복소수 판정궤환 필터를 이용한 8-VSB 신호의 채널등화

論 文
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Equalization of 8-VSB Signals using Complex-Valued Decision Feedback Filter

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Abstract - In this paper, we present an equalization scheme for 8-VSB signals for the ATSC DTV system. We propose a complex feedback filter and complex feedback sample generator for DFE to equalize 8-VSB signals in order to efficiently remove multipath distortions causing leakages from the qudrature component. We show that the proposed structure outperforms the conventional DFE used for the digital VSB which uses a real-valued feedback filter with real-valued decisions.

Key Words : Decision Feedback Equalizer, Adaptive Equalization, Vestigial Side Band, 8-VSB, Feedback Sample Generator, Hilbert Transformation Filter.

1. Introduction

The decision feedback equalizer (DFE) is a prevailing equalization scheme for digital data broadcasting receivers, where severe and long delay spread multipath channels are one of major obstacles. DFE outperforms linear equalizer with relatively simple structure. Especially in ATSC DTV systems, DFE plays a critical role to recover received 8-vestigial side band (VSB) signals [1].

Since 8-VSB is generated from 8-PAM by reducing the redundant half in frequency domain via Hilbert Transform filter [2], the conventional DFE for 8-VSB uses real-valued feedback equalizer after projecting the received complex 8-VSB signals to the real axis [1] [3]. Specifically, for the 8-PAM signal s_k, the quadrature component of 8-VSB is given by $\hat{s}_k = \sum_{k=-\infty}^{\infty} h[m] s_{k-m}$. where $h[m]$denotes Hilbert transform filter [4], and the VSB signal is given by $v_k = s_k + j\hat{s}_k$.

In the presence of multipath, however, the effective channel for the projected signal into the real axis becomes infinite length in practice, since the real component usually the quadrature component as well. For example, consider a baseband channel model $[1\ 0.1j]$. Then the received baseband VSB signal is given as $r_k = v_k + 0.1jv_{k-1}$ and its real projection is

$$Re(r_k) = s_k + 0.1\hat{s}_{k-1} \quad (1)$$

Hence, the effective channel for the real component becomes $\delta(0) + 0.1h * \delta(-1)$, i.e. infinite length due to the Hilbert Transform filter. Since it is measure zero event that a baseband channel is real valued only, the DFE with a real-valued feedback for 8-VSB signals always should deal with infinite length channel.

In this paper, we propose a DFE for 8-VSB signals equipped with a complex-valued feedback filter and a complex-valued feedback sample generator to overcome the drawback of the real-valued DFE. Simulation results shows that the proposed DFE successfully cancels post-cursor multipath distortion in the presence of the leakage from the quadrature component. The complex DFE outperforms the real-valued DFE in consideration of the fact that the computational complexity of a length N complex DFE is equivalent to a length $2N$ real-value DFE,

2. Complex DFE for 8-VSB

Consider a DFE system illustrated in Figure 1.

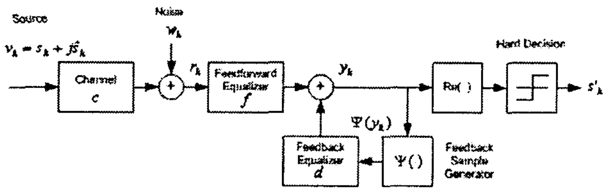

Fig. 1 Proposed Complex DFE structure for 8-VSB signals

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Let a complex vector $c=[c_0, c_1, \cdots c_{N_c-1}]$ denote the baseband channel model. Then the received signal is

$$r_k=\sum_{n=0}^{N_c-1} c_n v_{k-n}+w_k \tag{2}$$

We use a complex-valued feedforward filter f as the conventional DFE for VSB [3], but use a complex feedback filter d. To feedback through the complex feedback filter, we generate a complex feedback sample as the following

$$\Psi(y_k)=D(Re(y_k))+jIm(y_k), \tag{3}$$

where $D(\cdot)$ denotes the hard decision for 8-PAM (the minimum distance detector). Notice that the proposed feedback sample generator uses identity function for the imaginary and the hard decision for the real component. Then, the equalizer output is written as

$$y_k=\sum_{n=0}^{N_f-1} f_n r_{k-n}-\sum_{n=1}^{N_d-1} d_n \Psi(y_{k-n})+\sum_{n=0}^{N_f-1} f_n w_{k-n} \tag{4}$$

In this DFE structure, we define the optimal filter coefficients of f, d as the ones minimizing the decision error as the following cost function;

$$J(f,d)=E\{(Re(y_k)-D(Re(y_k))^2\} \tag{5}$$

To obtain the optimal coefficients, an adaptive approach can be used with the following update equations [5]

$$\begin{aligned} f_n(k+1)&=f_n(k)+\mu r_k^* Err(k), \\ d_n(k+1)&=d_n(k)+\mu\Psi^*(y_k)Err(k) \end{aligned} \tag{6}$$

where μ denotes a step-size and $Err(k)$ denotes instaneous error at k, $Err(k)=Re(y_k)-D(Re(y_k))$.

To theoretically evaluate the performance of the proposed DFE, one should show that the minima of the cost function (5) locate at the desired places. For example, in the noise free case, f, d should be Zero-Forcing DFE [6]. However, theoretical analysis of the cost function (5) is an extremely difficult task, since (5) has an IIR term with respect to y_k. Hence, we consider a simple one tap case for theoretical analysis and will provide simulation results for the general cases.

For a two-tap channel $[1\ c]$, we consider a one-tap feedback filter d, with $f=1$ in the absence of noise. and decision error. Then

$$y_k=v_k+cv_{k-1}-d\Psi(y_{k-1}) \tag{7}$$

and by the recursive nature of y_k, we have

$$\begin{aligned} Re(\Psi(y_k))&=s_k \\ Im(\Psi(y_k))&=\hat{s}_k+Im(cv_{k-1})-d_R Im(\Psi(y_{k-1})) \\ &=\hat{s}_k+\sum_{n=1}^{\infty}(-d_R)^{n-1} Im(\theta v_{k-n}) \end{aligned} \tag{8}$$

where $\theta=c-d$ denotes the parameter error and d_R denotes the real component of d. Hence,

$$\begin{aligned} Re(y_k)-D(Re(y_k))&=Re(y_k)-s_k \\ &=Re(v_{k-1}\theta)+d_I\sum_{n=2}^{\infty}(-d_R)^{n-2} Im(\theta v_{k-n}) \end{aligned} \tag{9}$$

where d_I denote the imaginary component of d, respectively. The cost function is now given by

$$\begin{aligned} J(d)&=E\{[(Re(y_k)-D(Re(y_k))]^2\} \\ &=E\{Re(\theta v_{k-1})^2\}+E\left\{\left[d_I\sum_{n=2}^{\infty}(-d_R)^{n-2} Im(\theta v_{k-n})\right]\right\} \\ &+2d_I\sum_{n=2}^{\infty}(-d_R)^{n-2}E\{Re(\theta v_{k-1})Im(\theta v_{k-n})\} \end{aligned} \tag{10}$$

From the nature of VSB signal, we have

$$E(\hat{s}_l\hat{s}_m)=\sigma^2\delta(l-m),\ E(s_l\hat{s}_m)=\sigma^2 h[m-l] \tag{11}$$

where $\sigma^2=E(s_k^2)$. Hence,

$$\begin{aligned} &E\{Re(\theta v_{k-1})^2\}=|\theta|^2\sigma^2 \\ &E\{Im(\theta v_{k-n})Im(\theta v_{k-m})\}=|\theta|^2\sigma^2\delta(n-m) \\ &E\{Re(\theta v_{k-1})Im(\theta v_{k-n})\}=-|\theta|^2\sigma^2 h[n-1] \\ &E\left\{\left[d_I\sum_{k=0}^{\infty}(-d_R)^k Im(\theta v_{k-2})\right]^2\right\}=|\theta|^2\sigma^2 d_I^2\sum_{n=2}^{\infty}(-d_R)^{2(n-2)} \end{aligned} \tag{12}$$

Therefore, finally,

$$\begin{aligned} J(d)&=|\theta|^2\sigma^2\left(\sum_{n=2}^{\infty}(h[n-1]+d_I d_R^{n-2})^2+\frac{1}{2}\right) \\ &=|\theta|^2\sigma^2\left(\frac{d_I^2}{1-d_R^2}-4d_I\sum_{n=0}^{\infty}\frac{d_R^{2n}}{\pi(2n+1)}+1\right) \end{aligned} \tag{13}$$

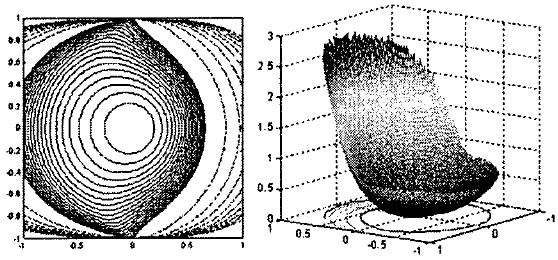

Fig. 2 Contour and Cost function of J(d)

Above equations show that I) $J(d)$ has a unique global minimum at $c=d$ $(\theta=0)$ as desired and the cost function is not diverge, hence the stochastic update algorithm is (statistically) stable. Figure 2 shows the contour of $J(d)$ for $-1 \leq d_R, d_I \leq 1$, and cost function.

3. Simulation Results

In this section we present simulation results showing the analysis for a one-tap feedback filter holds for general multi-tap cases even in the presence of noise.

The first channel is a following multipath channel in the absence of noise.

$$c = [1, 0, 0, 0, 0, 0.3j] \tag{14}$$

The feedback equalizer is initialized to 0 and adapted the update rule (6) with the step-size $\mu = 0.0007$ without help of training signals. In Figure 3 we compare the cluster variance($E\{(Re(y_k) - D(Re(y_k))^2\}$) trajectories of the proposed DFE output with feedback equalizer length $N_d = 5$, and convention DFE output with $N_d = 5$, and $N_d = 10$.

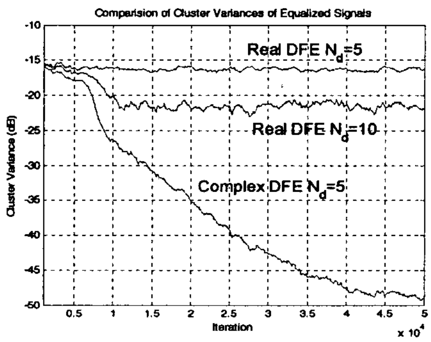

Fig. 3 Cluster Variance Comparison for c_1

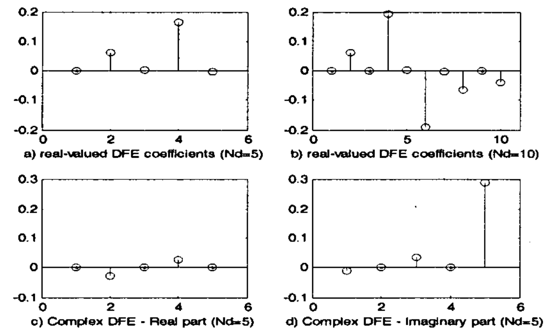

Fig. 4 Feedback Equalizer Coefficients Comparison

One can clearly observe outstanding performance of the proposed DFE. The converged coefficients plotted in Figure 4 explains why complex DFE works better. The conventional real-DFE converged to the truncated Hilbert converged to $0.3j$ to cancel the channel.

In the second simulation, we used a real-value dominating multi-tap multipath channel under 30dB SNR,

$$c_2 = [1,0,0.1+0.05j,-0.2+0.1j,0,0,0.1,0,0, \\ -0.1,0.1-0.06j] \tag{15}$$

Figure 5 shows the cluster variance trajectories of complex-valued DFE with $N_d = 10$, and real-valued DFE with $N_d = 10$ and $N_d = 20$. For this channel, the proposed DFE achieves better cluster variance (almost the noise floor) than the conventional DFE.

4. Conclusion

In this paper, we have proposed a DFE equipped with a complex feedback filter and complex feedback sample generator for digital VSB signals. Simulation results show that the proposed DFE outperforms the conventional DFE when the multipath causes quadrature leakage. Cost function analysis is done for a simple one-tap feedback filter case showing global convergence of the proposed algorithm.

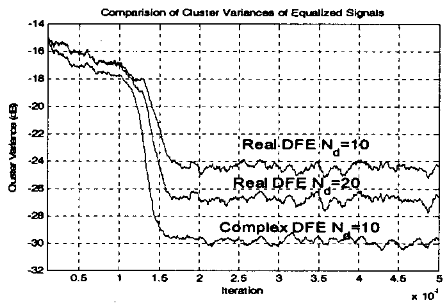

Fig. 5 Cluster Variance Comparison for c_2

References

[1] E. A. Henderson, et al "ATSC DTV receiver implementation," *Proceedings of the IEEE*, pp. 119-147, Jan. 2006.

[2] J. Proakis, *Digital Communications* .McGraw-Hill, 1995

[3] T. Endres et. al., "Carrier independent blind initialization of a DFE,"*IEEE Signal Processing Workshop SPAWC*, pp. 239--242, May 1999.

[4] A.V. Oppenheim and R.W. Shafer, *Discrete-Time Signal Processing*, Prentice Hall, 1998.

[5] J.R. Treichler, C.R. Johnson, Jr., and M.G. Larimore, *Theory and Design of Adaptive Filters*. Prentice Hall, 2001.

[6] J.M. Cioffi, G.P. Dudevoir, M.V. Eyuboglu, G.D. Forney Jr., "MMSE Decision-Feedback Equalizers and Coding, Part I: Equalization Results," *IEEE Trans. on Commun.*, vol 43, NO. 10, Oct. 1995.