Novel Blind Decision Feedback Structure for Channel Estimation in Severe ISI Environment

Yasamin Mostofi and Donald C. Cox
Stanford University, Stanford, CA 94305, USA
{yasi@leland.stanford.edu, dcox@nova.stanford.edu}

Abstract– **In any communication environment with delay spread comparable to the symbol period, a % of bandwidth is wasted sending a known training sequence. For time-variant channels, this percentage becomes higher due to the need for more frequent channel estimation. In GSM [1], 17.93% of bandwidth is dedicated to transmission of such a sequence. If a structure can estimate the channel blindly with an acceptable performance, it will save the bandwidth considerably. This motivates our work: the development of a new blind structure for severe ISI channel estimation.**

Keywords– delay spread, blind equalization, DFE

I. Introduction

In any communication environment with coherence bandwidth smaller than the bandwidth of the input signal, Inter-Symbol Interference (ISI) will occur. There are several ways to remove the effect of ISI. Since the receiver does not know both the channel and the input, there is a need for transmission of a known sequence for channel estimation. Transmission of such a sequence is a waste of bandwidth particularly in a mobile environment due to the need for more frequent estimation. This motivated researchers to come up with blind algorithms for channel estimation/equalization. Blind adaptation of both linear and non-linear structures has been explored. Linear blind structures [2] suffer from noise enhancement and fail to produce a reliable channel estimate for non-minimum phase channels. Since in a non-blind scenario, a decision-feedback equalizer (DFE) proved to be the most efficient among the low complexity sub-optimum techniques, it is worth exploring its performance in blind case. [3-4] investigated blind adaptation of a DFE structure with only a feedback section (FB-DFE). Their results indicate that for channels with pre-cursors (channels in which the first received path is not the strongest), FB-DFE fails to converge to its global minimum with high probability[1]. In building environments where there is no LOS[2] path most of the time, the first received path is not necessarily the strongest [5-6]. Therefore FB-DFE can not produce an acceptable performance. This motivates the work presented here, developing a blind structure that can estimate ISI in such an environment. We first address the ill performance of a decision feedback structure with both feedforward and feedback sections when blindly adapted. Then we introduce a new structure which implements the feedforward part of a DFE structure in its feedback section. This structure can produce a reliable estimate of channels with pre-cursors.

[1] More precisely, FB-DFE fails with high probability for channels in which $h(\delta) \geq \sum_{i=0}^{\delta-1} h(i)$ [3-4]

[2] Line of Sight

II. System Model

Consider a transmission system depicted in Fig. 1. The aim is to blindly estimate $h(k)$ from a block of output, $y(k)$, of length M. To simplify the analysis, we assume BPSK input and baseband fixed channels. However, the results can be extended to PSK input and passband time-variant channels.

III. A Decision Feedback Structure

A traditional DFE is shown in Fig. 2. When adapted non-blindly, this structure proved to be the best sub-optimum equalization structure. However, it fails when it comes to blind adaptation. There are a few studies [3-4] indicating its poor performance when it is used with no feedforward (FF) section (FB-DFE). Since the FB section can only effectively deal with post-cursors, FB-DFE fails when it comes to channels with precursors. If the structure with both its FF and FB sections is adapted blindly, it will converge to an undesirable minimum. In other words, the global minimum of the cost function $E\{(|A(k)|^2-1)^2\}$ or $E\{|A(k)-\hat{A}(k)|^2\}$ is reached when one FB coefficient is one and all the other FB and FF coefficients are zero. This will result in $A(k)=\hat{A}(k-t)$ where the delay t depends on which FB coefficient is one. Hence $\hat{A}(k)$ would be equal to $A(k)$ resulting in both cost functions reaching their lowest minimum (zero). Therefore, unconstrained minimization of this structure does not produce an agreeable result. Adding any constrain to prevent the aforementioned minimum[3] will increase

[3] Keeping the norm of FF filter or the value of its first coefficient constant are examples of possible constrains

the complexity of the blind adaptation and may not produce an agreeable result either. Hence we introduce a new decision feedback structure which can handle both pre- and post-cursors of the channel while maintaining low complexity.

IV. New 2-stage structure

Consider the DFE structure in Fig. 2 only with its FB section (FB-DFE) . If channel has no pre-cursors , with high probability FB-DFE will converge to its optimum solution. When the channel has its strongest path at time index p, FB-DFE tries to catch the first path and remove other paths some of which are now stronger than the first one. Hence, it fails with high probability. In the case of a channel of length N with both pre- and post-cursors and the strongest path at p, we can write, $y(k) = h(p) \times I(k-p) + \vec{I}_{post}(k) \times \vec{H}^t_{pre} + \vec{I}_{pre}(k) \times \vec{H}^t_{post}$ (1)
Where $\vec{I}_{post}(k) = [I(k-p+1) \dots I(k)]$,
$\vec{I}_{pre}(k) = [I(k-p-1) \dots I(k-N)]$,
$\vec{H}_{post} = [h(p+1) \dots h(N)]$, $\vec{H}_{pre} = [h(p-1) \dots h(0)]$
In the above equations, $\vec{H}_{post}$ and $\vec{H}_{pre}$ represent post- and pre-cursors of the channel respectively. Since h_p is the strongest path, the goal is to design a structure which estimates and removes the effect of $\vec{H}_{post}$ and $\vec{H}_{pre}$ from y_k. In other words, the first term on the right-hand side of Eq. 1 is the desirable term and the effect of last two terms should be removed. We know that a FB-DFE structure can handle a pre-cursor free channel. Next, consider the post-cursor free channel, depicted in Fig. 3a. It is easy to see that for a received block of length M the reverse of the input passing through the reverse of the channel will produce the reverse of the output sequence as shown in Fig. 3b. Define $h_{r,new}(k)$ to be $h_r(-k)$. Since $h_{r,new}(k)$ represents a pre-cursor free channel, we can pass $y(M-k)$ to a FB-DFE. Then the output and the coefficients of FB-DFE will estimate I_{M-k} (the reverse of the input bits) and $h_{r,new}(k)$ respectively. The new structure for estimating channels with both pre- and post-cursors is depicted in Fig. 4. The principle behind it is the observation that if a channel has no pre-cursor, it can be handled with FB-DFE with high global convergence probability. If a channel has no post-cursor, its time-reversal representation can be viewed as a pre-cursor free channel [7], which can then be estimated with FB-DFE as well. Hence, we split a channel with both pre- and post-cursors into a pre- and post-cursor free channel. The goal is to estimate and remove the last two terms on the right-hand of Eq. 1, hence capturing the strongest path. The forward stage acts similar to a FB-DFE. Its input is the received sequence $y(k)$. Its coefficients and output estimate $\vec{H}_{post}$ and $\vec{I}_{pre}$ respectively hence removing the effect of the third term on the right hand side of Eq. 1 from $y(k)$. The backward stage acts similar to the reverse structure. The input to this stage is $y(M-k) = h(p) \times I(M-k-p) + \vec{I}_{rev,post}(k) \times \vec{H}^t_{pre} + \vec{I}_{rev,pre}(k) \times \vec{H}^t_{post}$ (2)
$\vec{I}_{rev,post}(k) = \vec{I}_{post}(M-k), \vec{I}_{rev,pre}(k) = \vec{I}_{pre}(M-k)$
The coefficients of this stage estimate $\vec{H}_{pre}$ and its output produces an estimate of $\vec{I}_{rev,post}$, hence removing the effect of the second term on the right hand side of Eq. 2. To remove the effect of the second term on the right hand side of Eq. 1 from the input to the forward stage, the forward stage needs the estimates of $\vec{H}_{pre}$ and I_{post} which it does not produce itself. These estimates can be acquired from backward stage since its coefficients estimate $\vec{H}_{pre}$ and its output estimates $\vec{I}_{rev,post}(k)$ which is equal to $\vec{I}_{post}(M-k)$. In Fig. 4, $O(M-k)$ represents an estimate of $\vec{I}_{post}(k) \times \vec{H}^t_{pre}$ which is acquired from the backward stage. In the same manner, the backward stage needs the estimate of the third term on the right hand side of Eq. 2 which it does not produce itself. It hence acquires it from the forward stage. $Q(M-k)$ in Fig. 4 represents this estimate. This results in complete ISI removal. Both stages function the same except for the direction in which the data is processed and the order in which the estimates of $I(k)$ are produced. At time instant k, the forward stage needs the estimate of $\vec{I}_{post}(k)$ which is $[I(k-p+1) \dots I(k)]$. At the same time instant, the backward stage has produced the estimates of $I(M-(k-1)-p)$, $I(M-(k-2)-p)$, ... , $I(M)$. Hence the estimates from the backward stage will be available for the forward stage to use when $k \geq (M+1-p)/2$ which happens approximately when k passes the midpoint of the received block ($M/2$ samples of $y(k)$) . The same argument holds for backward stage. In other words, $O(M-k)$ and $Q(M-k)$ become available for the other stage to use after the midpoint. Therefore both stages perform partial ISI estimation and reduction before the midpoint and begin to remove the ISI completely afterward. After processing

the whole received block, the coefficients of the forward and backward stages have estimates of post- and pre-cursor channels respectively. Since ISI levels on the two sides of the channel may differ noticeably, the performance of the stages are not necessarily the same. For instance, consider a channel in which $\vec{H}_{post}$ has higher ISI level than $\vec{H}_{pre}$. When in partial ISI removal (up to the midpoint of the received block), both stages act independently so forward and backward stages do not have an estimate of $\vec{H}_{pre}$ and $\vec{H}_{post}$ respectively and have to tolerate the ISI caused by them. Since in our example, $\vec{H}_{pre}$ has lower ISI level than $\vec{H}_{post}$, the forward stage has to tolerate less uncompensated ISI than backward stage. Therefore, it is more probable that forward stage produces more reliable results. To avoid error propagation, we can choose to feed the estimates of a stage to the other one after the midpoint, only if it has produced an agreeable performance. If we can measure the performance of each stage based on its behavior up to the midpoint, we can choose to feed only the reliable estimates to the other stage. This will limit the error propagation hence producing a better overall performance.

V. Decision at Midpoint

There are different measures that can be used to evaluate the performance of a stage at midpoint. One such measure is convergence variance. The smaller convergence variance is, the better the performance of a stage would be. Also for a PSK modulated waveform with amplitude one, the smaller the ISI in a stage is, the closer the absolute value of the signal before the slicer would be to one. Hence the deviation of the amplitude of $A(k)$ and $B(k)$ from 1 can serve as a good measure to compare the performance of the stages. If the values produced by a measure function for both stages differ considerably, only the estimates of the more reliable stage will be used by the other stage.

VI. Adaptation Algorithm

The coefficients of both stages are updated through a Decision Directed-based (DD) algorithm with cost functions $E\{|A(k) - \hat{A}(k)|^2\}$ and $E\{|B(k) - \hat{B}(k)|^2\}$. The adaptation formulas for the forward stage coefficients will be (similar for the backward stage) $w_f^{(k+1)}(i) =$
$w_f^{(k)}(i) + \Delta \times w_{0f}^{(k)} \times (A(k) - \hat{A}(k)) \times \hat{A}(k-i)$
$w_{0f}^{(k+1)} = w_{0f}^{(k)} + \Delta \times w_{0f}^{(k)} \times (A(k) - \hat{A}(k)) \times \hat{A}(k)$
$i = 1, \ldots, L$ and Δ is the adaptation step.

VII. Minima of the 2-Stage Structure

If $[h(0)\ h(1)\ \ldots\ h(N)]$ represents a channel profile with its most powerful path at $p \leq N$, a desirable minimum is reached when w_{0f} and w_{0b} reach $h(m)$ and the backward and forward stages estimate $[h(m-1)\ h(m-2)\ \ldots h(0)]$ and $[h(m+1)\ h(m+2)\ \ldots\ h(N)]$ respectively, where $m = p$. But there are other **acceptable** minima too. In general, the 2-stage structure breaks the channel into two complementary parts and estimates each part with one of the stages. Theoretically, the break point (m) can be any point from 0 to N. If the 2-stage coefficients produce two sides of any of these break points, the ISI will be removed completely. Therefore these minima are all acceptable. In practice, it is easier to converge to some of these minima than others. The choice of the channel and how two stages interact at the midpoint (that is which stage is fed to the other) are important factors in determining which minimum the structure will eventually converge too. In most of the cases, the break point m is equal to p. However, since all of them are acceptable, it does not change the performance if any of them is reached.

VIII. Simulation Results

Fig. 5-7 (a,b) show three channels in both time and frequency domain. These channels differ in their ISI levels from mild to severe (from 1-3). Convergence of coefficients of both stages (except for w_{0f} and w_{0b}) are depicted in Fig. 5-7(c,d). Since the speed of convergence is not of concern in this paper, we chose an arbitrary M (length of the received block, in our case 1000) long enough to reach steady state. In this part, noise is not considered. The number of feedforward and feedback coefficients was each set to ten. The extra coefficients (not shown in the figures) converge to zero. We observe the deep nulls in the frequency response of the channels, especially channel 3. The 2-stage starts with partial ISI removal. After the midpoint (500 here), the effect of switching from partial to complete ISI removal is noticeable. The dashed lines indicate the optimum solutions and the coefficients converge toward them after the midpoint. For this part, measure functions were not used. For channels 1 and 2, two-way feeding was performed and for channel 3 one-way feeding was implemented. The most reliable part of the estimates is halfway between midpoint and the end of the block. The reason behind this is as follows. When the stages enter the complete ISI removal mode after the midpoint, they use the results of the other stage (if approved by the measure function) starting from the most recent results and moving toward older ones. Since the most recent results are less erroneous, they lead to more reliable performance. To investigate the effect of noise and different measure functions, we evaluate the % of normalized estimation error vs. SNR for two different midpoint measuring methods of section V. Fig. 8 shows the % of normalized estimation error vs. SNR for these two methods. From Fig. 8, for all the channels when SNR is greater than 15db, % of normalized estimation error is under 5%. This shows that the new 2-stage structure can estimate channels with severe ISI. Also, both error propagation control functions produce similar results and prove to function reliably.

IX. Conclusion

A blind decision feedback structure for severe ISI channel estimation was introduced. Its performance was analyzed under different ISI levels and noise. It was shown that up to high levels of ISI, the structure can estimate the channel successfully. To control the error propagation, two measure functions were introduced. Our simulations showed that both functions act robustly and can be utilized for error propagation control.

X. References

1. M. Mouly, M. Pautet, "The GSM System for Mobile Communications," Cell & Sys, 1992

2. "Blind Equalization for Short Burst Wireless Communications," Thesis by B. J. Kim, Stanford University, 1998

3. Kenney, Brian D. Anderson and R. Bitmead, "Blind Adaptation of Decision Feedback Equalizers: Gross Convergence Properties," International Journal of Adaptive Control and Signal Processing, Vol. 7, 1993

4. "Blind Adaptive Decision Feedback Equalization: A Class of Bad Channels," R. Casas, Cornell University, 1996

5. D. Devasirvatham, "Time Delay Spread and Signal Level Measurements of 850MHz Radio Waves in Building Environments," IEEE Trans. on AP,Vol. 34, Nov. 1986

6. L. Chang, "Dispersive Fading in CDMA Radio Systems," IEEE ICUPC, Oct. 1992

7. S. Ariyavisitakul, "A Decision-Feedback Equalizer with a Time-Reversal Structure," IEEE Trans. on Comm., May 1991

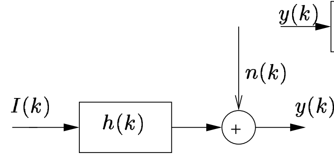

Fig. 1 System Model

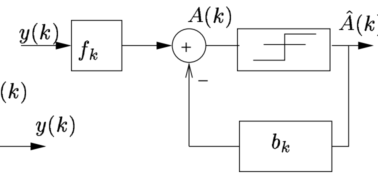

Fig. 2 DFE Structure

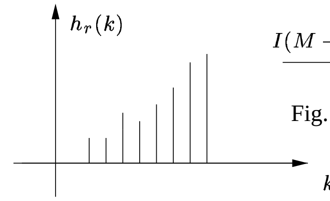

Fig. 3a Post−Cursor Free Channel

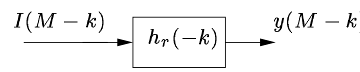

Fig. 3b Reverse Channel Model

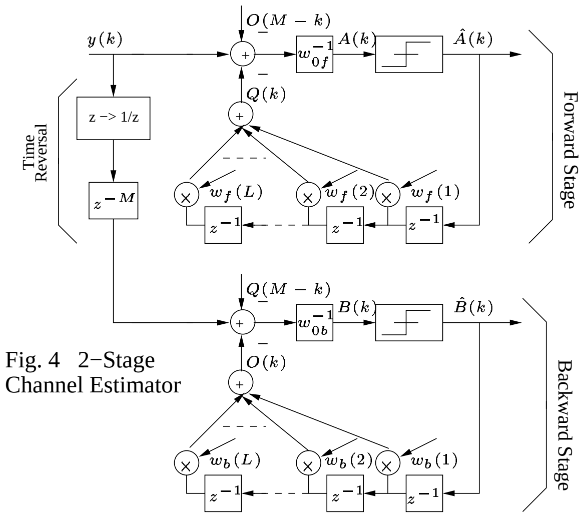

Fig. 4 2−Stage Channel Estimator