A LOW COMPLEXITY DECISION FEEDBACK EQUALIZER FOR WIDEBAND WIRELESS APPLICATIONS

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ABSTRACT

In this paper a new efficient Decision Feedback Equalizer (DFE) is proposed. The new technique is appropriate for channels with long and sparse impulse response (IR) as those encountered in many wireless communications applications. The proposed technique consists of two distinct parts. In the first one, the time delays of the multipath components are estimated by properly exploiting the form of the channel impulse response in the frequency domain. In the second part, a modified DFE scheme is applied to the channel output. The involved feedback filter has a significantly reduced number of taps, which are selected so as to act only on time positions associated with the estimated time delays of the involved multipath components. The new DFE exhibits considerable computational savings and faster convergence as compared to the conventional DFE, offering the same steady state performance.

1. INTRODUCTION

In many wireless communication systems the involved multipath channels exhibit a long time dispersion and delay spreads of up to $40\mu s$ are often encountered. If a wideband signal is transmitted through such a highly dispersive channel then the introduced Intersymbol Interference (ISI) has a span of several tens up to hundreds of symbols. This in turn implies that quite long adaptive equalizers are required at the receiver's end in order to reduce effectively the ISI component of the received signal. Wideband mobile communication systems and digital video terrestrial transmission are typical applications of the kind. In the latter case the involved channel impulse response (IR) may last up to several hundreds of baud intervals. Note that the situation is even more demanding whenever the channel frequency response exhibits deep nulls.

The adaptive DFE has been widely accepted as an effective technique for reducing ISI [1]. Moreover, it has been shown that the DFE structure is particularly suitable for multipath channels, since most part of ISI is due to the long postcursor portion of the IR (see for instance [7]). Recall that an important feature of the DFE is that the postcursor ISI is almost perfectly cancelled by the Feedback (FB) filter, provided of course that the previous decisions are correct. Moreover since noise is involved only in the output of the Feedforward (FF) filter, the DFE exhibits less noise enhancement effects as compared to linear equalizers. In applications of the type described above the DFE has a large number of taps (mainly due to the long FB filter). Thus, in high speed wireless applications, not only the implementation of a real time equalizer becomes a difficult task due to the very small symbol period, but also the equalizer itself has an increased complexity.

During the last decade there have been many efforts in different directions towards developing efficient implementations of the DFE. As such directions we mention IIR methods, block adaptive implementations, efficient algebraic solutions, modified DFE schemes etc. The works [2]-[9] are typical examples of these efforts. In the applications of interest, the involved multipath channel has a discrete sparse form. Efficient DFE schemes which exploit the sparseness of the IR have been derived in [10]-[13].

Recently a new DFE algorithm, appropriate for sparse multipath channels, was presented in [14]. This al-

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gorithm consists of two steps. In the first step, the time delays of the multipath components are estimated by properly exploiting the channel IR form. In the second step the DFE is applied, with the FB filter having a significantly reduced number of taps. These taps are selected so as to act only on time positions associated with the estimated time delays of the involved multipath components. Unfortunately, the performance of this algorithm deteriorates in cases the channel IR contains strong precursor echoes.

In this paper, a new DFE scheme is proposed, extending the approach introduced in [14]. The new technique overcomes the drawbacks of the previous work and retains its good performance even unter difficult conditions. This technique consists of the following four steps. First, the time delays of the multipath components are detected as in [14]. Second, the respective IR coefficients are estimated in an adaptive fashion. In the third step, a significant part of the postcursor ISI is removed before FF filtering, and finally in the forth part, a reduced size DFE with a sparse FB filter is applied to complete the equalization process. Moreover several other important issues are investigated such as, threshold determination for the detection of multipath components, tracking behavior, etc.

The paper is outlined as follows. In Section 2 the multipath channel is described and a useful formula for the inverse channel is derived. In Section 3 the method for estimating the time delays is presented and the new DFE is developed. Finally, in Section 4 some indicative experimental results are provided.

2. THE INVERSE MULTIPATH CHANNEL

In the applications of interest, the multipath channel is assumed to be invariant, or at least wide sense stationary, over a relatively small-scale time or distance interval. In such a case the baseband IR of the multipath channel may be expressed in the simple form

$$h(t) = \sum_i \alpha_i e^{j\theta_i} \delta(t - \tau_i) \tag{1}$$

where α_i and τ_i are the real amplitude and excess delay, respectively, of the $i-th$ multipath component, and θ_i is the corresponding phase shift (due to propagation along the $i-th$ path plus any other phase shifts). Then without loss of generality, the symbol spaced IR of the overall channel (including transmitter and receiver filters) can be written as

$$h(n) = \sum_{l=0}^{L} h_{n_l} \delta(n - n_l) \tag{2}$$

where L is the number of the dominant IR components appearing at the symbol spaced time instants, h_{n_l} is the complex amplitude of the $l-th$ component and n_l its respective delay. Delay n_0 corresponds to the main signal ($n_0 = 0$), while the remaining ones correspond either to causal ($n_l > 0$) or to anticausal ($n_l < 0$) components. The symbol spaced IR spans k_1 precursor and k_2 postcursor symbols respectively. That is the symbol spaced channel IR can be written in the vector form $\mathbf{h} = [h_{-k_1} \ \ldots \ h_0 \ \ldots \ h_{k_2}]^T$. From the total of the $(k_1 + k_2 + 1)$ IR coefficients only L of them are nonzero, located at the n_l positions.

Let $G(\omega) = \frac{1}{H(\omega)}$ be the frequency response of the inverse channel. Then, by making the assumption that $\sum_{l=1}^{L} |h_{n_l}| < |h_0|$, after some manipulations we end up with the following approximation formula for the inverse channel IR

$$\begin{aligned} g(n) &\approx h_0^{-1}\delta(n) - h_0^{-2}\sum_{l=1}^{L} h_{n_l}\delta(n-n_l) \\ &+ h_0^{-3}\sum_{l=1}^{L}[h_{n_l}]^2\delta(n-2n_l) \\ &+ 2h_0^{-3}\sum_i\sum_{j\neq i} h_{n_i}h_{n_j}\delta(n-n_i-n_j) \end{aligned} \tag{3}$$

Note that the above assumption is valid in most practical situations of interest (see for instance the HDTV test channels in [15]). Furthermore, as will be seen later on, in the case of the proposed DFE, this assumption is significantly relaxed. Formula (3) will be exploited in the following section to obtain the locations of the significant FB taps.

3. DERIVATION OF THE NEW ALGORITHM

Taking into account (2), the sampled output of the multipath channel can be written as follows

$$x(n) = \sum_{l=0}^{L} h_{n_l} u(n - n_l) + w(n) \tag{4}$$

where $\{u(n)\}$ is the i.i.d. symbol sequence with variance σ_u^2 and $\{w(n)\}$ is zero-mean complex white Gaussian noise independent of the input sequence. The channel span is considered to be N, i.e. $N = k_1 + k_2 + 1$. In a typical multipath channel, the postcursor part is much longer that the precursor part of the channel. The latter, in practice, consists of a small number of strong echoes (usually 1 or 2), located very close to the main signal (see for instance [12]).

As mentioned previously, the DFE is an appropriate equalization structure for the multipath environment under study. The conventional LMS-based adaptive DFE is given by the following set of equations

$$\hat{u}(n) = \sum_{k=-M+1}^{0} c_k(n)x(n-k) + \sum_{k=1}^{N} b_k(n)\tilde{u}(n-k) \quad (5)$$

$$\tilde{u}(n) = f\{\hat{u}(n)\} \quad (6)$$

$$e(n) = \hat{u}(n) - \tilde{u}(n) \quad (7)$$

$$c_k(n+1) = c_k(n) + \mu^c x(n-k)e(n)\ , \quad k = -M+1, \ldots, 0 \quad (8)$$

$$b_k(n+1) = b_k(n) + \mu^b \tilde{u}(n-k)e(n)\ , \quad k = 1 \ldots, N \quad (9)$$

where $\{x\}$ and $\{\tilde{u}\}$ denote the equalizer's input and decision sequences, respectively, c_k are the coefficients of the M-length FF filter, and b_k are the coefficients of the N-length FB filter (N is equal to the channel span [6]). $f\{\cdot\}$ is the decision device function, and μ^c, μ^b, are the step sizes.

Recall that, following an MSE approach and assuming that the previous decisions are correct, the FB coefficients can be expressed as [1]

$$b_k = -\sum_{j=-M+1}^{0} c_j h_{k-j} \quad k = 1, 2, \ldots, N \quad (10)$$

The main idea behind the technique described in this paper is to take advantage of the special form of the channel and reduce the computational complexity by considering only the effect of the non-zero coefficients of the channel. Such an approach, however, presumes a technique for determining the positions n_l of the dominant IR components. In the following, we present a method for obtaining the unknown n_l, based on a frequency domain expression for the cross-correlation between the channel input and output.

3.1. Estimation of the time delays

A well-established non-parametric procedure for estimating the time delays of the IR components is based on a proper crosscorrelation of the training symbols with the corresponding channel output samples. In a time domain implementation, the estimation of the crosscorrelation sequence for N lags requires $O(N)$ operations per sample. To reduce this complexity, we adopt a frequency domain scheme of complexity $log(N)$ per sample. The proposed scheme stems from an appropriate partitioning of both channel input and output sequences and is described below.

Let us first formulate the following $2N$-length DFT sequences for $k = 0, 1, \ldots, 2N-1$

$$U(k) = \sum_{m=p}^{N+p-1} u(n+m)e^{-j\frac{2\pi}{2N}mk} \quad (11)$$

$$X(k) = \sum_{i=0}^{2N-1} x(n+i)e^{-j\frac{2\pi}{2N}ik} \quad (12)$$

where p is an overestimated value of the non-causal size of the channel IR. Note that $X(k)$ is based on $2N$-length output sequence, while $U(k)$ results from an N length input sequence padded with N zeroes. If we consider the expected value of the product of the above sequences we get

$$\mathcal{E}\{X(k)U^*(k)\} = \sum_{i=0}^{2N-1} \sum_{m=p}^{N+p-1} \mathcal{E}\{x(n+i)u^*(n+m)\}e^{-j\frac{\pi}{N}(i-m)k} \quad (13)$$

where $\mathcal{E}\{\cdot\}$ denotes the expectation operator. If we now substitute eqn. (4) to eqn. (13) and take into consideration that the input sequence is i.i.d. and independent of the noise, then after some manipulations we get

$$\mathcal{E}\{U^*(k)X(k)\} = N\sigma_u^2 \sum_{l=0}^{L} h_{n_l} e^{-j\frac{\pi}{N}n_l k} \quad (14)$$

for $k = 0, 1, \ldots, 2N-1$. That is, we end up with a sum of complex sinusoids at normalized frequencies $n_l/2N$. Applying the $2N$-IDFT to the resulting sequence, the echo locations n_l are determined theoretically at the non-zero points of the IDFT.

In a practical situation, time averaging is used instead of $\mathcal{E}\{\cdot\}$ and a forgetting factor λ is included to compensate for time variations in the channel IR. The expression of the crosscorrelation in the frequency domain is then given by

$$C_{UX}^{(R)}(k) = \sum_{r=0}^{R-1} \lambda^{(R-1-r)N} X_r(k)U_r^*(k) \quad (15)$$

where $\lambda \leq 1$ and

$$X_r(k) = \sum_{i=0}^{2N-1} x(n+rN+i)e^{-j\frac{\pi}{N}ik}$$

$$U_r(k) = \sum_{m=p}^{N+p-1} \lambda^{N+p-m-1} u(n+rN+m)e^{-j\frac{\pi}{N}mk}$$

When a new N-length block of input-output samples is available, $C_{UX}^{(R)}(k)$ can be updated as follows

$$C_{UX}^{(R+1)}(k) = \lambda^N C_{UX}^{(R)}(k) + X_R(k)U_R^*(k) \quad (16)$$

We see from eqn. (15) that RN samples of $\{u\}$ and $\{x\}$ are used to compute $C_{UX}^{(R)}(k)$. The $L+1$ IDFT points of (15) with the highest amplitudes are then chosen as the desired locations. The number L of the undesired IR components can be preset by the designer by taking into account a worst case scenario for the specific application. Alternatively, L can be computed from the data using rank determination techniques. Another more realistic strategy would be to set a threshold and select the locations of the IDFT points of (15) having amplitude which exceeds this threshold.

It can be shown that the choice of a proper threshold quantity depends on the channel and input sequence characteristics. Let us consider the case of a sparse channel with real coefficients and an input sequence taking values from a binary alphabet ± 1. Then, it can be proven that the estimated cross-correlation, for a specific lag m, is a r.v. with normal distribution $\mathcal{N}(h_{n_k}, \frac{1}{K}\sum_{i \neq k} h_{n_i}^2)$, if $m = n_k$, and $\mathcal{N}(0, \frac{1}{K}\sum_{i=0}^{L} h_{n_i}^2)$ otherwise, where K is the total number of samples used. Thus, in order to be able to estimate all the existing echo positions with probability $\mathcal{P}$, the threshold t should be chosen as follows

$$t \leq (h_{n_k})_{min} - q\sqrt{\frac{\sum_{i \neq k_{min}} h_{n_i}^2 + \sigma_w^2}{K}} \qquad (17)$$

where q depends on $\mathcal{P}$, $(h_{n_k})_{min}$ is the minimum amplitude channel coefficient and σ_w^2 is the variance of the additive white gaussian noise.

3.2. The algorithm

In the proposed DFE, we focus our attention to the demanding FB part and reduce the computational load by properly selecting $O(L)$ number of taps out of N taps. Then the FB filter operation is replaced by an equivalent procedure which is applied to a restricted set of tap positions.

In the initial stage of the algorithm, the method described in the previous section is used for an adequate number of blocks $R = R_0$ and the time delays n_l are determined. Such an approach introduces a delay to the algorithm, which increases as the number of blocks R_0 increases. However the greater the parameter R_0, the higher is the degree of accuracy in selecting the correct positions. As it will be shown below, the initial delay of the algorithm is fully compensated by the fast convergence achieved by the new DFE.

It can be shown that for the type of sparse channels under consideration the MMSE FF filter approximates the anti-causal part of the inverse channel IR. Thus, having estimated the locations of the non-negligible channel IR coefficients, and combining Eqs. (3) and (10), we can easily determine the non-zero FB tap positions. Indeed, if we keep up to first order terms for the anti-causal inverse channel IR, we get the FB filter coefficients up to second order approximation. Specifically, it turns out that [14]:

a) there are "primary" non-zero taps at the positions where $n_l > 0$ in the channel IR (causal components) , and

b) for each "primary" non-zero tap $n_l > 0$, there are "secondary" non-zero taps at the positions $n_l - |n_k|$, where $n_k < 0$ are positions of the anticausal components in the channel IR.

Consequently, the FB filter can be restricted to act only to the above positions. Thus, we are led to an algorithmic scheme of low computational complexity, without sacrificing the performance of the full DFE [14].

In cases where there exist strong precursor echoes in the channel IR, the first order approximation of the inverse channel IR may not be sufficient. As a result a number of FB taps which were considered equal to zero, may now have non-negligible values, hence they should not be ignored. One possible solution is to consider a higher number of terms in the expression of the inverse channel IR. Such an approach, though, leads to an increase of the computational load required in the FB section of the DFE. The drawback of the algorithm in [14] can be overcome if the ideas presented in the previous sections are applied to the modified DFE which was introduced in [12].

The main feature of the so-called PFE equalizer presented in [12] is that a significant portion of the postcursor ISI is removed before FF filtering. Specifically, if we assume that the causal channel IR coefficients h_k are available, the operation of the DFE in the general case can be equivalently described by the following equations

$$\tilde{x}(n) = x(n) - \sum_{k=M}^{N} h_k \tilde{u}(n-k) \qquad (18)$$

$$\hat{u}(n) = \sum_{k=-M+1}^{0} c_k(n)\tilde{x}(n-k) + \sum_{k=1}^{M-1} b_k(n)\tilde{u}(n-k) \qquad (19)$$

$$\tilde{u}(n) = f\{\hat{u}(n)\} \qquad (20)$$

$$e(n) = \hat{u}(n) - \tilde{u}(n) \qquad (21)$$

$$c_k(n+1) = c_k(n) + \mu^c \tilde{x}(n-k)e(n), \quad k = -M+1, \ldots, 0 \qquad (22)$$

$$b_k(n+1) = b_k(n) + \mu^b \tilde{u}(n-k)e(n), \quad k = 1, \ldots, M-1 \qquad (23)$$

We observe from these equations that the input to the equalizer is the sequence $\{\tilde{x}\}$, which results after removing from the channel output signal $\{x\}$ a significant part of the postcursor ISI. Note that the size of the FB filter is $M-1$, where we recall that M is the

size of the FF filter.
In a sparse multipath scenario, the postcursor ISI is due to the restricted set of non-zero coefficients of the causal part of the channel IR. The required non-zero causal IR coefficients can be obtained by applying an LMS-based channel estimation procedure. Since the time delays of the multipath components have already been detected, the LMS channel estimator is restricetd to the respective positions. The input to the equalizer (eqn. (18)) is then written as

$$\tilde{x}(n) = x(n) - \sum_{\hat{n}_l \geq M} \hat{h}_{\hat{n}_l} \tilde{u}(n - \hat{n}_l) \qquad (24)$$

where $\hat{h}_{\hat{n}_l}$ and $\hat{n}_l$ are the estimated sparse channel coefficients and time delays respectively. Moreover, the second term in eqn. (19), which corresponds to the FB filter output, can be replaced by the output of a filter operating as described in the previous section (points a and b), but restricted to the range 1 to $M-1$. Note that by adopting this modified scheme, the expression for the inverse channel IR (eqn. (3)) is valid for a broader class of multipath channels. This is so because the equalizer input sequence $\{\tilde{x}\}$ corresponds to a sparse channel IR, whose causal part length equals to M. Thus the assumption that led to eqn. (3) now becomes $\sum_{n_l<M} |h_{n_l}| \leq h_0$, which is plausible in most practical cases.

The basic steps of the proposed algorithm are summarized below:

1. Compute $C_{UX}^{(R)}(k)$ from (15).
2. Compute the IDFT of $C_{UX}^{(R)}(k)$.
3. Estimate the echo locations.
4. Compute the secondary echo locations for $n_l < M$.
5. For the next N iterations:
   i) apply the LMS algorithm to estimate the sparse channel coefficients
   ii) compute the input to the equalizer from (24)
   iii) apply the DFE with the FBF acting to a restricted set of tap positions in the range 1 to $M-1$.
6. Update $C_{UX}^{(R)}(k)$ and repeat from step 2.

In Table 1, the computational complexity (number of complex multiplications per sample) of the proposed DFE is compared to that of the conventional DFE, and the algorithm described in [14], under the assumption that N is a power of 2. In Table 1, L_1, L_2 stand for the number of causal and non-causal echoes respectively

Algorithm	Computational Complexity
Conventional DFE	$2M+2N$
DFE of [14]	$2M+2L_1(L_2+1)+3\log_2(N)+5$
New DFE	$2M+2(L+1)+2L_{11}(L_2+1)+$ $L_{12}+3log_2(N)+5$

Table 1: Number of complex multiplications

$(L_1+L_2=L)$. L_{11} is the number of causal echoes for which $n_l < M$ and $L_{12} = L_1 - L_{11}$. It can be easily verified that the new DFE and the algorithm of [14] have similar complexities which are significantly lower compared to that of the conventional DFE. For instance, if $M=30$, $N=256$, $L_{11}=2$, $L_{12}=5$ and $L_2=2$ the required multiplications for the DFEs of Table 1 are 572, 131 and 126 respectively. As illustrated in the next section, the performance of the proposed DFE is superior compared to the performance of both other schemes, especially for channels with strong precursor echoes.

4. SIMULATION RESULTS - DISCUSSION

The channel used in our experiments consisted of 5 undesired multipath components with amplitudes $-14dB$, $-15\ dB$, $-17dB$, $-18dB$, and $-16dB$ respectively, and the corresponding time delays with respect to the main peak were $-6T$, $20T$, $37T$, $58T$ and $91T$, respectively, where T is the symbol time interval. The echo phases were chosen randomly. Note the presence of a strong precursor echo in the channel IR. The input to the channel was a QPSK signal, while white complex Gaussian noise was added to the channel output. The SNR was 30db. Three mean squared error curves are depicted in Fig.1. Curve a corresponds to the conventional DFE ($M=15, N=128$), curve b corresponds to the DFE scheme proposed recently in [14], and curve c to the proposed DFE algorithm. We observe that the new algorithm outperforms the conventional DFE in terms of convergence speed, for the same steady-state performance. The algorithm of [14] performs poorly in this case, because it neglects higher order terms which may have non-negligible values as explained in the previous section.

In order to investigate the tracking capabilities of the new algorithms in a time varying environment we consider the following scenario. A new postcursor component appears gradually in the channel IR after 7000 iterations. More specifically, in the interval $7000T - 7063T$ the amplitude of the new component increases linearly up to the value 0.192. In fig. 2, the MSE

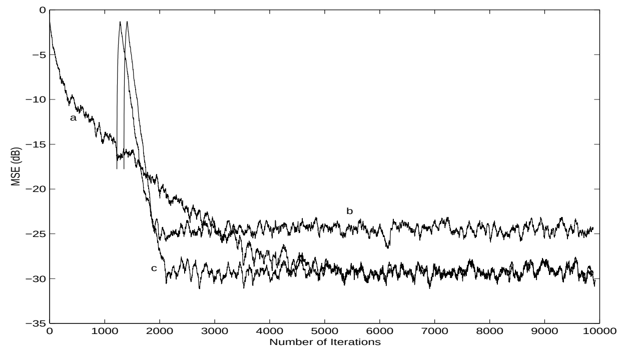

Figure 1: Mean squared error curves

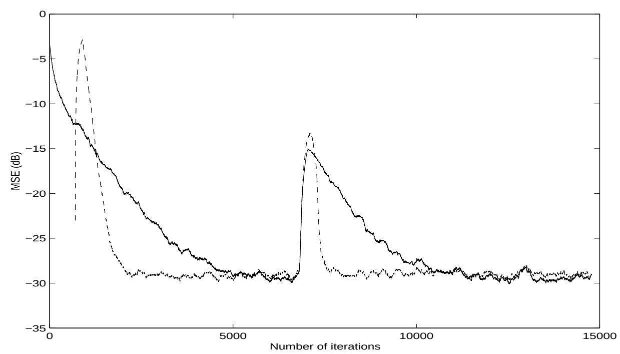

Figure 2: MSE curves under time varying conditions

of the new algorithm (dashed line) with $\lambda = 0.999$, is compared to that of the conventional DFE. We see that the proposed algorithm tracks immediately the change in the environment and converges very fast to the steady-state. On the contrary, the long FB filter of the conventional DFE drastically affects its tracking ability, resulting in slow convergence of the MSE to the steady-state error.

In conclusion, the new DFE offers considerable savings in complexity, fast convergence, and robustness with respect to channel conditions.

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