MULTISTAGE SYMBOL-BY-SYMBOL BAYESIAN INTERFERENCE CANCELLATION FOR UMTS-CDMA LINKS AFFECTED BY SEVERE MULTIPATH

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Abstract

A simple multiuser detector (MUD) is proposed for UMTS-CDMA links over channels affected by severe multipath. After coherent combining and despreading, for each user a suitable "Bayesian" memoryless non-linearity gives symbol-by-symbol the expected values of the transmitted data. These are employed for soft removal of inter-symbol and multiple-access interference from the received sequences. The procedure is iterated in a multistage structure until final hard-decisions are taken. The proposed Multistage Bayesian MUD exhibits low complexity and better performance (close to the ideal canceller) than other known solutions, even in presence of channel estimation errors. [1]

1. Introduction: The Problem and the System Model

Third-generation radio-mobile communication systems will employ the CDMA radio access technique and FDD/TDD systems are candidates for IMT-2000 [1]. In particular, the TDD system exhibits the spreading factor Q = 16 corresponding (at the chip rate of 3.84 Mcps) to the symbol rate f_s = 240 kbaud. The received signal is then impaired by inter-symbol (ISI) and multiple-access (MAI) interference spanning over a large number of received data samples (up to 6 or 7 samples when the channel time spread is about 20 μsec, as in non-urban open areas). Moreover, in TDD the number K of active (synchronous) users is small (max. K = 8) and traditional single-user receivers designed for the AWGN channel give poor performance.

In such environment Multi-User Detection (MUD) not only is feasible (because K is small) but is also mandatory to satisfy the required quality of service [2]. On the other hand FDD allows a larger spreading factor (up to Q=256) and the number of users is large too; MUD is then considered only for particular environments and constitutes an option for the base station but not for the mobile receiver (due to complexity reasons). Classic MUD solutions are the Zero Forcing (ZF) and the Minimum Mean Square Error [3]. However, MUD is a very time-consuming task even for small K and alternative solutions are currently investigated with the purpose to ameliorate the cost/performance trade-off. In this Paper we assume that the received signal, sampled at chip rate f_c, is preliminary subject to coherent combining (CC) and despreading for each of the K users. The modulation is QPSK and short-length spreading codes with periodicity P=Q are employed, as in TDD (and also in FDD, whenever the MUD option is activated). The following K (complex) sequences are obtained (at symbol rate f_s) after CC:

$$y^{(k)}(n) = \sum_{i=1}^{K} \sum_{m=-L_g}^{L_g} [g^{(i,k)}(m) d^{(i)}(n-m)] + w^{(k)}(n), \; k=1,..K, \quad (1.1)$$

where $\{d^{(i)}(n)\}$ are the K transmitted data sequences; $\{w^{(k)}(n)\}$ are K additive (Gaussian) observation noise sequences obtained after CC and despreading of the thermal noise; $\{g^{(i,k)}(n)\}$ are the baseband discrete-time (sampled at f_s) equivalent channel impulse responses (S-CIRs) between the i-th user and the output of the k-th despreader, having (maximum) length $2L_g+1$. As a consequence of the CC operation, the S-CIRs are non-causal.

2. The Multistage PIC/SIC Receiver

The general scheme of a multistage parallel interference canceller (MPIC) for multipath channels and (generally) complex modulation is obtained by extending the approach proposed in [4] for BPSK transmissions over AWGN channels. Here, a block of N consecutive received samples for each user (after CC and despreading) is collected in the vector $\underline{y} \equiv [y^{(1)}(1),...,y^{(1)}(N),...,y^{(K)}(1),..\,.,y^{(K)}(N)]$. In the first stage KxN "tentative decisions" $\tilde{d}_1^{(k)}(n) = F[y^{(k)}(n)]$ are computed and collected in the vector $\underline{\tilde{d}}_1 = [\tilde{d}_1^{(1)}(1),\dots,\tilde{d}_1^{(1)}(N),\dots,\tilde{d}_1^{(K)}(1),\dots,\tilde{d}_1^{(K)}(N)]$. $F[\ ]$ is an instantaneous (generally,complex) non-linearity which, for BPSK modulation, can be a hard-limiter, a hyperbolic tangent, or others (see Fig.5 of [4]). The total interference $\tilde{I}_1^{(k)}(n)$ due to both ISI

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and MAI affecting user #k at epoch n, with n = 1,…, N and k = 1,…, K, is then estimated from the tentative decisions as

$$\tilde{I}_1^{(k)}(n)=\sum_{i=1,i\neq k}^{K}\Big[\sum_{m=-L_g}^{L_g} g^{(i,k)}(m)\,\tilde{d}_1^{(i)}(n-m)\Big]+\sum_{m=-L_g,m\neq 0}^{L_g} g^{(k,k)}(m)\,\tilde{d}_1^{(k)}(n-m). \qquad (2.1)$$

After posing: $\underline{\tilde{I}}_1=[\tilde{I}_1^{(1)}(1),..,\tilde{I}_1^{(1)}(N),..,\tilde{I}_1^{(K)}(1),..,\tilde{I}_1^{(K)}(N)]$, the difference $\underline{y}$ - $\underline{\tilde{I}}_1$ constitutes the received sequences "cleaned" from the estimated interference. As in [4], at the output of the first stage a "soft-decision" is computed (and collected in the vector $\underline{\tilde{y}}_1$) as the weighted sum between the received sample $\tilde{y}_0^{(k)}(n) = y^{(k)}(n)$ and $\underline{y}$ - $\underline{\tilde{I}}_1$, i.e.

$$\tilde{y}_1^{(k)}(n)= p_1\,[\,y^{(k)}(n) - \tilde{I}_1^{(k)}(n)\,] + (1 - p_1)\,\tilde{y}_0^{(k)}(n), \qquad (2.2)$$

with $0 < p_1 < 1$. The structure of the first stage is repeated in the next stages of the detector but tentative decisions are now computed from the soft-decisions (2.2) and not from the original received sequence. The basic equations for the s-th stage are still (2.1),(2.2) with the subscript 1 replaced by the stage index s = 1, …, N_s and the subscript 0 by s-1. Final decisions are obtained by thresholding the soft-decisions at the output of the last stage.

The above-described scheme can be also implemented in a sequential (MSIC) version: this simply means that whenever a tentative decision is computed at time epoch n for any user, it is immediately employed for ISI&MAI calculation by the remaining users in the same stage and by all users at following epochs. The users are preliminary ordered on the basis of the received power. Computer simulations showed that MSIC offers a slight performance improvement over MPIC or a computational saving (same performance with one less stage) at the price of increased processing delay.

MPIC and MSIC may employ a different non-linearity *F*[ ] at any stage and different weigths p_1, p_2, …, p_{Ns}. In [4] the different stages employ the same non-linearity but growing weighs, so that more and more emphasis is given to the "cleaned" sequence in the last stages. In this case a straightforward solution is obtained for QPSK modulation by considering the following *hyperbolic tangent* non-linearity:

$$\tilde{d} = F[y] = \begin{cases} \tanh(\alpha y_R) & \text{if} \quad abs(y_R) \geq abs(y_I) \\ j\cdot\tanh(\alpha y_I) & \text{if} \quad abs(y_R) < abs(y_I) \end{cases} \qquad (2.3)$$

having posed $\tilde{d} = \tilde{d}_s^{(k)}(n)$, $y = y_R + j\,y_I = y^{(k)}(n)$ and having employed the QPSK constellation symbols {+1,-1,+j, -j}; α is a scale factor. In the following we call MPIC and MSIC the parallel and serial MUDs using (2.3).

An alternative approach is to consider equal weights at different stages and a (complex) memoryless non-linearity *F*[ ] which is smooth in the first stage and gradually harder in the next stages. The theoretically optimum solution for *F*[ ] is obtained by calculating the expected value of the symbol $d^{(k)}(n)$ conditional to the observation of the actual data sample $y^{(k)}(n)$. This can be carried out by assuming that the sum of ISI and MAI constitutes a zero-mean Gaussian noise of variance σ^2_I so that the total variance is $\sigma^2_{WI} = \sigma^2_W + \sigma^2_I$, σ^2_W being the variance of the observation noise $w^{(k)}(n)$. From the classic Bayes'rule its expression is calculated as

$$\tilde{d} = F[y] = E\{d \mid y\} = \frac{\sum_{m=0}^{M-1}(c_m + js_m)\exp\{(y_R c_m + y_I s_m)/\sigma^2_{WI}\}}{\sum_{m=0}^{M-1}\exp\{(y_R c_m + y_I s_m)/\sigma^2_{WI}\}} \qquad (2.4)$$

where $d \equiv d(n)$ and c_m+js_m, m=0,…,M-1, are the constellation symbols (e.g., M=2 in BPSK, M=4 in QPSK). We underline that the Gaussianity of ISI and MAI is assumed only to calculate the non-linearity (2.4) and that (2.4) itself is valid for any complex modulation. A typical behaviour of the non-linearity (2.4) is shown in Fig.1. A large value of σ^2_{WI} is employed in the first stage and decreasing values in the next stages, thus reflecting the circumstance that ISI and MAI are gradually reduced by moving from one stage to the other. The multistage Bayesian (MB) receiver then employs $p_1=\ldots=p_{Ns} = 1$ and (2.4) instead of (2.3). It can be implemented in both parallel and sequential versions, but in the following only the latter case will be considered.

Complex memoryless Bayesian non-linearities similar to (2.4) were calculated in previous works on blind equalisation of single-user ISI channels [9,10]. The non-linearity in (2.4) could be also derived from the symbol-by-symbol MAP algorithm (see, e.g., the recent contribution in [11]) in the limit case of zero-memory flat-fading channel.

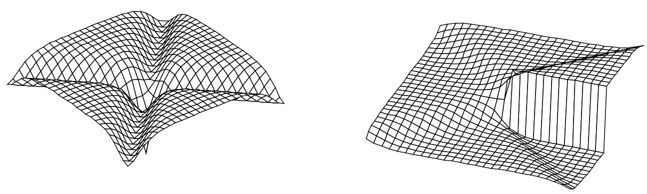

Fig.1 – Magnitude (left) and phase (right) of the non-linearity *F* of (2.4) for QPSK modulation and $\sigma^2_{WI} = 0.15$.

In the context of CDMA multiuser detection, a soft non-linearity is employed in [5] at the output of the coherent combiners while in [6] a multilevel quantiser is considered. Other MUD receivers with soft decisions have been recently proposed, for example, in [7],[8]. The above solutions (and many others found in the literature) assume that ISI is absent because the symbol interval is much larger than the channel time spread or because it has been fully suppressed by the CC, so that the detector is basically designed for the AWGN channel.

3. Simulation Results with Perfect Channel Estimation

The performance of the MUDs described in the previous section has been evaluated via computer trials simulating at chip rate the whole CDMA transmission system model. For every trial a large number of timeslots has been generated, each constituted by N independent symbols. The CIRs are generated at chip rate (C-CIRs) following the classic Wide-Sense Stationary Uncorrelated Scattering (WSSUS) random model with Rayleigh-distributed magnitude, uniform-distributed phase and assigned power-delay profile. Channel realisations are kept constant along each timeslot and are independent from a timeslot to another. While simulating the uplink they are also independent from an user to another, even though the same power-delay profile is assumed for all the users. The CC takes into account all nonzero C-CIR coefficients (known or estimated) from which the S-CIRs are computed and then also employed for ISI and MAI calculation. In Figs.2,3 a TDD system with QPSK modulation, K=8 users, spreading factor Q=16, chip rate of 3.84 Mcps and Orthogonal Walsh spreading codes have been assumed [1], while packet length is N = 40. The following test channel profiles have been considered in the simulations:

Veh.B		Veh.A	
Delay	Power (dB)	Delay	Power (dB)
0	-14.1	0	-16.8
$2T_c$	0.0	$2T_c$	0.0
$4T_c$	-0.6	$4T_c$	-2.8
$6T_c$	-12.3	$6T_c$	-8.6
$8T_c$	-17.1	$8T_c$	-14.5
$38T_c$	-12.7	$10T_c$	-14.9
$55T_c$	-8.8	$12T_c$	-20.0
$83T_c$	-16.4		

For comparison purposes we also report the performance of the ZF receiver [3] and of the ideal interference canceller where ISI and MAI are calculated (after CC) from the error-free hard-decisions. The number N_S of stages has been selected so that increasing it does not substantially improve the receiver performance.

For MPIC and MSIC solutions, optimised performance was obtained at all considered SNRs by assuming N_S = 4 with weights 0.5, 0.7, 0.9, 1 while setting α=3 in (2.3) for the first N_S-1 stages and $\alpha=\infty$ (i.e., hard-limiting) in the last stage.

Regarding the MB, in a similar way we selected optimised values for the parameter σ^2_{WI} in (2.4) and assumed $\sigma^2_{WI}=0$ (i.e., hard limiting) in the last stage. However, in this case a simpler and more “objective” strategy leading to nearly the same results was found by selecting σ^2_{WI} for user #k at step n as the minimum between the squared distances $|y^{(k)}(n) - g^{(k,k)}(0)\,(c_m+js_m)|^2$, with m = 0,…, M-1, which represents a “local” measure of σ^2_{WI}. In this way the receiver does not need to estimate the noise and interference power. Referring to the case of known channel, from Figs.2,3 we verify that the MPIC, MSIC and MB receivers outperform ZF and that MB is better than MSIC&MPIC because (2.3) implies a preliminary hard-decision while (2.4) is completely “soft”; this also explains why the weights $p_1,\ldots,p_{Ns}$ are not useful when using (2.4). In Fig. 4 we also report the performance of the MB receiver for the uplink case with spreading factor Q=128 (as allowed in FDD) for all users. It is verified that in this case the MB solution supports more than 80 users.

4. Channel Estimation for Downlink and Uplink

A realistic performance evaluation must take into account the presence of channel estimation errors, which constitutes a critical point for ICs [2]. For the downlink case of Fig.2 a training sequence of length p=512 chips (as described by ETSI specifications) is inserted at the center of the transmitted timeslot *after* having summed up the K spreaded data sequences. The receiver computes the cross-correlation between such (known) midamble and the received one, thus directly obtaining the channel estimate. From Fig.2 it is verified that the performance loss due to imperfect channel estimation is less than 1 dB.

For the uplink case of Fig.3, each user employs a different training sequence of length p=512 (as described by ETSI specifications) which is inserted at the center of the transmitted timeslot. In this case the single-user cross-correlation technique employed for the downlink is not effective, due to the presence of MAI between midambles. The Maximum Likelihood (ML) channel estimator of [12] gives satisfactory results but it implies a matrix inversion so that the computational burden is quite large.

An alternative technique based on cross-correlation and mul-

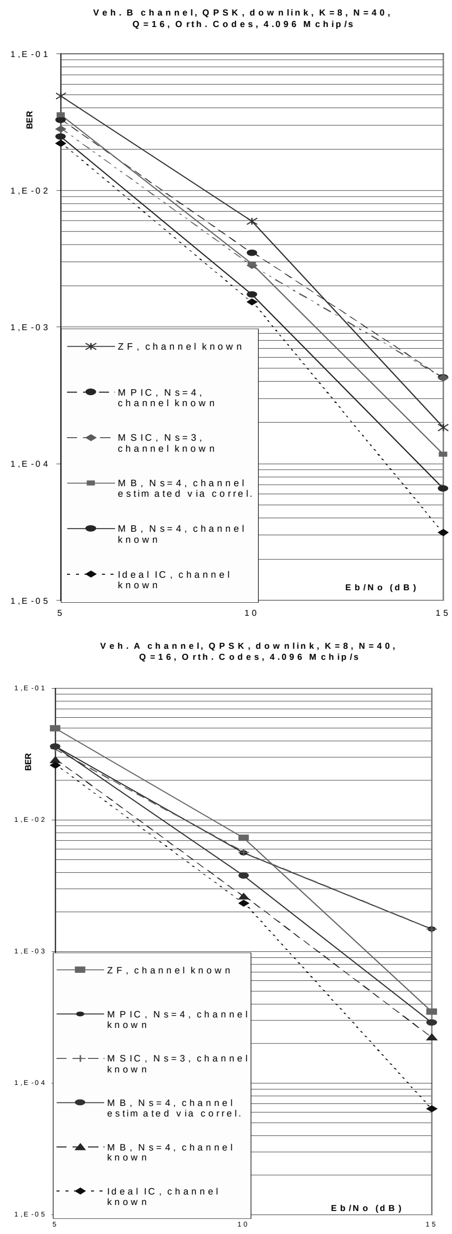

Fig.2 – Simulated BER vs SNR performance of the proposed MB receiver and comparison with ZF and MPIC/MSIC solutions for the TDD *downlink* Veh. B (top) and Veh. A (bottom) WSSUS test channels. Ns is the number of stages employed by the receiver. The case when the channel is estimated from a midamble sequence of p=512 chips via the cross-correlation technique and the ideal case of perfect IC are both considered.

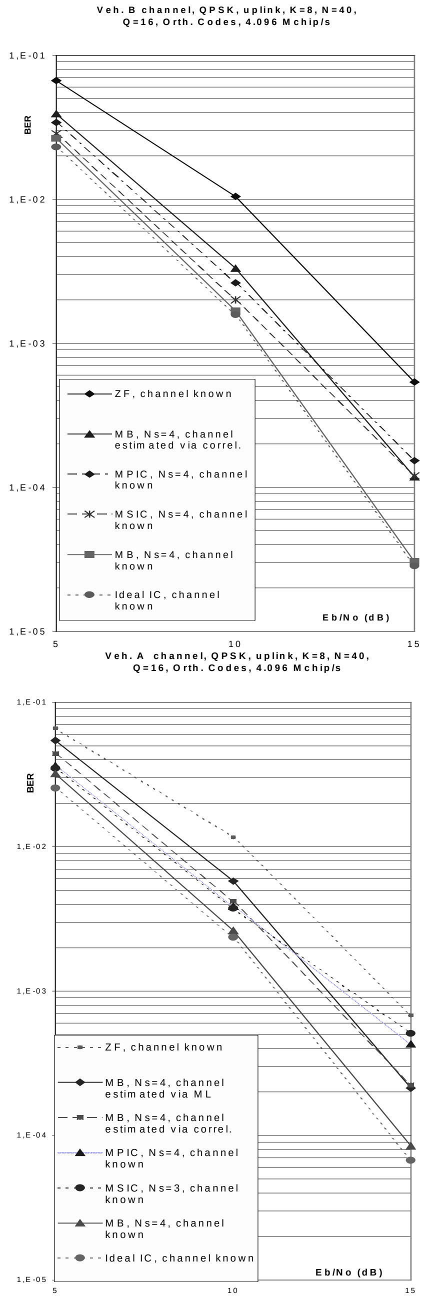

Fig.3 – Same as in Fig.2, for the *uplink* case. Both ML and cross-correlation (with four-stage interference cancellation, as described in the text) channel estimation techniques based on a midamble sequence of p=512 chips have been considered, together with the ideal case of perfect IC.

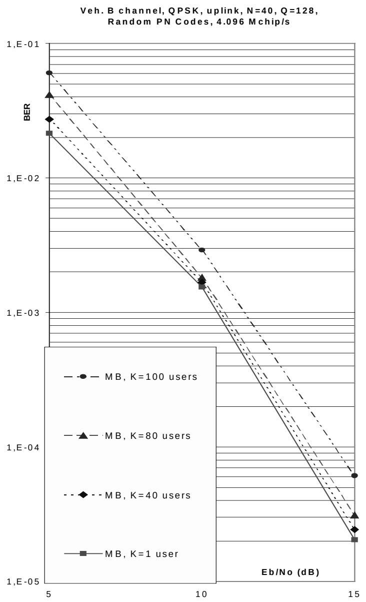

Fig.4 Simulated BER vs SNR performance for the Veh.B WSSUS test channel (uplink case, known channel) and for large spreading factor Q=128, as allowed in FDD (Ns=4).

tistage interference cancellation has been considered for the uplink, as described here. For user #i, the cross-correlation method gives a "tentative" channel estimate; this is employed by each of the other users to regenerate (at chip interval) the received midamble pertaining to user #i and subtract it from the received midamble before estimating (via cross-correlation) its own channel. The procedure is repeated until that a satisfactory channel estimate is obtained for all users. From Fig.3 we verify that quite good results (also better than ML @SNR=10 dB) are obtained, with a reduced computational cost.

5. Computational Issues and Conclusions

The arithmetical complexity of the MPIC, MSIC and MB solutions is that associated to eq.(2.1), i.e., $K(2L_g+1)-1$ complex products per user and per received sample, plus that of the CC (L complex product per user and per sample if L is the number of the non-negligible C-CIR coefficients) plus that of the non-linearity *F*[ ]. Although (2.4) seems slightly more complex than (2.3), both them can be computed via low-cost operations. The complexity of MPIC, MSIC and MB is then nearly the same and is much smaller than that of the ZF.
The effectiveness of the proposed MB-MUD receiver is then fully verified so that it constitutes a good candidate for TDD and FDD receivers. In particular, for the reference case of known channel its performance is close to the ideal interference canceller and is clearly better than the ZF while the performance loss due to imperfect channel estimation is not large.

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