JOINT EQUALIZATION AND INTERFERENCE SUPPRESSION FOR HIGH DATA RATE WIRELESS SYSTEMS

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Abstract - Enhanced Data Rates for Global Evolution (EDGE) is currently being standardized as an evolution of GSM in Europe and IS-136 in the United States as an air interface for high speed data services for third generation mobile systems. In this paper, we study space-time processing for EDGE to provide interference suppression. We consider the use of two receive antennas and propose a joint equalization and diversity receiver. This receiver uses front-end filters on each diversity branch to perform minimum mean-square error (MMSE) cochannel interference suppression, while leaving the intersymbol interference to be mitigated by the subsequent equalizer. The equalizer is a delayed decision feedback sequence estimator (DDFSE), consisting of a reduced-state Viterbi processor and a feedback filter. The equalizer provides soft output to the channel decoder after deinterleaving. We describe a novel weight generation algorithm and present simulation results on the link performance of EDGE with interference suppression. These results show a significant improvement in the signal-to-interference ratio (SIR) performance due to both diversity (against fading) and interference suppression. At a 10% block error rate, the proposed receiver provides a 20 dB improvement in SIR for both the typical urban and hilly terrain profiles.

I. INTRODUCTION

Enhanced Data Rates for Global Evolution (EDGE) is currently being standardized as an evolution of GSM in Europe and IS-136 in the United States as an air interface for high speed data services [1]. EDGE reuses the GSM time slot structure, carrier bandwidth (180.05 MHz), and symbol rate (270.833 kbaud), but can provide a 3 times higher data rate through the use of 8-PSK modulation with partial response pulse shaping. EDGE is being introduced as an IS-136 and GSM overlay using a 1/3, 3/9, or 4/12 reuse pattern (instead of the 7/21 reuse pattern in current IS-136 systems); thus, cochannel interference severely limits the radio link performance. Adaptive array techniques, using multiple receive antennas for interference suppression (as used in IS-136 (see, e.g., [2])), can mitigate this problem.

In this paper, we study the use of a joint equalization and diversity receiver to provide interference suppression in EDGE. The proposed receiver is similar to the original EDGE receiver [1], except for an additional receiving branch. The front-end filters of the diversity receiver perform minimum mean-square error (MMSE) cochannel interference suppression, while leaving the intersymbol interference to be mitigated by the subsequent equalizer. This equalizer is a delayed decision feedback sequence estimator (DDFSE), consisting of a reduced-state Viterbi processor and a feedback filter. This equalizer provides soft output to the channel decoder after deinterleaving. We describe a novel weight generation algorithm and present simulation results on the link performance of EDGE with interference suppression. Our results show a significant improvement in the signal-to-interference ratio (SIR) performance due to both diversity (against fading) and interference suppression. At 10% block error rate (BLER), the proposed receiver provides a 20 dB improvement in SIR for both typical urban and hilly terrain profiles.

In Section II we describe the system and in Section III describe the computer simulation model. In Section IV we present performance results. Conclusions are presented in Section V.

II. SYSTEM MODEL

The EDGE system (see, e.g., [1]) uses a TDMA format with a burst length of 576.92 μs, a frame of 8 bursts (5 ms), and a block for interleaving of 4 bursts out of each frame. The modulation currently being considered is 8-

0-7803-5565-2/99/$10.00

PSK with linearized GMSK pulse shaping, with a symbol rate of 270.833 kbaud (symbol period T = 3.692 μs). The receiver filter is a square-root Nyquist filter with a bandwidth of 180.05 kHz. Each burst contains 116 payload symbols, with 26 training symbols as a midamble, and 6 tail and 8.25 guard symbols. Coding is 1/3 rate convolutional coding with a constraint length of 7, with block interleaving over 20 ms.

Figure 1 shows a block diagram of our receiver. It is similar to the original EDGE receiver [1], except for an additional receiving branch. The front-end filters of the diversity receiver perform MMSE cochannel interference suppression, while leaving the intersymbol interference to be mitigated by the subsequent equalizer. This equalizer is a DDFSE, consisting of a reduced-state Viterbi processor and a feedback filter. This equalizer provides soft output to the channel decoder after deinterleaving.

The three key techniques used with this receiver are the soft output method, the timing recovery algorithm, and the equalizer weight training algorithm. These techniques are as follows.

Soft-Output DDFSE: The soft output of the DDFSE is computed using Lee's method [3] which is used to compute *a posteriori* probabilities recursively, as described in detail in [4].

Timing Recovery: Timing recovery is based on the technique described in [5]. This technique is optimum for a decision feedback equalizer (DFE) with a single feedforward tap, but it is also found to give near-optimum performance in a general case over all multipath delay profiles we have tested, compared to a brute-force timing search method. Therefore, $T/2$-spaced taps are not needed, i.e., the front-end filter can use symbol spaced taps. In this study, we use this timing recovery technique to determine the decision delay of the front-end filter (with common timing for both branches).

Equalizer Training: The front-end filter in Figure 1 has F symbol-spaced taps, while the Viterbi equalizer has a memory of μ, and the feedback filter has $B-\mu$ taps. Our receiver is trained as if it was a DFE with F feedforward taps per branch and B feedback taps, using the MMSE criterion. The method we use is based on the recursive least square (RLS) algorithm, where the weights are calculated using the training sequence, and then held fixed over the TDMA burst. After training the feedforward filter coefficients are used as the coefficients of the front-end filter, the first μ feedback coefficients $b_1, b_2, \cdots b_\mu$ are used to compute the metrics in the Viterbi equalizer, and the remaining coefficients $b_{\mu+1}, \cdots, b_B$ are used to set the feedback filter of the DDFSE. The rationale for this training method is as follows.

II.1. Rationale for the Space-Time Equalizer Training Algorithm

For the infinite-length case, according to [6,(54)], the optimum maximum likelihood sequence estimation (MLSE) and DFE filters have the following relationship,

$$\mathbf{W}_{MLSE} = \mathbf{W}_{DFE} \frac{1+\Gamma(f)}{C[1+\Gamma(f)]} \cdot \frac{C[\Gamma(f)]}{\Gamma(f)} \quad , \qquad (1)$$

where $\Gamma(f)$ is the signal-to-interference-plus-noise power density ratio at frequency f, and $C[\cdot]$ denotes the *canonical* factor as defined in [6].

Applying the results of [6, (16)] or [6,(43)] to the case of 2 antennas ($M = 2$) and 1 interferer ($L = 1$), we obtain

$$\Gamma(f) = \frac{a+b}{N_o(|I_1(f)|^2+|I_2(f)|^2+N_o)} \quad , \qquad (2)$$

where

$$a = N_o(|H_1(f)|^2+|H_2(f)|^2) \quad , \qquad (3)$$

$$b = |H_1(f)I_2(f)-H_2(f)I_1(f)|^2 \quad , \qquad (4)$$

$H_i(f)$ is the transfer function of the desired channel on the ith antenna, $I_i(f)$ is the transfer function of the interference channel on the ith antenna, and N_o is the two-sided noise density. Without loss of generality, we assume here that the system has a small excess bandwidth such that all transfer functions approximate a Nyquist system.

We see from (2) that, at high SNR, i.e., as $N_o \to 0$,

$$\Gamma(f) \approx 1+\Gamma(f) \to \frac{|H_1(f)I_2(f)-H_2(f)I_1(f)|^2}{N_o(|I_1(f)|^2+|I_2(f)|^2+N_o)} \qquad (5)$$

Therefore,

$$\mathbf{W}_{MLSE} \approx \mathbf{W}_{DFE} \quad . \qquad (6)$$

(In our experience, the only special case where $\mathbf{W}_{MLSE}$ cannot be approximated by $\mathbf{W}_{DFE}$ at high SNR is the case where $L = 1$ and $M = 1$, which is not relevant here, and assuming no excess bandwidth in this case,

$$\Gamma(f) = \frac{|H(f)|^2}{|I(f)|^2+N_o} \text{ and } 1+\Gamma(f) = \frac{|H(f)|^2+|I(f)|^2+N_o}{|I(f)|^2+N_o} .)$$

We have shown above that, for an ideal receiver with infinite complexity, the front-end filter of the MLSE behaves similarly to the feedforward filter of the DFE at high SNR. This proof, however, does not extend to the finite-length case, and certainly not to the case of DDFSE. In these practical cases, we need to consider the tradeoff between maximizing the output SINR of the front-end filter and the ability to suppress both the ISI and CCI with a short postfiltering equalizer. This has been studied before in other publications (e.g., [7]), but no optimum solution has been found. In our case, we have no guarantee that our training method is optimum, and we have no benchmark for the finite-length case with which to compare the performance of our receiver (since there is no optimum solution available to date). However, we are confident that, based on our training method, the receiver at least performs better than a receiver that uses DFE for both training and data detection. This is because the Viterbi equalizer part of the DDFSE always provides more accurate estimates of the received symbols than a hard slicer. Moreover, since there is only a small gap (1-2 dB) in the ideal performance (with infinite length) between the DFE and MLSE (see, e.g., [6, Fig. 5]), and this is also generally the case for practical receivers, we deduce that our training method is close to optimum (not to mention that it is simple and guaranteed to converge, since it is based on MMSE).

III. SIMULATION MODEL

The simulation assumptions are similar to those used in [1] for the EDGE system as described above. Exceptions include the following: First, the number of payload symbols in each burst was reduced to 80 (for the convenience of simulation only). Second, the number of training symbols was variable to study the effects of a longer training sequence. Third, a preamble instead of midamble training sequence was used (this results in more pessimistic performance in fast fading).

The channel models are the typical urban (TU) and hilly terrain (HT) - with $\tau_{\max} = 17.2\mu s$ - models, as shown in Figure 2, with Doppler frequency f_D up to 200 Hz, corresponding to 108 km/h at 2 GHz. The equalizer uses an 8-state Viterbi algorithm with 5 prefilter and 4 feedback taps. Therefore, F=5 and B=5 in our case. We consider a single dominant interferer with random symbol alignment. All the imperfections associated with the above timing and weight estimations/training are included in the simulations.

IV. PERFORMANCE RESULTS

The measure of performance is the BLER, where each block is interleaved over 4 TDMA frames. Overall, we found the proposed 8-PSK scheme performs about 5 dB poorer than standard QPSK with Nyquist filtering. The use of linearized GMSK pulse shaping results in only a small degradation (within 1 dB) compared to square-root raised cosine filtering.

Figure 3 shows the SIR performance for the TU profile at a Doppler frequency of 4 Hz. Three groups of results are given. "1-branch" represents our version of the original EDGE results. "2-branch MMSE" represents the results of our interference suppression technique. "2-branch selection" assumes selection diversity based on the total received signal power. The performance differences between these results and the "2-branch MMSE" results indicate the interference suppression gain.

Each group of curves includes results for two different numbers of training symbols: 26 symbols (the EDGE standard) and 30 symbols, as well as "perfect training", which assumes perfect knowledge of all the channel impulse responses for the desired and interfering signals.

Although "2-branch selection" provides some improvement (about 5 dB) over "1-branch", "2-branch MMSE" provides about 20 dB improvement. The results for 30 training symbols and perfect training show that additional gain (up to 8 dB) could be provided by better weight estimation. Compared to the case without diversity, the BLER with MMSE diversity is more sensitive to the number of training symbols.

This effect is further illustrated in Figure 4, which shows results for MMSE diversity as a function of the number of training symbols. The training sequence used in each case was chosen to have zero autocorrelation sidelobes - this is considered optimum for space-time equalization with unknown interference. For every 4 symbols added, the results show an improvement by 2 to 3 dB in the required SIR.

We note that our equalizer uses 5 feedforward taps (F=5) on each branch and four feedback taps, along with the one symbol Viterbi equalizer (B = 5). Thus, the weight training algorithm estimates 15 weights ($2F+B$) using the training symbols.

From [8], using the Direct Matrix Algorithm for weight estimation (note that we are using RLS, though, which should give similar results), one would expect a 3 dB degradation in signal-to-noise ratio (SNR) as compared to perfect estimation when the number of training symbols is twice the number of weights. Here, we see about a 7 dB degradation in SIR with 30 symbols, which is in line with [8].

With perfect training, the performance should improve with the number of feedforward taps. However, with a fixed training sequence length, this is not always the case, since the weight estimation error increases with the number of feedforward taps. This is illustrated in Figure 5, where we show that even with 30 training symbols a 9-tap feedforward filter does substantially worse than a 5-tap feedforward filter. This is to be expected since, with 9 feedforward taps, 23 weights must be calculated using only 30 symbols, which leads to such a large weight estimation error that the gain due to diversity is lost (see Figure 3). Thus, the above results indicate the potential for even better performance with further improvements in the weight tracking algorithm.

Figure 6 shows the results for the HT profile. Here, the performance with MMSE diversity is even more sensitive to the number of training symbols - increasing the number of training symbols from 26 to 30 provides a 4 dB improvement in the required SIR at a 20% BLER.

Note that the results for "1-branch" and "2-branch selection" in Figure 6 exhibit error floors. This indicates that there are not nearly enough equalizer taps in the receiver when the HT profile is assumed. The results for "2-branch MMSE", however, do not have these floors. This is because MMSE diversity is effective against both cochannel and intersymbol interference.

The results given in Figures 7 and 8 and Table 1 assume the use of 30 training symbols. Figure 7 shows the SNR performance with and without MMSE diversity when interference is not present. The SNR at a 10% BLER is decreased by 10 dB for the TU model with MMSE diversity, and even more for the HT model.

Figure 8 shows the performance (with MMSE diversity) as a function of the Doppler frequency. The SIR and SNR values were chosen to provide BLER's between 10% and 20% at a low fading rate. Due to the use of coding and interleaving, the receiver is seen to be quite robust against fast fading.

Table 1 lists the values of the required SNR and SIR to achieve BLER's around 10% to 20% when the SNR and SIR are comparable. With interference suppression, the system can operate at an SIR of 10 dB and an SNR of 18 dB; whereas the original EDGE system would require an SNR and SIR of 20 dB. Thus, we conclude that in environments where the noise and interference levels are comparable MMSE diversity provides a 10 dB improvement in SIR performance and a 2 dB improvement in SNR performance.

V. CONCLUSIONS

In this paper we presented a receiver structure and algorithms for joint interference suppression and equalization with 2 antennas in EDGE. This receiver provided from 10 to 20 dB interference suppression and from 2 to 10 dB improvement in SNR performance (with and without interference, respectively). We also showed improved robustness against delay spread and the effect of increased training sequence length.

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Required SIR and SNR

1-Branch ($f_D = 4$ Hz)

Channel	SIR	SNR	BLER
Flat Fading	20	20	11.2%
TU Profile	20	20	18.6%
HT Profile	20	20	26.2%

2-Branch ($f_D = 4$ Hz)

Channel	SIR	SNR	BLER
Flat Fading	10	18	9.7%
TU Profile	10	18	11.8%
HT Profile	10	18	19.4%

TABLE 1

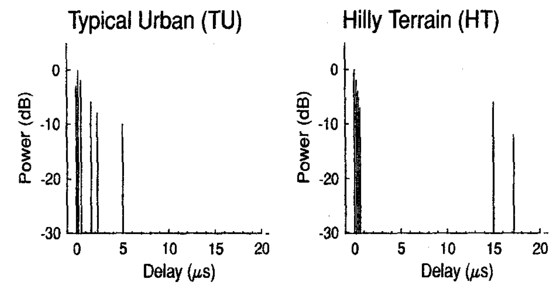

FIGURE 2
Delay profiles for the typical urban and hilly terrain models.

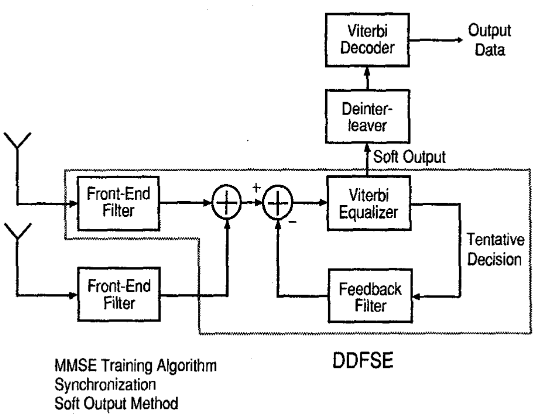

FIGURE 1
Receiver structure.

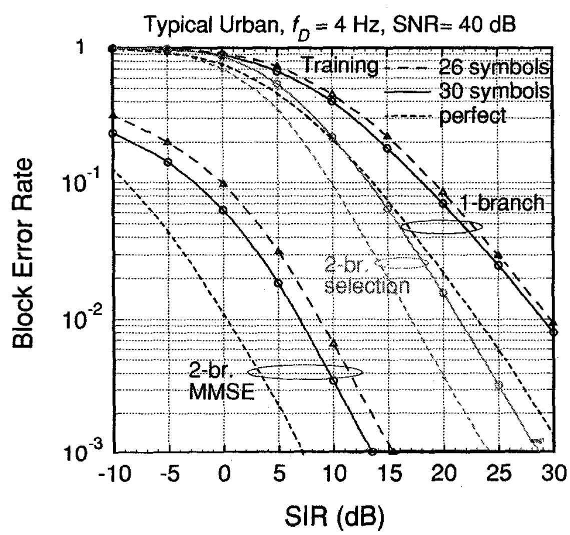

FIGURE 3
The SIR performance for the typical urban profile at a Doppler frequency of 4 Hz.

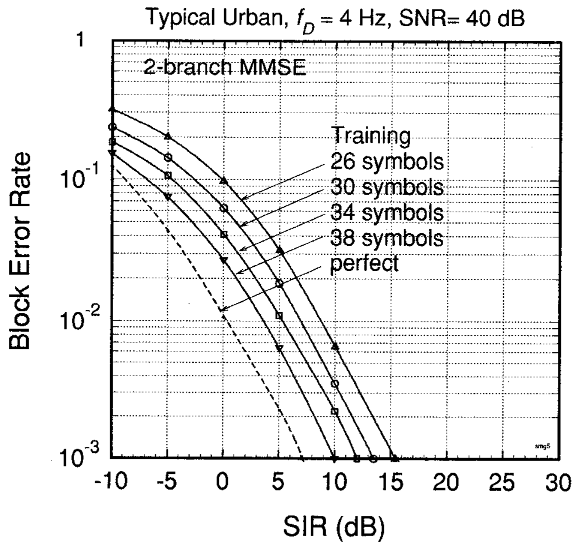

FIGURE 4
The SIR performance for MMSE diversity as a function of the number of training symbols.

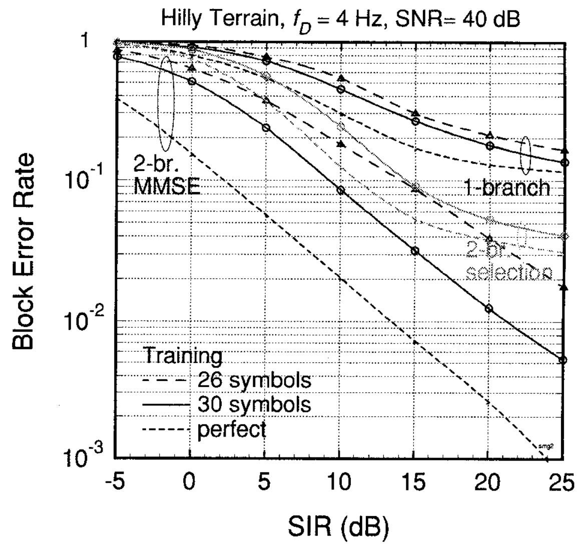

FIGURE 6
The SIR performance for the hilly terrain profile at a Doppler frequency of 4 Hz.

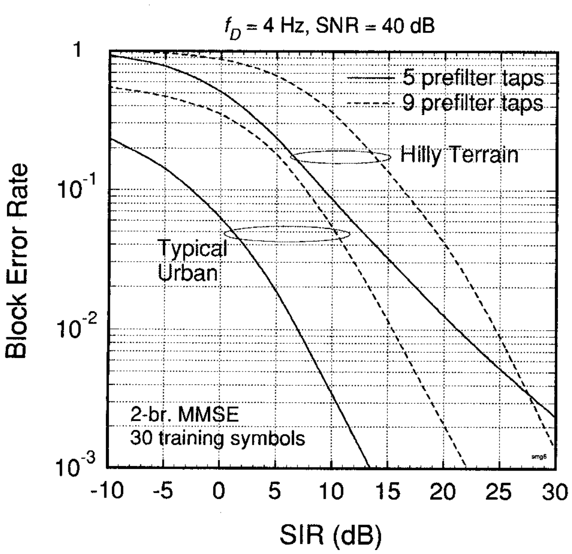

FIGURE 5
Comparison of results using 5 and 9 prefilter taps on each diversity branch.

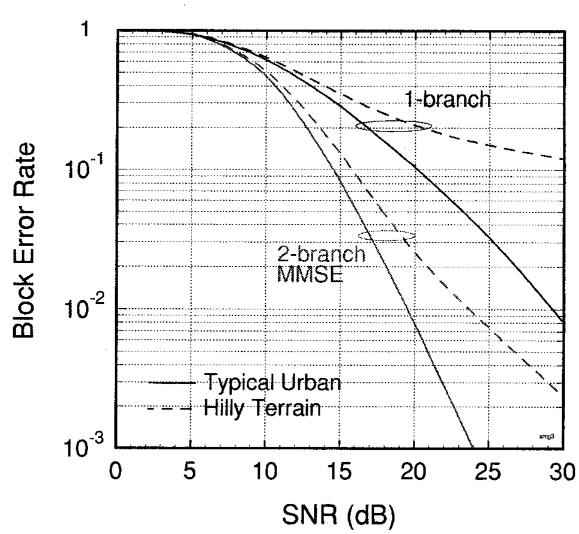

FIGURE 7
The SNR performance with and without MMSE diversity.

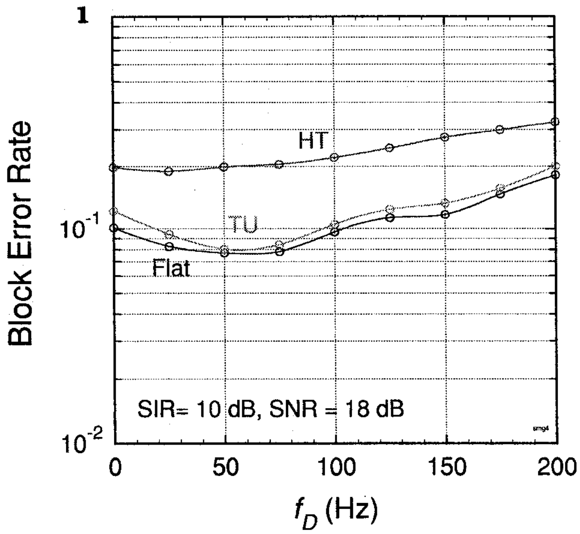

FIGURE 8
The performance with MMSE diversity as a function of the Doppler frequency.