A Fast Selective-Direction MMSE Timing Recovery Algorithm for Spatial-Temporal Equalization in EDGE

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Abstract

For the EDGE system with multiple antennas, spatial-temporal equalization can reduce the effect of multipath fading, intersymbol interference and co-channel interference, thereby increasing the capacity and range. With time varying delay spread, accurate timing recovery is crucial for good equalizer performance especially when the equalizer length is short because of the multiple receivers with a limited number of training symbols. In this paper, we propose a fast selective-direction minimum mean-square error (MMSE) timing recovery algorithm. The new timing recovery algorithm determines the estimated burst timing and processing direction for the equalizer by computing the MSE for a decision feedback equalizer in both the forward and reverse time directions. Simulation results show that a 2-branch receiver with our techniques requires about 3 dB lower signal-to-interference ratio than a previous approach for 1% raw BER in EDGE.

I. Introduction

The radio interface EDGE, Enhanced Data rates for Global Evolution, is currently being standardized as an evolutionary path from GSM and TDMA-IS136 for third-generation high-speed data wireless systems. A major limitation on the system capacity of wireless systems such as EDGE is intersymbol interference (ISI), caused by multipath fading, and co-channel interference (CCI). Spatial-temporal equalization (STE) using multiple antennas is an effective approach to jointly suppress ISI and CCI [1], [2], [3], [4].

As shown in Figure 1, the STE [1] uses space-time prefilters for combining, followed by a temporal equalizer for signal detection. In EDGE systems, the burst and symbol phase at the receiver vary from burst to burst due to multipath fading. Burst and symbol timing recovery is crucial for good equalizer performance especially when the prefilter length is short because of the multiple receivers with a limited number of training symbols. In addition, since the impulse responses of fading channels are usually not symmetric, a selective time-reversal nonlinear equalizer can improve the performance [5]. In [5], a timing recovery algorithm based on exhaustive search is proposed to determine the estimated burst and symbol timing and to select the processing direction as either the forward or reverse time direction. However, this method requires multiple equalizer training which is computationally costly especially in EDGE systems. In contrast to the exhaustive search method, fast computation of the MSE for a decision-feedback equalizer (DFE) has been developed by exploring the DFE structure [6]. The main contribution in this paper is to extend the above method to a selective time-reversal equalizer in EDGE. Our new timing recovery algorithm computes the MSE in both the forward and reverse time directions. The estimated burst and symbol timing as well as the processing direction are then determined to minimize the MSE.

The performance of the modified receiver using the MMSE timing recovery algorithm is evaluated for EDGE. At 1% raw BER, a 2-branch receiver requires about 3 dB lower signal-to-interference ratio (SIR) than the previous approach [1] for the hilly terrain (HT) channel profile. Compared with a 1-branch receiver, the 2-branch receiver requires a 4 dB lower signal-to-interference-plus-noise ratio (SINR) in the noise dominant case, In the interference dominant case, the gain

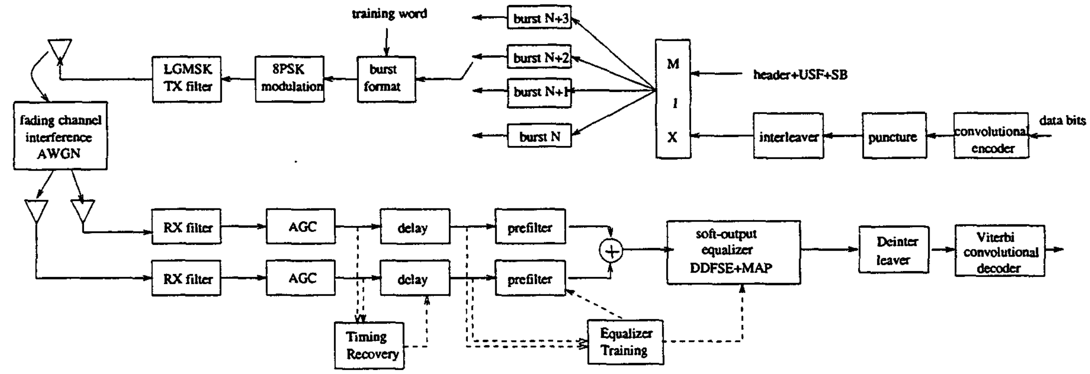


Fig. 1. EDGE system diagram with the STE.

is 10 dB for the HT profile and 30 dB for the typical urban (TU) channel profile. Thus the 2-branch receiver using the proposed timing algorithm can significantly suppress strong CCI, especially in the typical urban environment, and therefore is an efficient technique for EDGE capacity enhancement.

The rest of this paper is as follows: In Section II we derive the new MMSE timing algorithm. In Section III we demonstrate the performance improvement of the new timing algorithm, as well as 2-branch interference suppression, via simulations. Conclusions are given in Section IV.

II. Fast Selective-Direction MMSE Timing Recovery Algorithm

As shown in Figure 1, our STE has prefilters followed by a temporal equalizer. The prefilters suppress noise and cochannel interference and also shorten the overall system impulse response for the temporal equalizer. In particular, the weights of our STE are obtained based on the MMSE-DFE criterion where the prefilters and shortened channel are viewed as the feedforward filters and feedback filter of the DFE, respectively [1]. Furthermore, the STE operates in a preselected time direction as shown in Figure 2. Thus, our STE can be viewed as a delay-and-direction-optimized MMSE-DFE. The direction and the equalizer delay (i.e., burst timing) are obtained from the timing recovery. Below, we present our new timing recovery algorithm.

As shown in Figure 3, the T-spaced received signal sample at the AGC output in m-th branch is given by

$$x_m(k) \;=\; \sum_{i=-\bar{L}}^{L} h_m(i)s(k-i)+n_m(k) \qquad (1)$$

where $\{h_m(i), i=-\bar{L},\cdots,L\}$ is the m-th channel impulse response after the AGC, $s(k)$ is the k-th transmitted desired symbol, and $n_m(k)$ is the sample of the interference plus noise. The DFE has M (L_f+1)-tap feedforward filters and an L_b-tap feedback filter, where M is the number of receiver branches. The slicer input is given by

$$\begin{aligned}\hat{s}(k) \;=\; & \sum_{m=1}^{M}\sum_{i=0}^{L_f} f_m^*(i)x_m(k-d+L_f-i) \\ & -\sum_{i=1}^{L_b} f_b^*(i)s(k-i)\end{aligned} \qquad (2)$$

where integer d is the equalizer delay and $()^*$ denotes complex-conjugate transpose. Using vector representation, we have

$$\mathbf{x}(k) \;=\; \mathbf{H}\mathbf{s}(k)+\mathbf{n}(k) \qquad (3)$$

where

$$\mathbf{x}(k) \triangleq \begin{pmatrix} x_1(k-d+L_f) \\ \vdots \\ x_1(k-d) \\ \vdots \\ x_M(k-d+L_f) \\ \vdots \\ x_M(k-d) \end{pmatrix} \qquad (4)$$

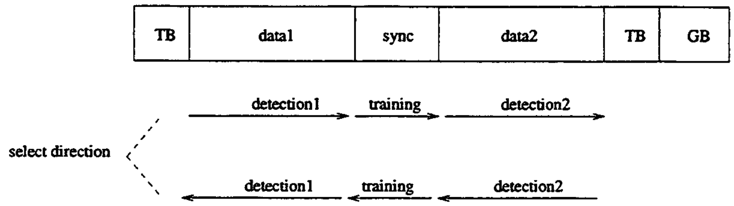


Fig. 2. Selective directional equalization.

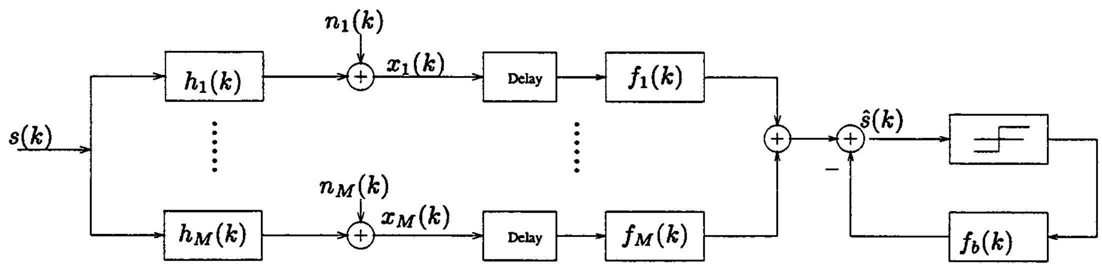


Fig. 3. Baseband signal representation of a DFE with M branches.

$$\mathbf{s}(k) \triangleq \begin{pmatrix} s(k-d+L_f+\bar{L}) \\ \vdots \\ s(k) \\ \vdots \\ s(k-d-L) \end{pmatrix} \quad (5)$$

$$\mathbf{H} \triangleq \begin{pmatrix} h_1(-\bar{L}) & \cdots & h_1(L) & & 0 \\ & \ddots & & \ddots & \\ 0 & & h_1(-\bar{L}) & \cdots & h_1(L) \\ & & \vdots & & \\ h_M(-\bar{L}) & \cdots & h_M(L) & & 0 \\ & \ddots & & \ddots & \\ 0 & & h_M(-\bar{L}) & \cdots & h_M(L) \end{pmatrix} \quad (6)$$

and $\mathbf{n}(k)$ is defined similarly to $\mathbf{x}(k)$. For convenience, define the feedforward and feedback coefficient vectors as follows:

$$\mathbf{f} = [f_1(0), \cdots, f_1(L_f), \cdots, f_M(0), \cdots, f_M(L_f)]^T \quad (7)$$

$$\mathbf{b} = [\underbrace{0, \cdots, \cdots, 0}_{-d+L_f+\bar{L}}, 1, \underbrace{f_b(1), \cdots, f_b(L_b)}_{\mathbf{f}_b^T}, \underbrace{0, \cdots, \cdots, 0}_{d+L-L_b}]^T \quad (8)$$

Under the assumption of correct past decisions, the error in the slicer input is given by

$$e(k) \;=\; \hat{s}(k) - s(k) = \mathbf{f}^*\mathbf{x}(k) - \mathbf{b}^*\mathbf{s}(k). \quad (9)$$

To derive the optimal values of $\mathbf{f}$, $\mathbf{x}(k)$ and $\mathbf{b}^*\mathbf{s}(k)$ are considered as the input and desired signal, respectively. From adaptive filter theory [7, p. 170], the MSE is given by

$$\begin{aligned} MSE &= E\{|\mathbf{b}^*\mathbf{s}(k)|^2\} - E\{\mathbf{x}(k)\mathbf{b}^*\mathbf{s}(k)\}^* \\ &\quad E\{\mathbf{x}(k)\mathbf{x}(k)^*\}^{-1}E\{\mathbf{x}(k)\mathbf{b}^*\mathbf{s}(k)\} \\ &= \mathbf{b}^*\underbrace{(\mathbf{I} - \mathbf{H}^*(\mathbf{H}\mathbf{H}^* + \mathbf{R}_{nn})^{-1}\mathbf{H})}_{\mathbf{\Phi}}\mathbf{b} \end{aligned} \quad (10)$$

where $\mathbf{R}_{nn} \triangleq E\{\mathbf{n}(k)\mathbf{n}(k)^*\}$ is the covariance matrix of the noise vector, and the transmitted signal samples are assumed uncorrelated, i.e., $E\{\mathbf{s}(k)\mathbf{s}(k)^*\} = \mathbf{I}$. We partition $\mathbf{\Phi}$ according to

$$\mathbf{\Phi} = \begin{pmatrix} \times & \times & \times & \times & \times \\ \times & p & \mathbf{q}^* & \times & \times \\ \times & \mathbf{q} & \mathbf{P} & \mathbf{r} & \times \\ \times & \times & \mathbf{r}^* & t & \times \\ \times & \times & \times & \times & \times \end{pmatrix} \quad (11)$$

where p is the $(-d+L_f+\bar{L}+1)$-th diagonal element, and t is the $(-d+L_f+\bar{L}+1+L_b+1)$-th diagonal element. The MSE and MMSE at delay d are then given by

$$MSE(d) \;=\; (p - \mathbf{q}^*\mathbf{P}^{-1}\mathbf{q}) +$$

$$(\mathbf{f}_b + \mathbf{P}^{-1}\mathbf{q})^*\mathbf{P}(\mathbf{f}_b + \mathbf{P}^{-1}\mathbf{q}) \quad (12)$$

$$MMSE(d) = p - \mathbf{q}^*\mathbf{P}^{-1}\mathbf{q}, \quad (13)$$

respectively. The estimated equalizer timing $\hat{d}$ is the one that minimizes $MMSE(d)$.

From the above discussion, we need to calculate $\mathbf{\Phi}$ and $\mathbf{P}^{-1}$'s for different d's to obtain the delay estimation for the forward DFE. Below, we show that these $\mathbf{\Phi}$ and $\mathbf{P}^{-1}$'s can also be used for the time-reversal DFE. In the time-reversal DFE, the received signal samples and the transmitted symbols are reversed in time. The corresponding channel impulse is also reversed. Let $\bar{x}_m(k) = x_{M-m+1}(-k)$, $\bar{s}(k) = s(-k)$, $\bar{h}_m(k) = h_{M-m+1}(-k)$, where the overbar implies a time reversal system. It is straightforward to verify that

$$\bar{x}_m(k) = \sum_{i=-L}^{\bar{L}} \bar{h}_m(i)\bar{s}(k-i) + \bar{n}_m(k). \quad (14)$$

To derive the MSE of time-reversal DFE, it is convenient to define the transform

$$\mathbf{J} = \begin{pmatrix} 0 & & 1 \\ & \cdot^{\cdot^{\cdot}} & \\ 1 & & 0 \end{pmatrix} \quad (15)$$

with a given dimension. For a matrix $\mathbf{A}$, $\mathbf{JA}$ flips the rows of $\mathbf{A}$ from top to bottom. Similarly, $\mathbf{AJ}$ flips the columns of $\mathbf{A}$ from left to right. It can be shown that

$$\bar{\mathbf{H}} = \mathbf{JHJ}, \quad \bar{\mathbf{R}}_{nn} = \mathbf{JR}_{nn}\mathbf{J}, \quad \bar{\mathbf{\Phi}} = \mathbf{J\Phi J}. \quad (16)$$

By choosing $\bar{d} = (L_f + L_b + 1) - d$, we have

$$\bar{\mathbf{P}} = \mathbf{JPJ}, \quad \bar{p} = t, \quad \bar{\mathbf{q}} = \mathbf{Jr} \quad (17)$$

The corresponding MMSE is given by

$$\overline{MMSE}(\bar{d}) = \bar{p} - \bar{\mathbf{q}}^*\bar{\mathbf{P}}^{-1}\bar{\mathbf{q}} = t - \mathbf{r}^*\mathbf{P}^{-1}\mathbf{r}. \quad (18)$$

As seen from (13) and (18), $\mathbf{P}^{-1}$ is involved in the MMSE calculation for both directions, and thus only one matrix inversion is required.

III. Simulation

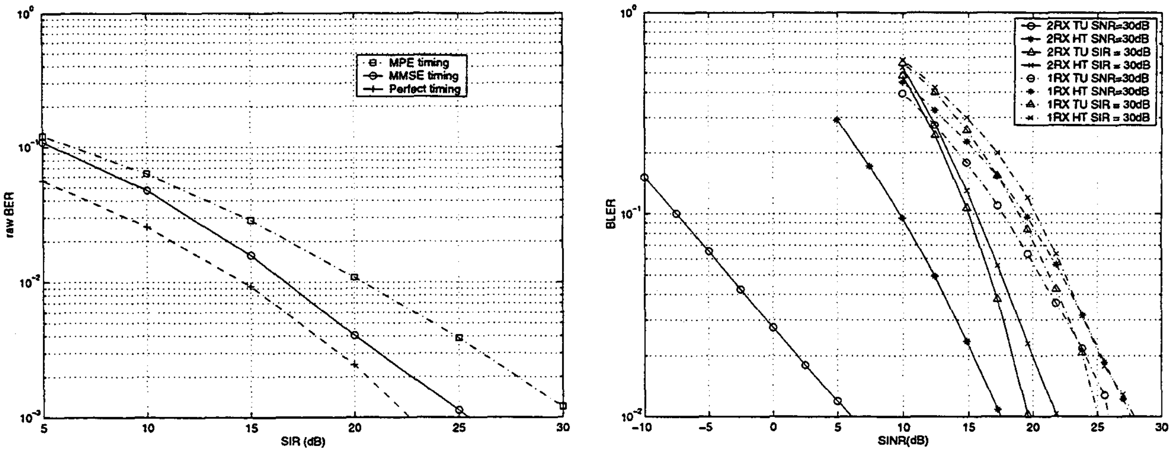
The performance of the new MMSE timing recovery algorithm was evaluated for the EDGE system shown in Figure 1. Modulation and coding scheme No. 5 (MCS 5) was chosen in the simulation [8]. In particular, a rate 1/3 convolutional code with constraint length of 7 was used. The burst format is the same as GSM. The modulation is 8-PSK with linearized GMSK pulse shaping, and the baud rate is 279.833 ksps The channel model is a multipath fading channel with a single interferer, a Doppler frequency of 4Hz, and no frequency hopping. In the receiver, the length of the prefilter and shortened channel are 5 and 6 respectively ($L_f = L_b = 5$). The soft-output equalizer uses an 8-state delayed decision-feedback sequence estimator (DDFSE) and 8-state maximum *a posteriori* probability (MAP) estimator [9].

In the simulation, we first compare the raw BER performance of the new MMSE timing recovery algorithm and a previous approach which estimated the symbol phase and equalizer delay by minimizing the ratio of precursor to cursor energy (MPE) [1], [10]. Figure 4 (a) shows the raw BER versus SIR for 2-branch receivers with the HT profile. The results are shown for the MMSE timing recovery algorithm (circle), the MPE timing recovery algorithm (square), and perfect MMSE timing recovery (plus). For a 10^{-2} BER, the required SIR with the MMSE timing recovery algorithm is 3 dB less than that required with the MPE approach. Note that this SIR is 2 dB higher than that required with perfect MMSE timing recovery.

Next, we study the system block error (BLER) performance for different channel conditions. Figure 4 (b) shows the BLER versus signal-to-interference-plus-noise ratio (SINR) for the typical urban (TU) and hilly terrain (HT) profiles. The results are summarized as follows:

- 1-branch receiver: The required SINR for 10% BLER is between 17 to 20 dB for all cases. The 1-branch receiver basically treats the CCI as the AWGN.
- 2-branch receiver: In the interference dominant cases (SNR=30dB), the required SINR for 10% BLER is -7.5 dB for TU, and 10 dB for HT. On the other hand, in the noise dominant case (SIR=30 dB), the required SINR is about 15 dB for both TU and HT.

Compared with 1-branch receiver, the 2-branch receiver has much greater CCI suppression especially in the TU case.



(a) (b)

Fig. 4. (a)Performance comparison of timing recovery algorithms with 2 antennas and the HT profile.(b) SINR performance of the modified EDGE receiver with the new MMSE timing recovery algorithm.

IV. Conclusion

In this paper, we have proposed a fast selective-direction MMSE timing recovery algorithm for spatial-temporal equalization. The new timing recovery algorithm computes the MMSE for DFE in both the forward and reverse time directions, and determines the estimated burst timing and processing direction. We applied the new algorithm for STE in EDGE. For a 2-branch receiver with 1% raw BER, the new timing algorithm requires 3 dB lower SIR than a previous approach. Furthermore, the 2-branch receiver with the proposed timing algorithm can significantly suppress strong CCI, especially in the typical urban environment, and therefore is an efficient technique for EDGE capacity enhancement.

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