A New Receiver for Digital Mobile Radio Channels with Large Multipath Delay[1]

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ABSTRACT

A new receiver structure is proposed for the equalisation of data links affected by large multipath delays. It is constituted by a symbol-by-symbol MAP detector which exploits the "sparse" nature of the channel, constituted by a few nonzero powerful taps spaced by many zero taps of negligible amplitude, to achieve a complexity tied to the number of nonzero taps and not to the overall channel length. When the channel is time-varying, an efficient nonlinear Kalman-like channel estimator is employed to track only the nonzero taps.

1. INTRODUCTION

Intersymbol interference (ISI) from multipaths constitutes one of the main factors limiting the performance of terrestrial radio mobile systems [7]. The temporal spread induced by the transmission medium can be as large as $t_s = 20$ μsec and, at the GSM bit rate of 270.8 Kbps, this gives a length L=6 for the equivalent discrete-time channel impulse response (CIR). Although the Viterbi Equaliser (VE) constitutes a feasible solution commonly implemented in commercial GSM receivers [7], a complexity reduction is of interest in order to save battery power or to implement some additional features. Referring to future third-generation radio-mobile systems, large symbol rates (up to 2 Mbit/sec) are envisaged and large values of L are expected, so that a direct application of the VE may result quite complex. Low-cost solutions (for example, a transversal filter or a DFE) generally exhibit a performance loss with respect to the VE. It can be observed that, even when the temporal spread t_s is large, only a few echoes are generally present. As a consequence only a few taps in the equivalent CIR sampled at symbol interval T_s are *nonzero*, i.e. their magnitude is much larger than that of the remaining coefficients (which are called *zero* taps) and the channel is classified as *sparse*.

A possible solution is constituted by an (adaptive) transversal filter which reduces the ISI associated to the farest echoes, followed by a VE [2]: however, echo cancellation is only partial and the filter also induces some noise enhancement.

A "pruned" version of the VE has been recently proposed [5], where the presence of zero CIR coefficients is taken into account in the calculation of the metrics associated to the VE trellis paths. However, the VE must be designed "ad hoc" for the actual configuration of the zero and non-zero CIR coefficients: such design seems not easy and no general procedure has been given so far.

In the present paper the "sparse" nature of the channel is exploited to derive a novel efficient equaliser (called SC-MAP equaliser) with a complexity proportional to the number of nonzero CIR coefficients. It belongs to the family of the Symbol-by-Symbol Maximum A Posteriori (MAP) equalisers, originally proposed in [1] and recently revisited and made more attractive from a computational point of view (see, in particular, [4],[8]).

2. THE MAP EQUALISER FOR NON-SPARSE CHANNELS

Following the classic baseband discrete-time (sampled at symbol interval T_s) equivalent channel model, we express the received sequence as:

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$$r(n) = \sum_{k=0}^{L-1} c_k\, b(n-k) + w(n), \qquad (1)$$

where b(n) is the transmitted data sequence, belonging to a (complex) constellation of size S; L is the CIR length; w(n) is an additive white noise sequence (typically, Gaussian) with zero mean and spectral density power $N_0/2$; $c_0,..., c_{L-1}$ are the CIR coefficients.

Let us assume for the moment that the channel is Non-Sparse (NS), i.e. all the L CIR coefficients are (generally) nonzero, and that it is also Time- Invariant (TI), i.e. the CIR does not significantly change during the transmission interval.

The MAP equaliser [4],[8], is based on the definition of the channel state vector $\underline{b}(n)\equiv[b(n) \ldots b(n-L+1)]^T$, which can assume $N = S^L$ different configurations $\underline{m}_1, ..., \underline{m}_N$, and on the introduction of the A Posteriori Probability (APP) vector:

$$\underline{p}(n) \equiv [P(\underline{b}(n) = \underline{m}_1 \mid \underline{r}_{1,n}) \ldots P(\underline{b}(n) = \underline{m}_N \mid \underline{r}_{1,n})]^T, \qquad (2)$$

which collects the probabilities of the N possible channel states at step n, conditional to the observation of the vector $\underline{r}_{1,n} \equiv [\ r(1) \ldots r(n)\ ]$ of the sequence received from step 1 to step n.

The matrix F of the channel state transition probabilities (TPs), with elements:

$$F_{i,j} = P(\underline{b}(n) = \underline{m}_i \mid \underline{b}(n-1) = \underline{m}_j), \quad i,j = 1,...,N, \qquad (3)$$

is also introduced and employed to calculate as $\underline{p}(n/n-1) = F\underline{p}(n-1)$ the one-step prediction of $\underline{p}(n)$, i.e. its estimate based on the observation of $\underline{r}_{1,n-1}$. At step n the MAP equaliser recursively computes the vector $\underline{p}(n)$ on the basis of its prediction $\underline{p}(n/n-1)$, of the actual observation r(n) and of the available CIR estimate. From $\underline{p}(n)$ the probabilities of the constellation symbols at step n-D (i.e., with a decision-delay D = L-1 equal to the channel memory) are then calculated and finally, following the MAP rule, it is decided for the symbol with maximum probability (see [8] for details).

When the channel is NS but also Time-Variant (TV), during the transmission interval the CIR coefficients vary with the temporal index n from symbol to symbol, and the model (1) is still valid after replacing the CIR taps $c_0,..., c_{L-1}$ with their time-varying counterparts $c_0(n), ..., c_{L-1}(n)$. In this case a Channel Estimator (CE) is employed to feed step-by-step the MAP equaliser with an updated CIR estimate, obtained on the basis of the received sequence r(n) and of the MAP outputs (or of the known data, in correspondence of the training sequence). In [8] an adaptive nonlinear Kalman-like filter has been proposed, driven by the channel APP vector $\underline{p}(n)$ available (with zero delay) from the MAP equaliser, which constitutes a "soft" statistic about the data sequence, and not by the "hard" decisions. This mitigates the error propagation phenomenon and results in good performance [8].

3. The SC-MAP Equaliser for Sparse Channels

When the channel is TI and Sparse (S), let $c_{i'1},..., c_{i'k}, ..., c_{i'NNZ}$ be the NNZ nonzero CIR coefficients, their positions in the CIR being determined by the set of integer indices i'1, i'2,...,i'k,..., i'NNZ-1, i'NNZ (with i'1 < i'2 < ...< i'k < ... < i'NNZ-1 < i'NNZ). The number and the positions of the CIR coefficients are assumed known or reliably estimated. In this case the model (1) becomes:

$$r(n) = \sum_{k=1}^{NNZ} c_{i'k}\, b(n-i'k) + w(n), \qquad (4)$$

and a vector $\underline{b}'(n) \equiv [\ b(n-i'1) \ldots b(n-i'NNZ)\ ]$ of the channel state "visible" at step n can be defined. It is constituted by the NNZ transmitted symbols which are located in correspondence of the nonzero CIR coefficients. The vector $\underline{b}'(n)$ can take on $N' = S^{NNZ}$ different configurations $\underline{m}'_1, ..., \underline{m}'_{N'}$, and their probabilities at step n are collected in the APP vector of the visible channel state:

$$\underline{p}'(n) \equiv [\, P(\underline{b}'(n) = \underline{m}'_1 \mid r_{1,n}) \ldots P(\underline{b}'(n) = \underline{m}'_{N'} \mid r_{1,n})]. \quad (5)$$

The transmitted symbols located in correspondence of the remaining L-NNZ zero CIR coefficients constitute the "hidden" channel state at step n. Among them, let us collect in the vector $\underline{b}''(n) \equiv [\, b(n-i''1) \ldots b(n-i''NZN)\,]$ the NZN symbols such that a nonzero CIR coefficient is located in correspondence of the subsequent position in the CIR; such symbols become visible at the next step n+1, and then constitute a "near" hidden (NH) state. Let us also collect in the vector $\underline{b}'''(n) \equiv [\, b(n-i'''1) \ldots b(n-i'''NZF)\,]$ the remaining NZF = L-NNZ-NZN symbols, which are still hidden even at the next step n+1 and then constitute a "far" hidden (FH) state. The above definitions are clarified in the example of Fig.1.

The vectors $\underline{b}''(n)$ and $\underline{b}'''(n)$ may take on respectively the $N'' = S^{NZN}$ configurations $\underline{m}''_1, \ldots, \underline{m}''_{N''}$ and the $N''' = S^{NZF}$ configurations $\underline{m}'''_1, \ldots, \underline{m}'''_{N'''}$; they have associated at step n the APP vectors $\underline{p}''(n)$ e $\underline{p}'''(n)$, defined similarly to (5) with the obvious substitutions (we observe that the configuration assumed by $\underline{b}'(n)$ at step n+1 depends not only on $\underline{b}'(n)$ itself, but also on $\underline{b}''(n)$).

The matrices F'_k, k=1,...,N'' of the TPs conditional on $\underline{b}''(n)$, with elements:

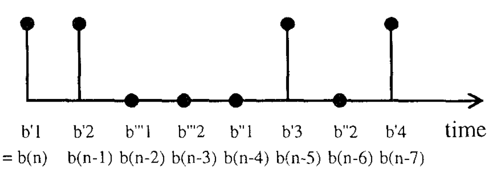

Fig.1 - Example of sparse channel of length L=8 with impulse response [1, 1, 0, 0, 0, 1, 0, 1] and positions of the NNZ=4 nonzero symbols b'(n) = [b'1 b'2 b'3 b'4], of the NZN=2 near hidden symbols b''(n) = [b''1 b''2] and of the NZF=2 far hidden symbols b'''(n) = [b'''1 b'''2].

$$F'_{k\,i,j} = P(\underline{b}'(n) = \underline{m}'_i \mid \underline{b}'(n-1) = \underline{m}'_j,\ \underline{b}''(n) = \underline{m}''_k),$$
$$i,j=1,\ldots,N',\ k=1,\ldots,N'', \quad (6)$$

are then introduced to express the one-step prediction $\underline{p}'(n/n-1)$ of $\underline{p}'(n)$ as:

$$\underline{p}'(n/n-1) =$$
$$= \{ \sum_{k=1}^{N''} [\, F'_k\, P(\underline{b}''(n) = \underline{m}''_k)\,]\,\}\ \underline{p}'(n)$$
$$= F(n)\, \underline{p}'(n), \quad (7)$$

where the matrix $F(n)$ constitutes the average of the matrices F'_k weighted by their probabilities at step n, and its elements are the unconditional TPs of $\underline{p}'(n)$ at step n:

$$F_{i,j}(n) = P(\underline{p}'(n) = \underline{m}'_i \mid \underline{p}'(n-1) = \underline{m}'_j),\ \ i,j = 1,\ldots,N'. \quad (8)$$

At step n, the vectors $\underline{p}''(n)$ and $\underline{p}'''(n)$ are computed on the basis of the APP vectors of the visible state calculated and stored in the previous steps (in principle, all the previous vectors $\underline{p}'(n-L+1), \ldots, \underline{p}'(n-1)$ are necessary for this purpose).

When the channel is S and TV, let us assume that the positions of the NNZ nonzero CIR coefficients do not vary during the transmission interval. In this case the channel model (4) is still valid after replacing the NNZ nonzero CIR coefficients $c_{i'1}, \ldots, c_{i'NNZ}$ with their time-varying counterparts $c_{i'1}(n), \ldots, c_{i'NNZ}(n)$.

We employ an Adaptive Nonlinear Kalman-like (ANKL) filter similar to that in [8] to track the trajectories of the NNZ nonzero CIR coefficients only. It is based on r(n) and on the APP vector $\underline{p}'(n)$ of the observable state $\underline{b}'(n)$, in place of the APP vector $\underline{p}(n)$ of the whole channel state $\underline{b}(n)$, as it is done in [8]. In correspondence of the training sequence a data-aided version of the ANKL filter is employed, taking into account that the channel state is known.

4. Simulation Results and the Timeslot-Adaptive Equalisation Strategy

The effectiveness of the proposed SC-MAP equaliser has been verified via computer simulations. We describe here only the case of Wide Sense Stationary Uncorrelated Scattering (WSSUS) Rayleigh TV channels with Land Mobile (LM) fading spectrum, with a given value of the parameter B_dT_s (the product between T_s and the Doppler spread B_d).

For every trial a large number of independent timeslots have been generated, each constituted by a preamble of p known symbols followed by d data symbols (framing (p,d)). For each timeslot, L CIR taps are preliminarly estimated from the preamble; the NNZ most powerful ones are then identified (with the reasonable assumption that their positions do not change during the timeslot) and all the others are neglected, i.e. they are treated as zero. The channel index sets are thus identified and then employed to compute the TP matrices F'_k, $k = 1,...,N''$ of (6).

The parameter NNZ is fixed a priori on the basis of the computing resources available at the receiver. The preliminary estimate of the L CIR taps is carried out via the simple cross-correlation method commonly employed in GSM receivers (SC-CC solution), or by simultaneously tracking the L CIR taps via the data-aided ANKL filter (SC-KF solution). For reference purposes, in our trials we have also considered the ideal case when the NNZ taps are adaptively chosen on the basis of the true CIR taps at the end of the preamble (SC-ID solution), and the case when they are non-adaptively a priori chosen from the assigned channel power-delay profile (SC-NA solution).

The NNZ selected CIR taps are re-estimated from the training sequence via the data-aided ANKL filter. After initialising the channel APP vectors from the training sequence, the symbol-by-symbol SC-MAP equaliser (with the ANKL channel estimator) is then applyed to process and decode the received data symbols.

Some results are reported in Figs.2,3,4 for the GSM Hilly Terrain (HT) test channel with LM fading spectrum and symbol rate 270.8 Kbps [3]. This constitutes an example of sparse channel with L=8, its power-delay profile being reported in Fig.5.

For reference purposes the performance of the adaptive MAP equaliser of Sect.2 operating over the first four or five CIR taps are also reported in Figs.3,4. Bad performance is obviously obtained because in both cases the receiver completely neglects the contribution of tap #6, which is the most powerful after tap #2 (see Fig.5).

We also verified that the SC-MAP with NNZ=2 (Fig.2) largely outperforms the MAP with five taps (Fig.4). Obviously, the MAP equaliser with 8 taps (also reported in Figs.3,4) is better than the SC-MAP with NNZ < 8.

As an example of more "sparse" channel, in Fig.6 the equivalent T_s-sampled power/delay profile of the HT channel is reported for a transmission rate of 500 Kbps, and the corresponding performance curves are shown in Fig.7. The effectiveness of the SC-MAP equaliser is then fully verified.

From the computer simulations we then verified that the timeslot-adaptive SC-CC and SC-KA solutions are quite close to the ideal SC-ID and largely outperform the non-adaptive SC-NA, so that their introduction is fully motivated.

Regarding the computational complexity of the SC-MAP equaliser, as an illustrative example we report in Tab.I the (relative) computing times measured from our computer simulations for some reference power/delay profiles. It is verified that the complexity of the MAP solution grows exponentially with L, while the overhead of the MAP equaliser is less than a factor 2 for L ranging from 2 to 7.

6. Conclusions

The SC-MAP equaliser for channels with large multipath delays proposed in this paper constitutes a practical solution for digital radio mobile receiver applications. The core of the receiver is the derivation of an efficient MAP detector which takes into account only the significant CIR coefficients, with any spacing between them, at the price of a small extra-load with respect to the MAP detector designed for non-sparse channels (with the same number of nonzero coefficients). This allows a flexible equalisation strategy, where a preliminary estimate of the CIR is carried out to select the most relevant CIR coefficients. In the case of time-varying channels such preliminary estimate is carried out once for every received timeslot and the SC-MAP equaliser is coupled with a suitable Kalman-like channel estimator which tracks only the nonzero CIR coefficients, thus obtaining an adaptive equaliser with reduced-complexity and good performance.

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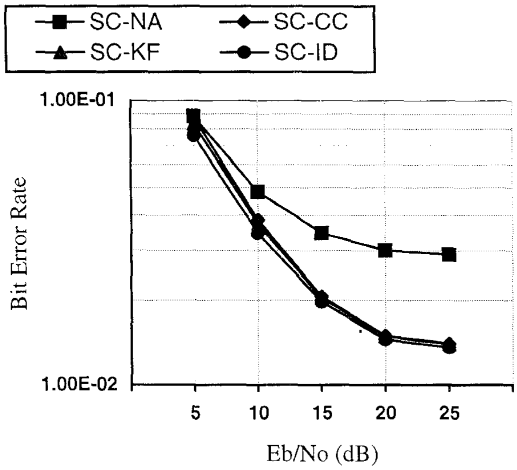

Fig.2 - Performance of the SC-MAP equaliser with NNZ=2 for the HT-GSM test channel of Fig.5 with L=8 taps. The fading spectrum is Land Mobile with $B_dT_s=10^{-4}$, the modulation is BPSK @270.8 Kbps (with no coding), the framing is (26,58).

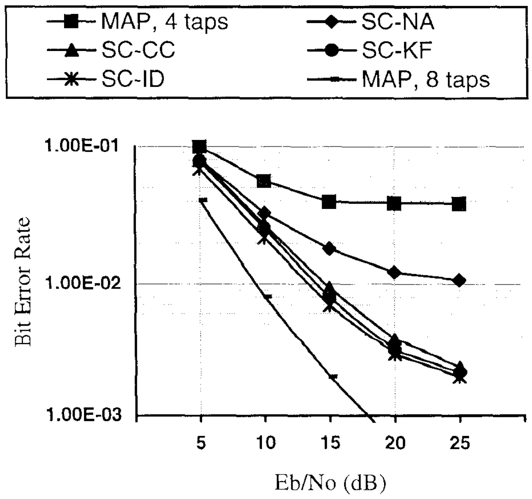

Fig.3 - Same as Fig.2, with NNZ=4.

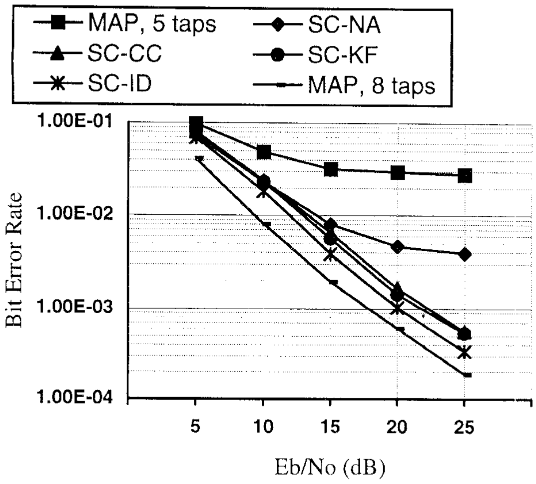

Fig.4 - Same as Fig.2, with NNZ=5.

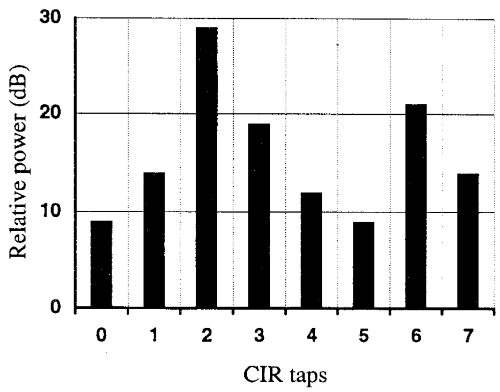

Fig.5 - Power/delay profile of the GSM-HT (Hilly Terrain) test channel [3] at 270.8 Kbps.

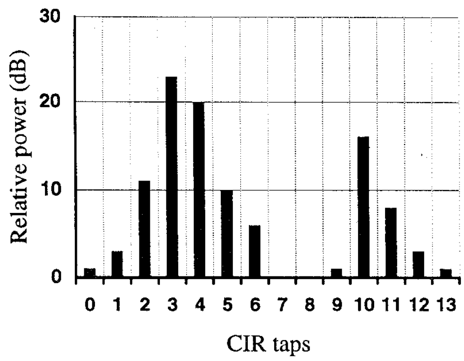

Fig.6 - Power/delay profile of the GSM-HT test channel at the bit rate of 500 Kbps.

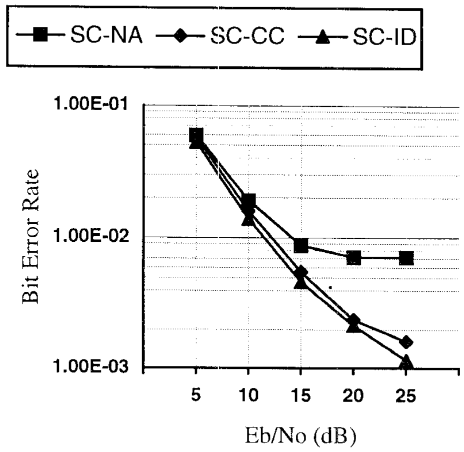

Fig.7 - Performance of the SC-MAP equaliser with NNZ=4 for the test channel of Fig.6. The fading spectrum is Land Mobile with $B_dT_s = 10^{-4}$, the modulation is BPSK @500 Kbps (with no coding), the framing is (52,116).

CIR power/delay profile	MAP equaliser	SC-MAP equaliser (NNZ=2)
[1/2 1/2]	1.0 (L=2)	1.0
[1/2 0 1/2]	2.6 (L=3)	1.1
[1/2 0 0 1/2]	10.6 (L=4)	1.4
[1/2 0 0 0 1/2]	44.2 (L=5)	1.7
[1/2 0 0 0 0 1/2]	198.2 (L=6)	1.8
[1/2 0 0 0 0 0 1/2]	1018.0 (L=7)	1.8

Tab.I - Comparison of the computing times measured for the MAP and SC-MAP equalisers, for a channel with NNZ=2 and L ranging from 2 to 7.