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Combined diversity sequence estimation receiver for wideband digital mobile radio

W.H. Lam, K.C. Chang and C.M. Lo

An investigation is presented into the performance of a novel type of sequence estimation equalisation receiver. The receiver incorporates space diversity to further enhance the signal reception in multipath fading environments. It is shown that the receiver complexity is reduced by ~ 90% with virtually no performance loss compared to that of the MLSE.

Introduction: Efforts have been made to increase the data rate over radio links to enable the provision of advanced wideband mobile radio services. Maximum likelihood sequence estimation (MLSE) implemented by the Viterbi algorithm (VA) is an optimum detection technique for digital signalling over ISI-corrupted channels. Adaptive MLSE can thus be obtained when a channel sounder is used for the periodical acquisition of the time varying channel impulse response. In this Letter, a novel type of combined space diversity and sequence estimation equalisation receiver is proposed. The proposed equaliser, which is equipped with a least sum of squared-error (LSSE) channel sounder [1], employs a combined version of two low complexity algorithms, namely, the decision feedback sequence estimation (DFSE) algorithm [2] and the T-algorithm [3]. In addition, maximal-ratio combining (MRC) and equal-gain combining (EGC) schemes [4] have been investigated.

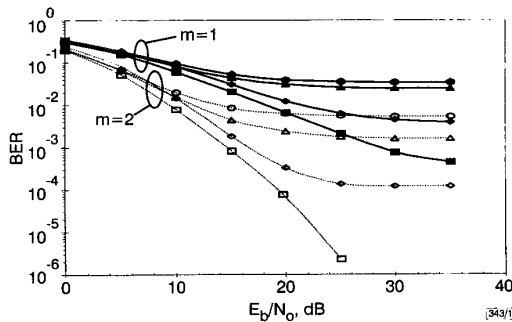


Fig. 1 BER performance of DFSE(m,k,t) and EDFSE(m,k,t) receiver over HT50 channel at three different bit rates

- 271 kbit/s DFSE (1,0,1)
- ▲— 500 kbit/s DFSE (1,0,1)
- ◆— 500 kbit/s EDFSE (1,0,1)
- 1 Mbit/s DFSE (1,0,1)
- 271 kbit/s DFSE (2,0,1)
- △— 500 kbit/s DFSE (2,0,1)
- ◇— 500 kbit/s EDFSE (2,0,1)
- 1 Mbit/s DFSE (2,0,1)

Table 1: Excess delays, mean delays, delay spreads and normalised delay spreads of two different GSM specified radio channels

Profile type	Typical excess delay τ	Mean excess delay $\bar{\tau}$	rms excess delay τ_{RMS}	Normalised delay spread τ_n at various bit rates		
				271 kbit/s	500 kbit/s	1 Mbit/s
BU	μs	μs	μs	0.686	1.266	2.531
HT	20	3.936	6.677	1.808	3.338	6.677

Transmission system: In our study, the radio channel was assumed to be wide-sense stationary and we adopted the GSM specified mobile radio channel COST 207 models [5], including the badly urban (BU) and hilly terrain (HT) models. As shown in Table 1, the root-mean-square (RMS) delay spread τ_{RMS} , the mean excess delay $\bar{\tau}$ and the normalised delay spread τ_n with respect to the bit duration T are used for characterising a time varying channel. Without loss of generality, we adopted the GSM data signalling format but with different transmission bit rate. Each TDMA timeslot consists of $L = 148$ binary data bits α arranged as

$$\alpha = \{\alpha_i\}_{i=0, \dots, L-1} = (\alpha_{L-1}, \alpha_{L-2}, \dots, \alpha_1, \alpha_0)^T$$

$$\alpha_i \in \{\pm 1\} \quad (1)$$

where the superscript T signifies the transpose operator. The binary data sequence α is 0.3GMSK modulated and has a

modulation index h_f of 0.5. Notice that LSSE channel sounding could provide an estimated channel impulse response of up to 10 bit periods, i.e. $L'_c = 10$.

Combined diversity sequence estimation: The maximum likelihood (ML) detector is used to estimate an ML sequence α from every possible transmitted sequence $\tilde{\alpha}_i = (\tilde{\alpha}_i, \tilde{\alpha}_{i-1}, \dots, \tilde{\alpha}_0)^T$. Each of the possible transmitted sequences is assigned a metric to indicate the degree of likelihood to the transmitted sequence α . The VA utilises a trellis state diagram to implement the MLSE (VA-MLSE). At any time instant, the path metric corresponding to state S_i can be written in recursive form as

$$\Gamma_i(S_i) = \Gamma_{i-1}(S_{i-1}) + \gamma_i(S_{i-1} \xrightarrow{\tilde{\alpha}_i} S_i) \quad (2)$$

where $\Gamma_i(\cdot)$ is the accumulated path metric for state S_i , $\gamma_i(\cdot)$ is the branch transition metric, a function of the state transition from S_{i-1} to S_i given the input of $\tilde{\alpha}_i$, defined by

$$\gamma_i(S_{i-1} \xrightarrow{\tilde{\alpha}_i} S_i) = |r_i - \tilde{r}_i(S_{i-1}, \tilde{\alpha}_i)|^2 \quad (3)$$

where r_i is the baseband received signal, $\tilde{r}_i = \hat{\mathbf{h}}^T \tilde{\mathbf{s}}_i$ is the noiseless received signal, $\hat{\mathbf{h}}$ is the estimated channel impulse response of \mathbf{h} and $\tilde{\mathbf{s}}_i = (\tilde{s}_i, \tilde{s}_{i-1}, \dots, \tilde{s}_{i-L_c})^T$ is the transmitted signal vector corresponding to the hypothetical transmitting sequence. When two or more transmitted sequences converge to the same state S_i , only the survivor having the best metric value (i.e. of the smallest value) is retained. As the number of trellis states increases exponentially with the total length of the system memory L_T , the implementation is thus prohibitive for large L_T . The complexity of the VA-MLSE can be reduced by adopting a smaller trellis having shorter memory to represent the same transmission system. The DFSE achieves this goal by truncating the Viterbi equaliser's memory L_{VD} and incorporating a DFE-like feedback mechanism into the structure of the Viterbi equaliser. The trellis state of a DFSE can be obtained by removing k of the state variables from the ML state as follows:

$$\begin{cases} \text{ML state } S_i : (\tilde{\alpha}_i, \tilde{\alpha}_{i-1}, \dots, \tilde{\alpha}_{i-L_T+1}, \theta_i) & L_{VD} = L_T \\ \text{DFSE state } S_i^k : (\tilde{\alpha}_i, \tilde{\alpha}_{i-1}, \dots, \tilde{\alpha}_{i-L_T+1-k}, \theta_i) & L_{VD} = L_T - k \end{cases} \quad (4)$$

where S_i^k denotes the DFSE state having k state variables removed and θ_i is the phase state due to the continuous phase GMSK modulation. The branch metric corresponding to the DFSE state transition

$$S_{i-1}^k \xrightarrow{\tilde{\alpha}_i} S_i^k$$

is given by

$$\gamma_i(S_{i-1}^k \xrightarrow{\tilde{\alpha}_i} S_i^k) = |r_i - \hat{\mathbf{h}}^T \tilde{\mathbf{s}}_i(S_{i-1}^k, \tilde{\alpha}_i)|^2 \quad (5)$$

For the T-algorithm receiver, it is basically the same as the VA-MLSE with the insertion of a rejection test after each trellis update. The rule will reject a path if its path metric Γ_i exceeds the best-valued metric Γ_B by a certain threshold value t_0 . To cope with the dynamic nature of the channel, a time-varying threshold parameter t_0 was proposed [6] and given by $t_0 = (\Gamma_W - \Gamma_B) \times t$, where Γ_W is the worst-valued (largest-valued) metric and t is a weighting parameter in the range (0,1]. Without loss of generality, the DFSE and the T-algorithm can therefore be combined and specified as a class of generalised DFSE receivers denoted as DFSE(k,t). Furthermore, the proposed receiver which incorporates a space diversity branch into the Viterbi equaliser is also investigated. The notation DFSE(m, k, t) signifies the m th order space diversity DFSE(k,t) receiver and the prefix 'E' is added, EDFSE(m,k,t), to denote the deployment of the enhanced channel estimate of $L'_c = 10$ instead of $L'_c = 5$. When MRC space diversity is employed, the branch transition metric γ_i^{MRC} is given by

$$\gamma_i^{MRC}(S_{i-1} \xrightarrow{\tilde{\alpha}_i} S_i) = \sum_{m=1}^{L_m} E_{\hat{\mathbf{h}}_m} \gamma_{m,i}(S_{i-1} \xrightarrow{\tilde{\alpha}_i} S_i) \quad (6)$$

where L_m is the number of diversity branches, $\gamma_{m,i}$ is the branch metric γ_i at the m th antenna and $E_{\hat{\mathbf{h}}_m}$ is the estimated channel energy. When EGC space diversity is employed, estimation of the

channel energy is not required and the resultant state transition metric γ_i^{FGC} is simply equal to the sum of all the transition metrics in the space diversity branches, i.e.

$$\gamma_i^{FGC} \left(S_{i-1} \xrightarrow{\hat{\alpha}_i} S_i \right) = \sum_{m=1}^{L_m} \gamma_{m,i} \left(S_{i-1} \xrightarrow{\hat{\alpha}_i} S_i \right) \quad (7)$$

Table 2: E_b/N_0 at BER = 10^{-2} and average number of survivors for DFSE(m,k,t) receivers over BU50 channel at transmission bit rate of 500kbit/s

k	m = 1					m = 2				
	t = 0.2	t = 0.4	t = 0.6	t = 0.8	t = 1.0	t = 0.2	t = 0.4	t = 0.6	t = 0.8	t = 1.0
E_b/N_0 in dB at BER = 10^{-2} for DFSE(m,k,t)										
0	14.5	12.0	11.8	11.8	11.8	11.0	7.2	7.0	7.0	7.0
1	15.2	12.0	11.8	11.8	11.8	11.6	7.5	7.0	7.0	7.0
Average number of survivors for DFSE(m,k,t)										
0	20	43	121	238	512	19	32	97	230	512
1	11	22	62	98	206	10	17	54	98	206

Results and conclusion: The BU50 and the HT50 GSM-specified channel models with a maximum Doppler shift frequency of 42Hz were employed for the transmission bit rates R , ranging from 271kbit/s to 1Mbit/s. The severity of the resultant time dispersion at three different bit rates is specified by the normalised delay spread τ_n as shown in Table 1. The memory for all the Viterbi equalisers remains at $L_{vp} = 7$ in the study. Fig. 1 depicts the error performance of the DFSE(m,k,t) and EDFSE(m,k,t) receivers for transmission over the HT50 channel. As the performances of the MRC and FGC combining scheme were almost the same, only one set of results for second order combining is presented. Observe that the BER increases with increasing transmission bit rate. Despite the second-order space diversity gain of 8.5dB at a BER of 10^{-2} and $R = 271$ kbit/s, irreducible BER floors appear at higher transmission bit rates due to the relative increase in excess delay. For instance, the BER floors for the DFSE(1,0,1) are equal to 3×10^{-2} and 4×10^{-2} for transmission bit rates of 500kbit/s (i.e. $\tau_n = 3.3$) and 1Mbit/s (i.e. $\tau_n = 6.7$), respectively. When $R = 500$ kbit/s, the deployment of EDFSE(1,0,1) reduces the BER floor to 4×10^{-3} . The use of second-order space diversity reception further reduces the BER floor by an order of magnitude over that of the DFSE(1,0,1) receiver. When the bit rate is further increased to 1Mbit/s, the introduction of space diversity reduces the error floor from 4×10^{-2} to 5×10^{-3} for the DFSE($m,0,1$) equaliser. As shown in Table 2, the receiver complexity can be reduced by using the DFSE algorithm and the T-algorithm specified by the parameters k and t , respectively. Observe that the DFSE(2,1,0.2) requires 11.6dB in E_b/N_0 to achieve a BER of 10^{-2} compared to that of 7.0dB for the DFSE(2,0,1). However, the complexity of the receiver is reduced by $\sim 90\%$ compared to that of the MLSE receiver with virtually no performance degradation when $k = 1$ and $t = 0.6$.

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Design of coded FPM with lower modulation index for fading channels

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The design of coded frequency and phase modulation (FPM) with lower modulation index for fading channels is discussed. An asymmetric constellation is introduced to improve its performance. The proposed trellis coded modulation scheme is particularly advantageous in Rician fading channels with less severe fading conditions.

Introduction: It is well known that the design criterion for trellis coded modulation (TCM) systems over additive white gaussian noise (AWGN) channels is the maximisation of the Euclidean distance of the shortest error event path d_{\min}^2 . However, in a Rayleigh fading channel, the criteria for optimum code design are the length L_{\min} of the shortest error event path and the product P of branch distance along that path [1]. For Rician fading channels where stronger direct signals exist, the performance is affected by all the three quantities, i.e. d_{\min}^2 , L_{\min} and P . Therefore, in mobile communications environments, the optimum trellis codes are those which are optimised with respect to these three quantities.

Padovani and Wolf [2] integrated TCM with combined frequency and phase modulation (FPM) and showed that TCM/FPM can achieve additional gains over conventional TCM/MPSK modulation. Moreover, Periyalwar and Fleisher [3] showed that MTCM/FPM with a comparatively higher modulation index ($h \geq 0.5$) can be optimised with respect to all three quantities and is optimum for both AWGN channels and fading channels. However, for the case of lower modulation index, MTCM/FPM does not perform well.

In [4], it is shown that coded FPM can be optimised by introducing asymmetric constellations and the optimisation is more efficient for lower modulation index. In this Letter we aim to investigate the design of coded FPM with small modulation index over fading channels. An asymmetric constellation is exploited to improve the performance of TCM/FPM over fading channels. It will be shown that coded TCM/FPM as well as MTCM/FPM with an asymmetric constellation can be optimised to satisfy the design criteria for all the three quantities.

Coded FPM with asymmetric constellation for fading channels: As in [4], the asymmetric constellation can be composed of two signal subsets with rotation angle α . For example, the combined two frequency and four phase constellation is plotted in Fig. 1, where ϕ represents the twisted angle of two frequency planes. The resulting constellation is incorporated into the trellis code in accordance with a mapping rule.

The design of TCM/MPSK and MTCM/MPSK schemes for fading channels has been studied in depth [5]. Optimum TCM codes for fading channels have been exhaustively investigated. Here, these TCM optimum codes are employed for TC-FPM. Similarly, the set partitioning methods [2, 3] developed for MTCM/MPSK over fading channels are also employed for MTCM/FPM:

$$E = \begin{bmatrix} 0 & 0 \\ 1 & 5 \\ 2 & 2 \\ 3 & 7 \\ 4 & 4 \\ 5 & 1 \\ 6 & 6 \\ 7 & 3 \end{bmatrix} \quad F = \begin{bmatrix} 0 & 4 \\ 1 & 1 \\ 2 & 6 \\ 3 & 3 \\ 4 & 0 \\ 5 & 5 \\ 6 & 2 \\ 7 & 7 \end{bmatrix} \quad G = \begin{bmatrix} 0 & 2 \\ 1 & 7 \\ 2 & 4 \\ 3 & 1 \\ 4 & 6 \\ 5 & 3 \\ 6 & 0 \\ 7 & 5 \end{bmatrix} \quad H = \begin{bmatrix} 0 & 6 \\ 1 & 3 \\ 2 & 0 \\ 3 & 5 \\ 4 & 1 \\ 5 & 7 \\ 6 & 4 \\ 7 & 1 \end{bmatrix}$$

The resulting TCM/FPM and MTCM/FPM schemes are both optimised by maximising the values of P and d_{\min}^2 with respect to α and ϕ .