A Transmission/Equalisation Procedure for Mobile Digital Radio Links using Interpolated Channel Estimates[1]

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Abstract: A transmission/equalisation strategy is proposed for frequency-selective, fast time-varying noisy digital channels. Each information data sequence is surrounded by a pair of training sequences of known symbols; these are employed by the receiver to estimate the channel impulse response (CIR) at the beginning and at the end of the timeslot, via a suitable Kalman-like filter with fast convergence properties. The CIR behavior along the data sequence is then linearly interpolated and exploited step-by-step by a Maximum Likelihood Sequence Equaliser (MLSE). A simple transmitter/ receiver structure is then obtained and its performance analysed via computer simulations.

1. Introduction

Performance of high-speed digital radio links is limited by the well-known multipath phenomenon which originates intersymbol interference and fading, generally with time-varying characteristics [4,Sect.1.4]. In radio-mobile applications the channel variability is due to the mobile speed; it can be measured by the coherence interval t_c (the time interval when the channel does not vary significantly) or, equivalently, by the Doppler spread $B_D = 1/t_c$ induced over the received spectrum. Typical values of the product B_DT_S between the Doppler spread and the symbol interval T_S are of the order of 10^{-3} or larger for fast-fading channels, while B_DT_S smaller than 10^{-4} denotes a quasi-static channel.

Nonlinear equalisation is mandatory to obtain satisfactory performance for fast time-varying channels. In particular, the well-known MLSE exhibits a reduced computational complexity, due to the Viterbi Algorithm (VA), and for quasi-static channels gives nearly-optimal performance because the decision over a data symbol is carried out on the basis of its preceeding and following symbols [3]. It operates with a decision-delay equal to D symbol intervals, generally chosen equal to five or six times the CIR length L [1]. Different versions of the MLSE have been proposed in the literature, and its effectiveness is so high that it constitutes a typical solution to equalise frequency-selective channels (for example, it is commonly implemented in current GSM cellular receivers).

If the channel is time-varying the MLSE must be assisted by a channel estimator feeding it with the actual estimated CIR [1]. It can be implemented via a gradient algorithm or, more effectively, via a Kalman filter, and typically operates on the basis of a training sequence (constituted by known symbols) during the set-up procedure, then updating the channel estimate on the basis of the decided data. This implies that the channel estimate is delayed of D symbols with respect to the true channel trajectory, with a consequent performance loss for the adaptive MLSE [6]. Alternative adaptive MLSE structures, which employ short-delay "tentative" decisions and large-delay "final" decisions, only mitigate the above problem. Another solution is based on the so-called Per-Survivor Processing principle [6], where a channel estimate is retained for any of the possible "surviving" paths of the VA trellis. Good performance is obtained in this way, but its complexity seems very large due to the simultaneous use of many channel estimators.

[1] Work supported by TELITAL SpA, Sgonico (Trieste), Italy

0-7803-5106-1/98/$10.00 © 1998 IEEE

In this paper a new transmission/equalisation strategy is described which is based on the MLSE solution but overcomes the above drawbacks by employing a suitable channel estimation procedure, thus achieving the high performance of the MLSE at a low computational cost. It is based on the circumstance that, even for very large values of $B_D T_S$ (at most, 10^{-2}) the channel variation is very slow with respect to the symbol rate. The fundamental idea is then to measure periodically the CIR on the basis of training sequences constituted by known symbols, and then to reconstruct the entire CIR behavior along the data sequences through an interpolation algorithm. This is accomplished by maintaining at the same time a large *transmission throughput* (the ratio between the number of information symbols and the overall number of transmitted symbols, including the training sequences) and a small *processing delay* (the interval between the arrival of a data symbol and the output decision about it), which are the fundamental system parameters. The idea of using interpolated channel estimates is not completely new but, at the best of the authors' knowledge, it was employed in the past by following very different equalisation strategies (see, e.g., [2]).

2. The Channel Model and the Proposed Transmission /Equalisation Strategy

Let us model the received sequence via the discrete-time (T_S-sampled) baseband (complex) channel model [1,Sect.6.3]

$$r(n) = \sum_{i=0}^{L-1} g_i(n)\, a(n-i) + w(n),$$

where a(n) is the transmitted (complex) data sequence (e.g., BPSK or QAM); w(n) is a complex additive white noise sequence, with independent components sharing a common variance equal to $N_o/2$, representing thermal noise and external interferences [4,Sect.1.4]; the L (generally, correlated) fading processes $g_0(n)$, ..., $g_{L-1}(n)$ constitute the discrete-time T_S-sampled CIR at time nT_s (including the multipath-faded radio channel and the transmitting and receiving filters) and are zero-mean (complex) Gaussian discrete-time stationary processes with independent real and imaginary components (Rayleigh fading).

Three different framing structures are considered, as described in Fig.1. In the *continuous data* (CD) transmission mode the alternance of a *training sequence* (TS) of L_{train} (known) symbols and a *data sequence* (DS) of (unknown) L_{data} symbols is transmitted. The TS acts as a *preamble* of length L_{pre}=L_{train} for the DS which follows and as a *postamble* of length L_{post}=L_{train} for the preceeding DS. A DS and its two surrounding TSs constitute a *timeslot*. In the *independent timeslots* (IT) transmission mode each timeslot is constituted by preamble, data and postamble of length L_{pre}, L_{data} and L_{post} respectively, and is transmitted separately and independently from the other timeslots. In the *independent blocks* (IB) transmission mode a *block* is constituted by N_m data sequences of length L_{data} (with N_m >1) separated by *midambles* of length L_{mid}, with a preamble at the beginning and a postamble at the end of the frame. The resulting net data throughputs are ρ=$L_{data}/(L_{train}+L_{data})$, $L_{data}/(L_{pre}+L_{data}+L_{post})$, L_{data} / ($L_{pre}+N_m$ xL_{data}+ (N_m-1)xL_{mid}+L_{post}) for the CD, IT and IB transmission modes respectively.

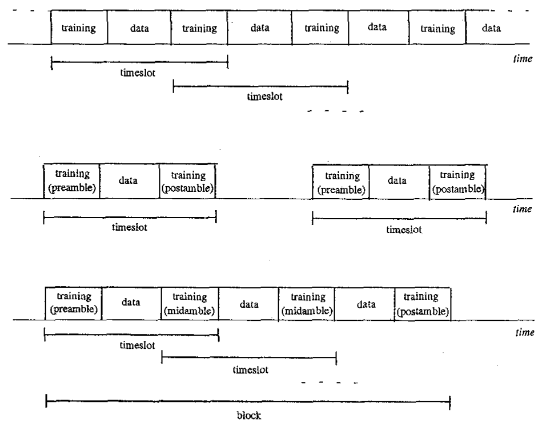

Fig.1 - Continuous data (a), independent timeslots (b) and independent blocks (c) transmission modes.

2.1 - Channel estimation in correspondence of the training sequences

The CIR is estimated in correspondence of the TS by employing a rapidly converging algorithm, so that L_{train} can be kept small and, as a consequence, the throughput is large. For this purpose a suitable choice is constituted by a fast Kalman-like estimator, for which many versions are available from the literature.

The channel estimation procedure requires a minimum length L_{train} of the training sequences in order to guarantee that the convergence is reached. In order to keep L_{train} small (so that the throughput is maximised) but at the same time ensure the convergence, it is possible to carry out the estimate N_{iter} times (with N_{iter}>1) using the same timeslot, starting at every iteration with the result obtained at the end of the previous iteration. Computer simulations (see Sect.3) showed that in general two or, at most, three iterations are sufficient to achieve convergence and improve the estimation accuracy, even for large $B_D T_s$.

2.2 - Channel interpolation along the data sequence

After receiving a timeslot (a preamble) the CIR is estimated as described in Sect.2.1. The subsequent data sequence is then ignored and the CIR is estimated again form the next timeslot (a postamble). The CIR behavior along the data sequence in between the two timeslots is then reconstructed via a simple linear interpolation exploiting the two above CIR estimates, as graphically illustrated in Fig.2. Employing the CD or IB transmission mode the CIR estimate at the postamble coincides with the preamble estimate for the next data block, which does not need to be re-calculated.

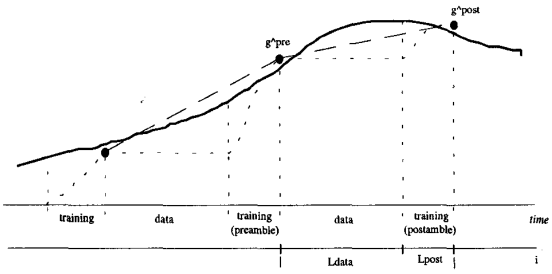

Fig.2 - Example behavior of the real (or imaginary) component of a CIR tap (bold solid line), trajectory of the estimates in correspondence of the timeslots (dot line) and linear interpolation of the CIR along the information symbols (dashed line). g^pre and g^post denote the preamble and postamble estimates, respectively.

If L_{data} is small (compared to the channel coherence interval t_c) the above linear interpolation results in a good piecewise linear channel approximation along the data symbols, and a small processing delay is also obtained. However, small values of L_{data} result in a small throughput, so that a trade-off value must be properly selected for L_{data}.

We remark that no error propagation effects or crash phenomena due to decision errors are present because *the decided data are not used for channel estimation purposes*. Moreover, the linear interpolation is useful to combat a deep fade eventually occurring during the data sequence. It also reduces the computational burded with respect to alternative solutions which continuously track the channel, also along the data sequences.

2.3 - The data detection procedure

Data detection is accomplished via a suitable version of the classic MLSE-VA employing the step-by-step CIR estimate obtained as in Sect.2.2. In particular, the metrics pertaining to the trellis nodes corresponding to the i-th time instant are computed employing the CIR estimate at the same instant. The VA assumes a certain channel length L_{VA} and starts from a known initial state (the last L_{VA}-1 symbols of the preamble), also terminating over a known state (the first L_{VA} -1 symbols of the postamble). The decision is taken in a whole for all the L_{data} information symbols, although for large L_{data} a VA with fixed decision-delay could be more appropriate to reduce the computational complexity, which is discussed in [1].

Regarding the computational complexity of the channel estimator, it is essentially that of the classic Kalman filter [1,Sect.6.8]. The overall computational burden per received symbol is indeed smaller, because the channel estimator is active only in correspondence of the training sequences. The channel interpolation along the data sequences exhibits a negligible computational cost. We can conclude that the proposed strategy can be then easily implemented via commercially available components.

3. Performance Analysis

The effectiveness of the equalisation strategy described in Sect.2 has been evaluated via extensive computer simulations in terms of Bit-Error-Rate (BER) vs Signal-to-Noise Ratio (SNR) curves by considering many different Rayleigh fading channels. In the simulations at least ten independent channel realisations of 64k samples each have been generated via a standard FFT/IFFT filtering procedure. The results in Figs.3-6 concern the case of Land-Mobile Doppler spectrum, typical of the terrestrial radio-mobile environment [4]. In Figs.3,4 the reference cases of two and three T_S-spaced equipowered uncorrelated taps are considered, while in Figs.5,6 the Hilly Terrain (HT) GSM test channel [5,Sect.IV] is taken into account. Larger values of L_{pre} and L_{VA} have been employed to equalise the HT channel, because its CIR length is about L=6; in particular, for the CD mode (Fig.5) L_{pre}=15 gives better results that L_{pre}=9, while L_{pre}=18 was necessary for the IT mode (Fig.6) because in this case the channel estimator is initialised to zero for each new timeslot.

From the above results we observe that, depending on the assumed system constraints, the processing delay can be kept small by assuming L_{data} not large (e.g., L_{data} = 21), while at the same time a small L_{pre} (e.g., L_{pre} = 9) can be employed due to the fast convergence properties of the channel estimation procedure along the timeslots, thus ensuring a large net throughput.

4. Conclusions

The proposed transmission/equalisation strategy is indeed very simple, but it performs very well and overcomes other previous solutions. In particular, in [8] the performance of the adaptive MLSE assisted by a Kalman channel estimator has been largely investigated by considering different decision delays and compared to the symbol-by-symbol MAP equaliser of [7], where the adaptive MLSE with tentative decisions has been also analysed. From the comparison with the above-reported techniques and with the other results available from the literature we conclude that the proposed solution gives better performance under the same system constraints (in particular, the same throughput and a reduced processing delay).

We finally remark that the basic differences between the proposed strategy and that proposed in [2], which also employs interpolated channel estimates, are that in [2] a block least-squares adaptive channel estimator is considered, the interpolation is carried out from many channel estimates via an (approximate) truncated Nyquist pulse, the equaliser is a DFE and only the CD transmission mode is considered.

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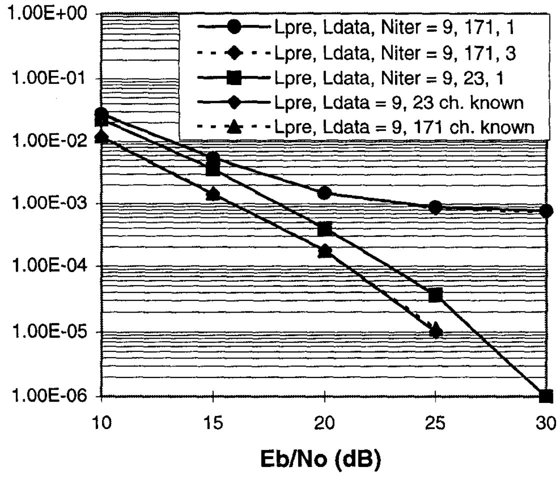

Fig.3 - BER performance of the proposed equalisation strategy for the three-tap reference channel with Land Mobile Doppler spectrum (T_s-spaced equipowered Rayleigh-distributed paths, $B_D T_s = 10^{-3}$). The transmission mode is CD, the modulation is QPSK; no channel coding or differential modulation is considered. The channel estimator performs N_{iter} iterations for each training sequence and the MLSE-VA assumes a channel length L_{VA}=3. For reference purposes the performance of the "ideal" VA with step-by-step known channel is also given.

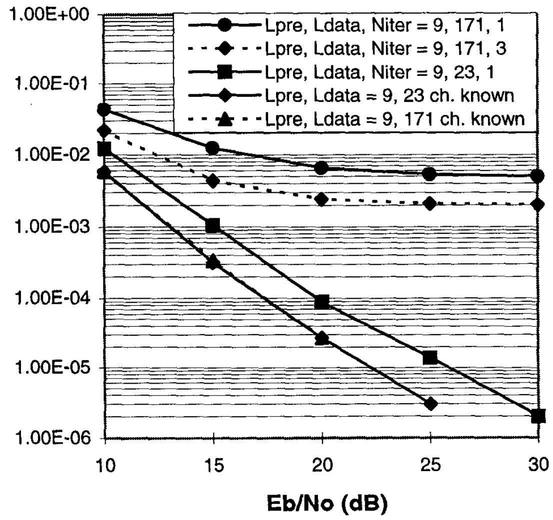

Fig.4- -Same as in Fig.3, but for a two-tap channel. The VA assumes $L_{VA} = 2$.

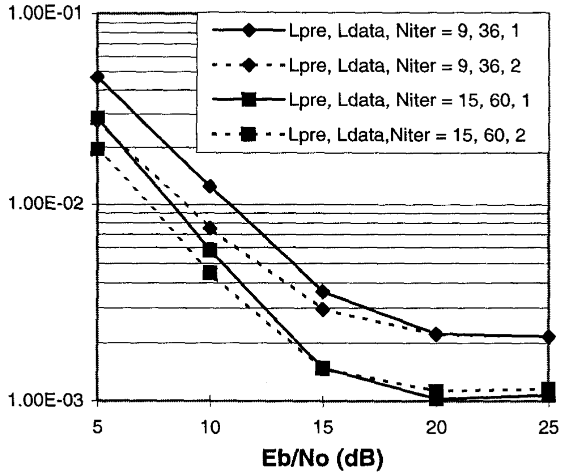

Fig.5 - BER performance of the proposed equalisation strategy for the Hilly terrain (HT) test channel [5]. The transmission mode is CD, the modulation is BPSK at 270 kbit/sec, the Doppler spread is B_D = 100 Hz (corresponding to a mobile speed of 120 km/h for a carrier frequency of 900 Mhz). The VA assumes $L_{VA} = 6$.

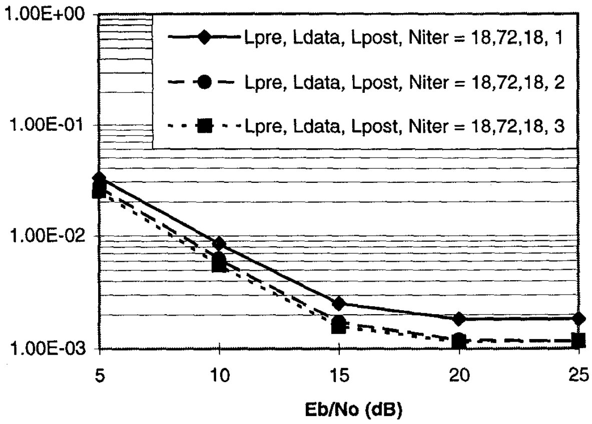

Fig.6 - Same as in Fig.5, but with IT transmission mode.