{n,i})\right) x_{n,i} = \exp\left(\Lambda_{II}(x_{n,i})\right) x_{n,i} \prod_{j \neq i} \exp\left(\Lambda_{II}(x_{n,j})\right) x_{n,j} \tag{7}$$

A SOVA algorithm has been selected in preference to the MAP algorithm in order to implement (7) because of computational cost consideration. The decoder in a TEQ is actually a parallel computing module that computes an updated LLR for each transmitted bit. Thus, use of SOVA decoder reduces complexity (as compared to the MAP or Log-MAP) by a factor proportional to $N \times n_0 \times 2^K$ where N is the received data block size and K is the constraint length of the convolutional code. As the SOVA does not have to store the forward and backward probabilities, the storage of the floating point probabilities is reduced by $2 \times N \times n_0 \times 2^K$.

As the SOVA decoder works in the log domain, for a rate $1/n_0$ outer forward error correcting (FEC) code, (7) becomes equal to

$$\ln \gamma_n = \sum_{i=1}^{n_0} \Lambda_{II}(x_{n,i}) x_{n,i} = \Lambda_{II}(x_{n,i}) x_{n,i} + \sum_{j \neq i} \Lambda_{II}(x_{n,j}) x_{n,j} \tag{8}$$

The second term in the RHS of (6) serves as the *apriori* about x_n

$$L_{ap}(x_n) = \log \frac{p(x_n = 1)}{p(x_n = 0)} \tag{9}$$

The extrinsic information produced by the decoder is computed as

$$\lambda_I(x_n) = \Lambda_I(x_n) - L_{ap}(x_n) \tag{10}$$

The problem boils down to finding an estimate of ISI from this extrinsic information for which we compute the soft bit estimates as [22]

$$\bar{x}_n = \tanh\left(\frac{\lambda_I(x_n)}{2}\right) \tag{11}$$

The residual ISI power σ_{ISI}^2 can be derived from averaging of $1-(\bar{x}_n)^2$ over $L-1$ samples of observation and hence leaves out the contribution from the desired bit.

$$v_n = 1 - (\bar{x}_n)^2 \tag{12}$$

In most of the literature, the z_k is fed to a filter or its equivalent signal processing stages for reduction of ISI to a level suitable for robust processing. In this paper, the possibility of reducing the σ_{ISI}^2 alongwith decision feedback derived as estimate of z_k is explored. It is to be noted that, $v_n = 1$ for which the LLR computation is being carried out following the turbo principle.

It may be noted from (12) that, the soft bit estimate approaches a value of ± 1 for very large values of the extrinsic information, hence

$$\lim_{\lambda_I(x_n) \to \pm\infty} \bar{x}_n = \lim_{\lambda_I(x_n) \to \pm\infty} \tanh\left(\frac{\lambda_I(x_n)}{2}\right) = \pm 1 \tag{13}$$

and thus, (12) tends to zero for these values of the soft bit estimates in (13).

An implication of (12)-(13) is that, the variance of the soft bit estimates tends to be smaller as the TEQ iteration proceeds for more iterations above a particular signal to noise ratio (SNR) that depends on the channel. This is because, the bits become more certain because of enhanced reliability and they approach their true values. The σ_{ISI}^2 variance is dominated by bits having smaller values of reliability which makes the uncertainty relatively higher. Asymptotically, $\sigma_{ISI}^2 \to 0$ and the only impairment that remains is due to channel noise variance only.

In the next section, we explore the possibility of reducing σ_{ISI}^2 by imposing soft decision feedback based interference components subtraction from z_k. This modification coupled with the turbo equalization iteration is studied to understand the convergence of algorithm at faster rate and improved error floor conditions. The interference estimate is made as $\sum_{\forall l, l \neq 0} h_l x_{k-l}$ by making the desired tap position zero similar to the SIC and is further subtracted from z_k.

2.1.1 The proposed SDFE receiver structure and performance analysis

The proposed receiver is schematically given in Fig.1. The observation z_k is used for LLR computation with the help of another parameter $z_k^{'}$ which in turn represents a refined form of z_k by extracting the effects of precursors and postcursors. In other words, z_k is used only for the initial iteration, considered the first iteration in our case, while the future iterations involve $z_k^{'}$ in LLR computation. The switching action gives a flavor of using $z_k^{'}$ for LLR computation for which the initial value of precursor as well as the postcursor is leveled to zero. However, in practice, getting the equalizer loop initialized to all zero condition with trivial zero tailing techniques is intractable.

The receiver structure stems from the fact that the observation z_k may be split into two parts, one representing the effect of precursors and postcursors combined while the second component is the residual value of z_k after subtracting the effect of the first part. In fact, it is the second component that is represented as $z_k^{'}$ as given below.

In fact, for BPSK modulation, the effective MF output, after interference cancellation becomes,

$$z_k^{'} = z_k - \sum_{\forall l, l \neq 0} h_l \overline{x}_{k-l} = h_0 x_k + \sum_{\forall l, l \neq 0} h_l x_{k-l} - \sum_{\forall l, l \neq 0} h_l \overline{x}_{k-l} + w_k \qquad (14)$$

The worst case situation may lead to a condition where all the precursors and postcursors get

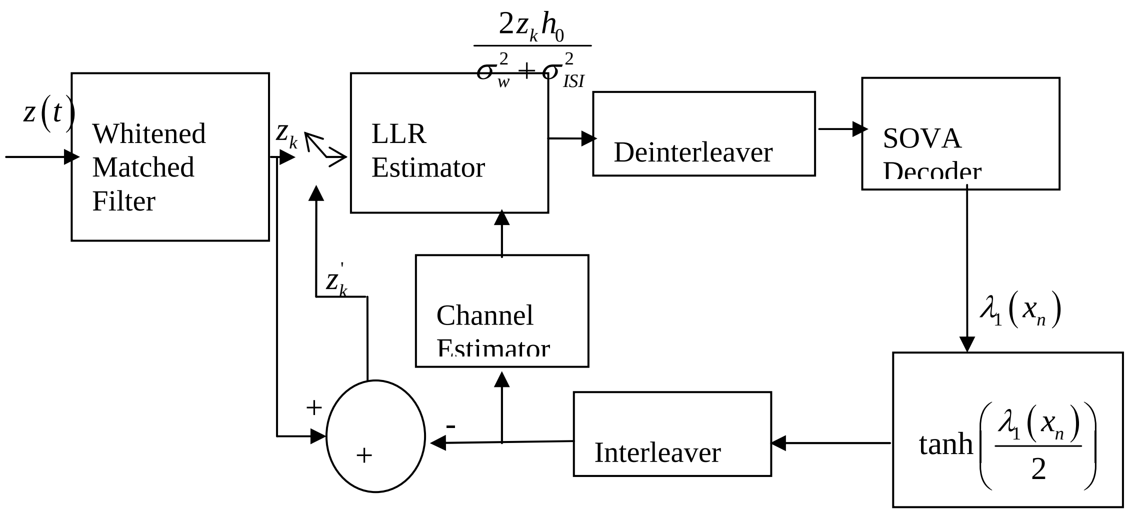

Fig.1 Block Schematic of the proposed SDFE

wrongly estimated because of sign mismatch. In that case, the $z_k^{'}$ will have a mean value of noise plus interference as $\pm 2 \sum_{\forall l, l \neq 0} h_l$.

$$z_k^{'} = h_0 x_k + \sum_{\forall l, l\neq 0} h_l x_{k-l} - \sum_{\forall l, l\neq 0} h_l \bar{x}_{k-l} + w_k = h_0 x_k \pm 2 \sum_{\forall l, l\neq 0} h_l + w_k \quad (15)$$

The BER in such a case after some algebraic manipulations, become

$$P_e = 0.5 erfc\left(\frac{m_{eq}}{\sqrt{2}\sigma}\right) \quad (16)$$

where $m_{eq} = \dfrac{\pm 2 \sum_{l, l\neq 0} h_l}{h_0}$

and $$\sigma_{eq}^2 = \frac{1}{h_0^2}\left(4\left(\sum_{l, l\neq 0} h_l\right)^2 + \sigma_w^2\right) \quad (17)$$

Success of the proposed scheme may be attributed to the fact that, from iteration to iteration, σ_{ISI}^2 as used in (2) also get updated gradually as shown by (12). As a result, $\Lambda_I(x_k)$ is enhanced.

The computational overhead due to σ_{ISI}^2 gain is compensated by the performance gain discussed in the next section. It may be noted here that, a similar reduction in the denominator of (2) has been attempted in [1-5] also. However, additional filter stages have been used in those implementations, which effectively is more rigorous than that of the averaging approach adapted in this paper. It is to be noted that, (14) reduces to an equivalent AWGN channel output when

$$\sum_{\forall l, l\neq 0} h_l x_{k-l} - \sum_{\forall l, l\neq 0} h_l \bar{x}_{k-l} = 0 \quad (18)$$

We note from (18) that, the asymptotic condition prevails when the error defined by the second and third terms in the RHS of (15) reduces to zero and we obtain an equivalent AWGN channel condition at the decoder input.

The SNR as used in the simulations is expressed as

$$SNR = \frac{1}{\sigma_w^2} \quad (19)$$

3 Problem Solution

The performance of the receiver is evaluated through simulation experiments for 4 different channels to illustrate its behavior. In all cases, a higher SNR compared to other schemes is needed to start the performance improvement. This is because, the equalizer block has not been used and performance solely depends on the quality of the soft outputs of the decoder. The outer FEC is chosen to be a rate ½,RSC code with constraint length 3 and generator polynomials [7,5]. A random interleaver has been used for each simulation run. An input data block of size 10000 bits has been used and the BER performance as averaged over 10 simulation runs is plotted as a function of SNR for three typical radio mobile channels. Standard BPSK modulation has been used. In all of these results, the SOVA has been used as the decoding algorithm in order to realize a low complexity receiver as compared to the MAP or its other version algorithm based decoders.

Performance gain of SDFE is compared with those of SOVA based decoding algorithm [23] and also in another attempt with that of Wiener filter based receiver structure. For the case of SOVA decoder, a 3-tap exponentially decaying channel is considered in which the one sided decay parameter is taken so as to represent a slow decaying and fast decaying conditions of the channel interference level.

For the sake of simplicity, the channel taps [0.8823 0.4241 0.2039] represent a fast decay while [0.7840 0.5180 0.3422] represent a slow decay. For SDFE, it is observed to have an additional 1.5 dB of SNR for similar BER performance. For slow decaying channel, the additional SNR requirement for SDFE comes out to be 1.8 dB. In both the cases, the comparison is made after 4 iterations get completed. Those results are summarized in Fig.2 and Fig.3.

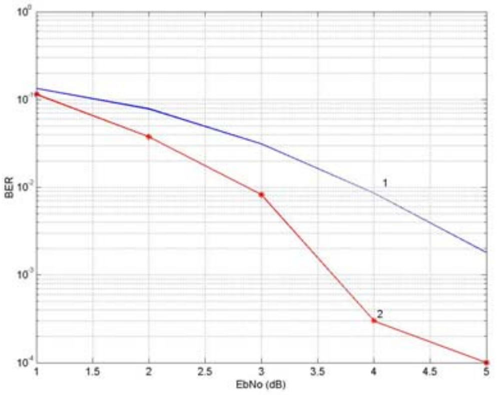

Fig.2 A comparison of performances of the SDFE and SOVA based turbo equalizer for a channel with fast exponential decay.

1: Performance of SDFE 2: Performance of SOVA based TEQ

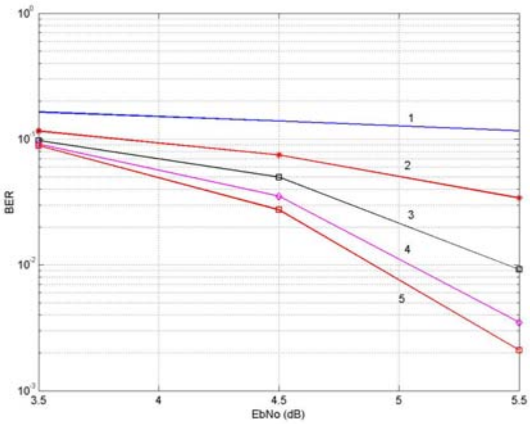

Fig.3 Performance of the proposed SDFE in a slow exponential decaying channel

The SDFE seems to require an additional 1.5 dB of SNR as compared to the SOVA based TEQ to achieve a BER of 0.0019. The advantage of SDFE lies in the fact that no trellis needs to be used for equalizer unlike the SOVA equalizer.

The Wiener filter stage in the receiver is studied in Fig.4 for slow decaying channel in which after 4 iterations as compared with Fig.3 comes close to that of SOVA decoder, however the latter requiring a marginal increment in SNR of 0.4 dB. It may be noted that, for Fig.3 and 4, the BER of 10^{-2} is taken as reference because it is also the coded AWGN bound at 3 dB SNR which is a channel SNR level in most of the practical cases as reported in literature.

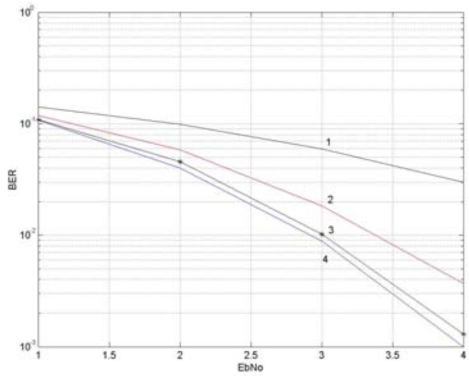

Fig.4 BER performance of MMSE 9-tap Wiener filter in slow exponential fading channel

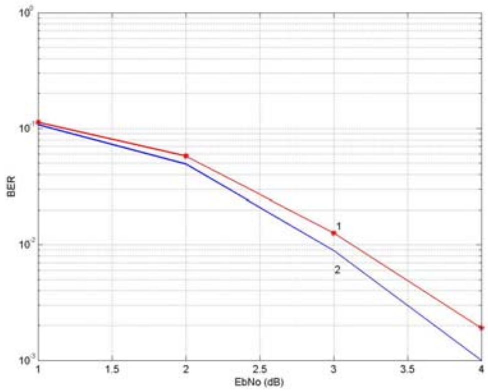

Fig.5 A comparison of performances of the 9-tap Wiener filter and SOVA based turbo equalizer for a channel with slow exponential decay.

1: Performance of SOVA based TEQ 2: Performance of Wiener filter

These two results are shown after 4 iterations

The Wiener filter based turbo equalizer seems to be better than the SOVA based turbo equalizer by about 0.2/0.4 dB for this particular channel.

Turbo equalizers have been extensively studied for channel models referred as Proakis-A [0.04 -0.05 0.07 -0.21 -0.5 0.72 0.36 0.21 0.03 0.07] and Proakis-B [0.407 0.815 0.407]. In Fig.6, we summarize the performance of SDFE for Proakis-A channel. The BER performance improves with respect to iterations significantly, particularly beyond 4 dB SNR. However, the deviation from the coded AWGN bound increases exponentially even after 7 iterations.

For the Proakis-A channel, the proposed receiver shows a gap of almost 3 dB from the coded AWGN bound at a BER of 10^{-4}. This may be attributed to the fact that the main tap occurs after a delay of about 5 bit periods which is a relatively long channel. However, the computational effort on the trellis based equalizers or the soft interference cancellation based filters may not be an appropriate choice to equalize this channel. In other words, channels with a long delay spread may find a better receiver in SDFE from a point of view of tradeoff between performance and computational complexity.

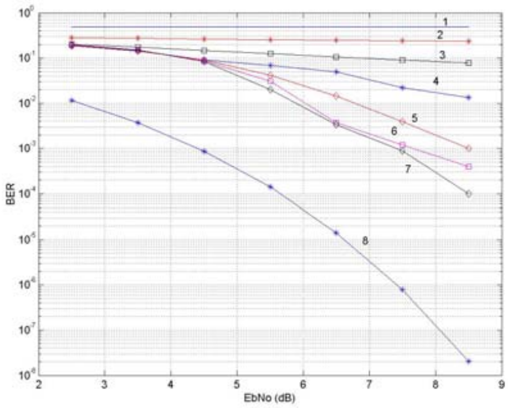

Fig.6 Performance of the proposed receiver in Proakis-A Channel

1:7 Seven iterations carried out by SDFE, 8: Coded AWGN bound

In Fig.7, the performance of SDFE is compared with Wiener filter in Fig.8 for the Proakis-B channel. The performance of the SDFE improves with iterations due to the availability of a cleaner signal at the matched filter output. As the decoder produces more reliable bit decisions, the interference estimate approaches its true value and hence, subtraction from the matched filter output results in a signal with less interference. Hence, the performance improves. The coded AWGN bound has been plotted as a reference for the sake of comparison. The performance shown by this receiver for this particular channel shows a gap of about 1 dB from the theoretical performance. In both the cases, the increase in number of iterations shows a monotonic improvement in performance. However, after fourth iteration, this improvement is observed insignificant.

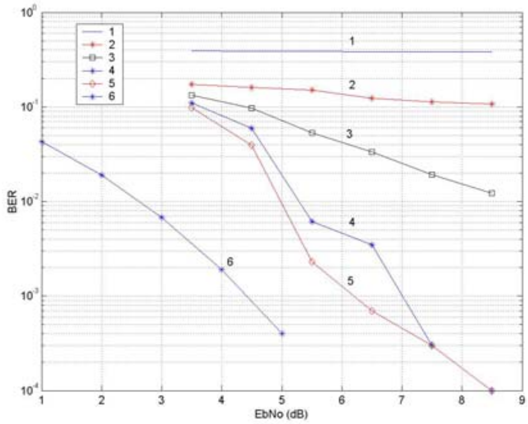

Fig.7 BER Performance of SDFE in Proakis-B

1:5 Five iterations carried out by SDFE, 6: Coded AWGN bound

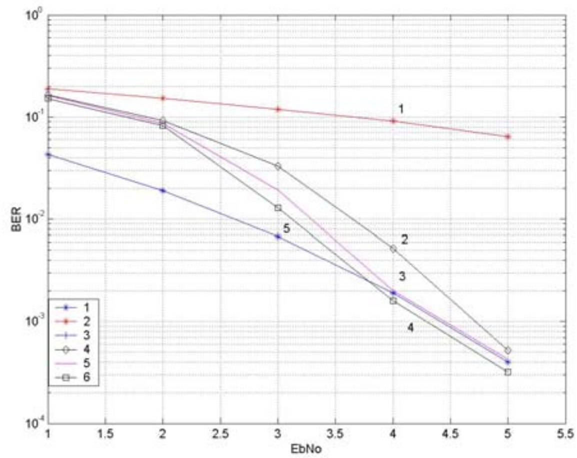

Fig.8 MMSE 9-tap Wiener filter in Proakis-B channel

1: 4 Four Iterations, 5: Coded AWGN bound

The numbers 1:5 indicate the number of iterations performed between the LLR computer unit and the channel decoder. These schemes seem to require higher SNR than the trellis or filter based receivers as is discussed previously. However, if the channel delay spread is large and it is known perfectly, the proposed receiver offers a better alternative from a performance *vs*. complexity trade-off. The proposed receiver is about 2.5 dB away from the theoretical bound at a BER of 8×10^{-4}. The performance does not improve much beyond five iterations.

The performance of a Wiener filter under the assumption of perfect channel tap gains is shown in this figure for the purpose of comparison. This receiver requires a computational effort proportional to $10000\times o\left(9^3\right)=10000\times729$ as compared to the SDFE for a received data block size of 10000. While instantaneous linear filtering is applied to $z_k^{'}$ in the Wiener filter based receiver, we apply $z_k^{'}$ directly as the input to the decoder. The *apriori* aided 9-tap Wiener filter approaches the coded AWGN bound for this particular code after five iterations.

A comparison of Fig.7 and Fig.8 shows the relative gain of defined as Wiener filter based receiver at 2, 3 and 4 dB SNR respectively which is presented in Table 1. The relative gain is defined as

(SDFE BER-Wiener Filter BER)/SDFE BER

Table No.1 Relative gain of the Wiener filter based receiver

Operating SNR (dB)	Relative gain (%)
2.0	78
3.0	94
4.0	98

However, this performance gain is compensated by the computational overhead that increases sharply with increase in input data block length and channel length. For our study, we have considered a Wiener filter of 9 taps for simplicity of matrix inversion. The SDFE has been compared for various iterations with those of Wiener filter and SOVA based approaches in Fig.9 for a slow exponential decaying channel.

The commonality between the performance curves obtained for four typical channels indicate that, (16) is true only for the initial iterations of the equalizer that corresponds to a wrong decision on the part of the decoder and this is shown as the flat performance. However, for higher iterations, (16) is not true and as the decoder is able to make more number of correct decisions because of a cleaner signal presented to its input. This improves the BER.

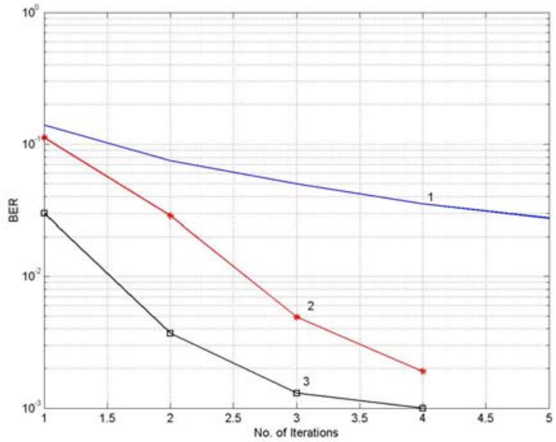

Fig.9 A comparison of 3 schemes for the slow exponential decay channel for 4 iterations at a given SNR

1: SDFE Performance at 4.5 dB SNR for slow exponential decay for 4 iterations

2: SOVA based TEQ at 4 dB SNR for the same channel for 4 iterations

3: 9-tap Wiener filter based performance at 4 dB SNR for the same channel

Table 2.Relative Advantages of SDFE over SOVA and Wiener Filter (per bit)

Type of Receiver	Computational Overhead	Storage	Latency
SOVA	$2^L+(10L+1)$	$2^{(L+1)}$	$2^L+(\delta+1)$
Wiener Filter	2x $o(3L)^3$	$(3L$x$4L-1)+3L$	$3L$

In this table, δ is the traceback depth of the SOVA equalizer which is usually 10 times the length of the ISI channel.

4 Conclusion

In filter based approaches the difficulty of large filter length taps iterative adaptation is realized in the form of turbo based iteration. The structure of SDFE although uses, subtraction of the estimated interference like the Wiener filter algorithm, the interference estimate is derived as extra information on the basis of LLR values. In effect, it is this SDFE architecture is a combination of filter based subtraction and SOVA/MAP based extrinsic information derivation.

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