Equalization of Digital Radio Channels With Large Multipath Delay for Cellular Land Mobile Applications

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***Abstract*— A new maximum *a posteriori* (MAP) equalizer is proposed for digital radio links affected by large multipath delays. The "sparse" nature of the channel, where a few *nonzero* powerful taps are spaced by many negligible taps, is exploited to achieve a complexity proportional to the number of nonzero taps. When the channel is time-varying, an efficient nonlinear Kalman-like channel estimator is employed to track only the nonzero taps.**

I. Introduction

INTERSYMBOL interference from multipaths is one of the main factors limiting the performance of terrestrial radio mobile systems, the time spread t_s being as large as 20 μs in open areas [1]. The length of the equivalent discrete-time (sampled at symbol interval T_s) channel-impulse response (CIR) is then about $L = 6$ samples at the GSM bit rate of 270.8 kb/s, and can be larger in future radio-mobile systems, where symbol rates up to 2 Mbit/s are envisaged. A complexity reduction with respect to the classic Viterbi equalizer (VE) is then of interest for practical applications [1], but low-cost solutions (e.g., a transversal filter or a DFE as in [2]) generally exhibit a nonnegligible performance loss with respect to the VE.

However, even when t_s is large, only a few echoes are generally present and a few CIR taps (called *nonzero* taps) exhibit a magnitude much larger than the others (*zero* taps). In [3] a transversal filter is employed before the VE to reject the farest echoes, but the rejection is only partial and the filter induces some noise enhancement. In [4] a "pruned" VE has been proposed, where the zero CIR taps are taken into account in the calculation of the metrics in the VE trellis. The selective use of the channel taps in the VE was also applied in [7].

A novel efficient equalizer for sparse channels (called SC-MAP) is proposed in this paper, belonging to the symbol-by-symbol maximum *a posteriori* (MAP) family of equalizers recently revisited, e.g., in [5], [6]. It simultaneously manages three different channel states ("visible", "near-hidden" and "far-hidden") to take into account the nonzero CIR coefficients, at the price of a small extra-load with respect to the MAP equalizer for nonsparse channels with the same number of nonzero taps. Although the soft-output information from the MAP equalizer can be further processed by a subsequent error control channel decoder, in this paper we consider only the case of uncoded transmission and do not evaluate the soft-output quality of the proposed reduced-state equalizer.

When the channel is time-variant the SC-MAP equalizer is supported by a suitable adaptive nonlinear Kalman-like filter which tracks the trajectories of the nonzero CIR taps only.

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II. The MAP Equalizer for Sparse Channels

The equivalent baseband discrete-time (sampled at symbol interval T_s) received sequence is:

$$(n) = \sum_{k=0}^{L-1} c(k)b(n-k) + w(n) \tag{1}$$

where $b(n)$ is the transmitted data sequence, with (complex) constellation of size S, $c(k)$ is the CIR, and $w(n$ is a zero-mean white noise sequence with spectral power density $N_0/2$. The MAP equalizer for time-invariant and (generally) nonsparse channels [5], [6] starts from the definition of the channel state vector $\underline{b}(n) \equiv [b(n) \cdots b(n-L+1)]^T$, which includes the current symbol $b(n)$ and can assume $N = S^L$ different configurations $\underline{m}\ , \cdots, \underline{m}_N$, and the associate *a posteriori probability* (APP) vector

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

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

Following [6], the matrix F of the channel state transition probabilities (TP's), with elements:

$$F_{i,j} = \Pr(\underline{b}(n) = \underline{m}_i \mid \underline{b}(n-1) = \underline{m}_j), \quad i,j = 1, \cdots, N \tag{3}$$

allows to calculate as $\underline{p}(n \mid n-1) = F\underline{p}(n-1)$ the one-step prediction of $\underline{p}(n)$, i.e., its estimate based on the observation of $\underline{r}_{1,n-1}$. As described in [6], at step n the MAP equalizer recursively computes the vector $\underline{p}(n)$ on the basis of its prediction $\underline{p}(n \mid n-1)$, the actual observation $r(n)$ and the available CIR estimate. From $\underline{p}(n)$ the APP's of the constellation symbols (Symbol APP's) are then calculated with a decision-delay $D = L-1$ equal to the channel memory and finally, following the MAP rule, it is decided for the symbol with maximum probability (see [6] for details).

When the channel is time-invariant and sparse, let $\{c(i_1), \cdots, c(i_R)\}$ be the R nonzero CIR taps, with positions

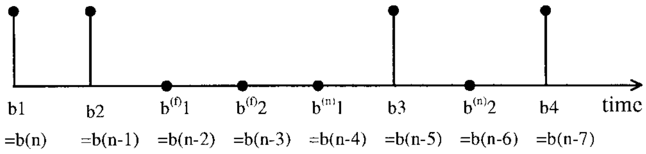

Fig. 1. Example of sparse channel of length $L = 8$ with impulse response $[1, 1, 0, 0, 0, 1, 0, 1]$ and positions of the $R = 4$ nonzero symbols $b(n) = [b_1 \;\; b_2 \;\; b_3 \;\; b_4]$, the $R_n = 2$ near hidden symbols $\underline{b}^{(n)}(n) = [b_1^{(n)} \;\; b_2^{(n)}]$ and the $R_f = 2$ far hidden symbols $b^{(f)}(n) = [b_1^{(f)} \;\; b_2^{(f)}]$.

given by the set of integer indices $i_1, \cdots, i_R$, in ascending order. The model (1) becomes

$$r(n) = \sum_{k=1}^{R} c(i_k) b(n - i_k) + w(n) \tag{4}$$

and the vector $\underline{b}(n) \equiv [b(n - i_1) \cdots b(n - i_R)]^T$ represents now the channel state "visible" at step n, constituted by the transmitted symbols corresponding to the nonzero CIR taps[1]. It can take on $N = S^R$ different configurations $\underline{m}_1, \cdots, \underline{m}_N$, whose probabilities are collected in a channel APP vector $\underline{p}(n)$ defined as in (2), but with the above-introduced definitions of N, $\underline{b}(n)$ and $\underline{m}_1, \cdots, \underline{m}_N$.

The transmitted symbols corresponding to the zero CIR coefficients constitute the "hidden" channel state. The vector $\underline{b}^{(n)}(n) \equiv [b(n - i_1^{(n)}) \cdots b(n - i_{R_n}^{(n)})]^T$ collects the R_n symbols which are not visible at step n but will be visible at step $n + 1$, thus constituting a "near" hidden state. The vector $\underline{b}^{(f)}(n) \equiv [b(n - i_1^{(f)}) \cdots b(n - i_{R_f}^{(f)})]^T$ collects the remaining $R_f = L - R - R_n$ symbols, which are still hidden even at step $n + 1$ and then constitute a "far" hidden state (see Fig. 1). The vectors $\underline{b}^{(n)}(n)$ and $\underline{b}^{(f)}(n)$ may take on respectively the $N_n = S^{R_n}$ configurations $\underline{m}_1^{(n)}, \cdots, \underline{m}_{N_n}^{(n)}$ and the $N_f = S^{R_f}$ configurations $\underline{m}_1^{(f)}, \cdots, \underline{m}_{N_f}^{(f)}$; their APP vectors $\underline{p}^{(n)}(n)$ and $\underline{p}^{(f)}(n)$ are defined again similarly to (2), with obvious substitutions.

The matrices G_k, $k = 1, \cdots, N_n$ of the TP's conditional on $\underline{b}^{(n)}(n)$, with elements

$$\begin{aligned} \{G_k\}_{i,j} &= \Pr\big(\underline{b}(n) = \underline{m}_i \mid \underline{b}(n-1) = \underline{m}_j, \underline{b}^{(n)}(n-1) = \underline{m}_k^{(n)}\big), \\ &\qquad i, j = 1, \cdots, N, \quad k = 1, \cdots, N_n \end{aligned} \tag{5}$$

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

$$\begin{aligned} \underline{p}(n \mid n-1) &= \left\{ \sum_{k=1}^{N_n} \left[G_k \Pr\big(\underline{b}^{(n)}(n-1) = \underline{m}_k^{(n)}\big) \right] \right\} \underline{p}(n-1) \\ &= E(n) \underline{p}(n-1) \end{aligned} \tag{6}$$

where $E(n)$ represents the statistical average of the matrices G_k, its elements being the unconditional TP's of $\underline{p}(n)$: $\{E\}_{i,j}(n) = \Pr(\underline{p}(n) = \underline{m}_i \mid \underline{p}(n-1) = \underline{m}_j)$, $i, j = 1, \cdots, N$. In (6), the term $\Pr(\underline{b}^{(n)}(n-1) = \underline{m}_k^{(n)})$ is directly obtained as an element of the vector $\underline{p}^{(n)}(n-1)$. The vectors $\underline{p}^{(n)}(n)$ and $\underline{p}^{(f)}(n)$ are then calculated from the Symbol APP's stored at the previous steps, when they were computed on the basis of the APP vectors of the visible state.

[1] For the sparse channel, we employ here the same notations $N, \underline{b}(n), \underline{p}(n), \underline{m}_k$ introduced for the nonsparse channel.

TABLE I
POWER/DELAY PROFILES OF THE GSM-HT (HILLY TERRAIN) TEST CHANNEL AT THE GSM BIT RATE OF 270.8 kb/s ($T_s = 3.69$ μsec) AND AT THE BIT RATE OF 500 kb/s ($T_s = 2$ μsec).

Tap position	@270.8 kbps (dB)	@500 kbps (dB)
1	-20	-12
2	-15	0
3	0	-3
4	-10	-13
5	-17	-17
6	-20	-∞
7	-8	-∞
8	-15	-∞
9	-∞	-∞
10	-∞	-7
11	-∞	-15

When the channel is sparse but also time-variant, let us assume that the positions of the nonzero CIR coefficients do not vary during the transmission interval. In this case the model (1) is still valid after replacing $c(i_1), \cdots, c(i_R)$ with their counterparts $c_n(i_1), \cdots, c_n(i_R)$, which vary with the index n. We employ an adaptive nonlinear Kalman-like (ANKL) filter similar to that in [6] to track the R nonzero CIR taps only and feed step-by-step the MAP equalizer with the CIR estimates. The filter is based on $r(n)$ and the APP vector of the visible state $\underline{p}(n)$ available (with zero delay) from the MAP equalizer, while the filter in [6] employs the APP vector of the whole channel state $\underline{b}(n)$. Similarly to [6], 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.

III. Simulation Results and the Timeslot-Adaptive Equalization Strategy

For each simulation trial many independent timeslots have been generated, each constituted by a preamble of p known symbols followed by d data symbols (framing (p, d)). For each received timeslot, L CIR taps are preliminary estimated from the preamble, the R most powerful ones are identified and all the others are treated as zero (R is fixed a priori on the basis of the allowed receiver complexity). Such preliminary estimate 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). We have also considered the ideal reference cases when the R taps are adaptively chosen from the true CIR taps at the end of the preamble (SC-ID solution) and when they are nonadaptively a priori chosen from the channel power-delay profile (SC-NA solution). The channel index sets are then employed to compute the TP matrices G_k, $k = 1, \cdots, N_n$ of (5) and the R selected CIR taps are re-estimated from the training sequence via the data-aided ANKL filter. The SC-MAP equalizer (with the ANKL

TABLE II
COMPARISON OF THE COMPUTING TIMES MEASURED FOR THE MAP AND SC-MAP EQUALISERS, FOR AN EXAMPLE WSSUS CHANNEL WITH $R = 2$ NONZERO INDEPENDENT TAPS OF EQUAL POWER, AND L RANGING FROM 2 TO 7.

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

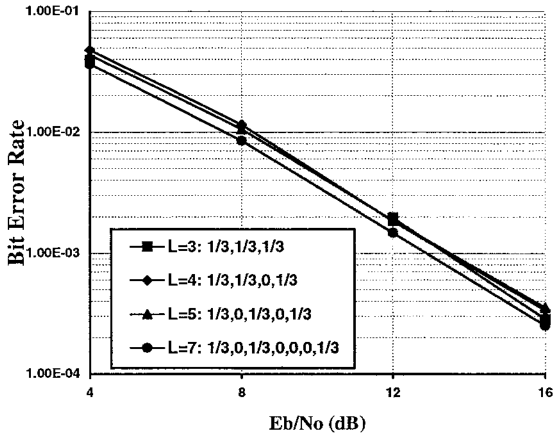

Fig. 2. BER performance of the SC-MAP equalizer for the reference WSSUS channel constituted by $R = 3$ independent Rayleigh-distributed nonzero taps with equal power, at different spacings. The different power/delay profiles are reported in the legenda. The fading spectrum is land-mobile with Doppler spread $B_d = 10^{-4}/T_s$, the modulation is BPSK (with no form of coding), the framing is $(7, 18)$.

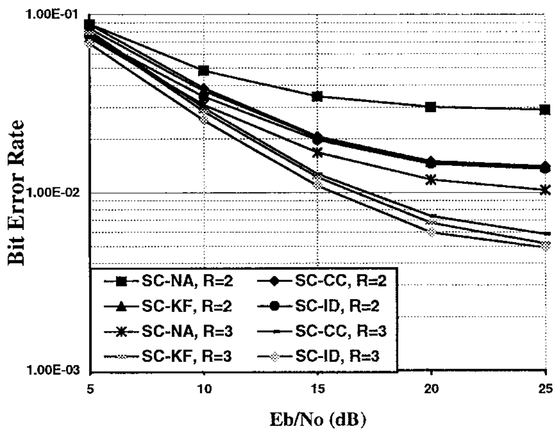

Fig. 3. BER performance of the SC-MAP equalizer (with $R = 2$ and $R = 3$) for the GSM-HT (Hilly Terrain) test channel @270.8 kb/s with $L = 8$ taps, its power/delay profile being given in Table I. The fading spectrum is land-mobile with $B_d = 10^{-4}/T_s$, the modulation is BPSK (with no coding), the framing is $(26, 58)$.

channel estimator) is then applied to process and decode the received data symbols.

Some results are reported in Fig. 2 for a reference channel with $R = 3$ nonzero taps, and in Figs. 3, 4 for the GSM-HT test channel @270.8 kb/s with the power/delay profile reported in Table I. The SC-MAP with $R = 2$ (Fig. 3) largely outperforms the "full" MAP with five taps (Fig. 4), which does

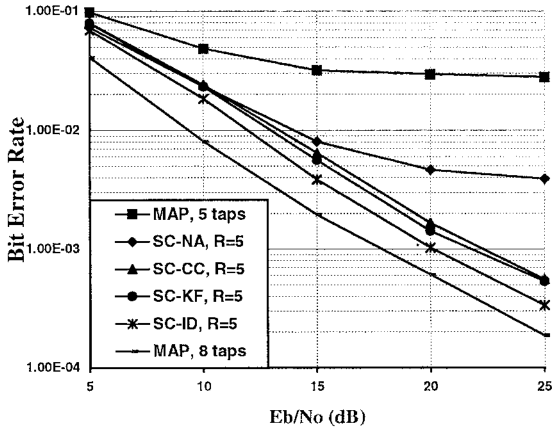

Fig. 4. The same as in Fig. 3, with $R = 5$. For reference purposes the performance of the MAP equalizer operating over the first five and eight CIR taps are also reported.

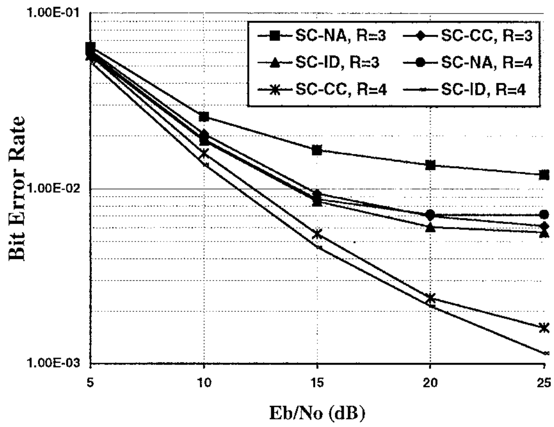

Fig. 5. BER performance of the SC-MAP equalizer (with $R = 3$ and $R = 4$) for the GSM-HT test channel @500 kb/s with $L = 11$ taps, its power/delay profile being given in Table I. The fading spectrum is Land Mobile with $B_d = 10^{-4}/T_s$, the modulation is BPSK (with no coding), the framing is $(52, 116)$.

not consider the presence of taps no. 6, 7, 8. Obviously, the "full" MAP with eight taps is better than the SC-MAP with $R < 8$, but with $R = 5$ the penalty is less than 2.5 dB; by fact, the SC-MAP approaches the full MAP whenever the R tracked CIR taps carry most of the received power.

At 500 kb/s the same GSM-HT channel becomes more "sparse" (see Table I) and the effectiveness of the SC-MAP equalizer is checked from Fig. 5. From our trials we verified that the timeslot-adaptive SC-CC and SC-KA solutions are quite close to the ideal SC-ID and outperform the non-timeslot-adaptive SC-NA, so that their introduction is fully motivated.

IV. COMPUTATIONAL COMPLEXITY AND CONCLUSIONS

We report in Table II the (relative) computing times measured for some reference power/delay profiles, showing that the complexity of the MAP equalizer grows exponentially with L, while the overhead of the SC-MAP equalizer is less than a factor 2, for L ranging from 2 to 7. We can conclude that the proposed SC-MAP equalizer for channels with large multipath delays constitutes a practical solution for digital radio-mobile receivers.

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