Iterative Low-Complexity Receiver for the UMTS downlink

Récepteur itératif de faible complexité pour la liaison descendante de la norme UMTS

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Abstract

In this paper, an iterative low-complexity receiver is proposed for Code Division Multiple Access (CDMA) systems with small spreading factors. The UMTS (Universal Mobile Telecommunication System) radio interface based on CDMA, has been designed to offer a wide range of data rates using variable spreading factors. High data rate services are obtained by using small spreading factors. For such services, the spreading sequences have bad autocorrelation properties causing the degradation of the Rake receiver performance because of the InterSymbol Interference (ISI). To improve the receiver performance, we propose to add a Decision Feedback Sequence Estimation (DFSE) equalizer at the Rake receiver output. The DFSE is able to take into account a priori probability ratios and to deliver a posteriori probability ratios on symbol digits in order to exchange randomize soft information with the channel decoder, so that the proposed receiver benefits from turbo-processing gains. Channel estimation is also treated in an iterative fashion. The complete receiver is well suited to the UMTS downlink system as it drastically reduces the ISI while keeping a reasonable computational complexity.

**Keywords**: UMTS, Rake receiver, Intersymbol interference, Equalization.
Résumé

Dans cet article, nous proposons un récepteur itératif de faible complexité pour les systèmes à Accès Multiples à Répartition par Codes (AMRC) de faible facteur d’étalement. L’interface radio de la norme UMTS est basée sur l’AMRC et permet d’assurer des débits variables en modifiant la longueur de la séquence d’étalement. Ainsi, des services nécessitant des débits élevés pourront être assurés grâce à l’utilisation de petits facteurs d’étalement. Pour de tels services, les séquences d’étalement possèdent de mauvaises propriétés d’autocorrélation ce qui dégrade les performances du récepteur en réseau à cause de l’Interférence Entre Symboles (IES). Pour améliorer les performances du récepteur, nous proposons de rajouter un égaliseur DFSE (Decision Feedback Sequence Estimation) à la sortie du récepteur en réseau. L’égaliseur proposé permet aussi de délivrer une probabilité a posteriori des symboles décodés, nous pourrons ainsi bénéficier du gain de turbo égalisation en utilisant la sortie souple du décodeur canal. L’estimation de canal est aussi effectuée de manière itérative. Le récepteur complet est particulièrement bien adapté au lien descendant de la norme UMTS dans la mesure où il réduit significativement l’IES tout en ayant une complexité de calcul raisonnable.

Mots clés : UMTS, Récepteur en réseau, Interférence entre symboles, Égalisation.
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1 Introduction

In the UMTS specifications [1], the chip rate is fixed and the transmitted data rate is increased by using small spreading factors. For such services, the spreading sequences have bad autocorrelation properties causing the appearance of InterSymbol Interference (ISI). The study of the Rake receiver performance degradation due to ISI was developed in [2]. This study showed that we must use some equalization techniques when the spreading factor is smaller than sixteen.

ISI reduction was studied only for the downlink since the uplink suffers more from the MultiUser Interference (MUI) which is essentially due to the near-far problem and also because high data rate services are especially used in the downlink. Moreover, there is likely a small number of high data rate users in the downlink. The other users employ significantly higher spreading factors and thus are received at lower power. Therefore, the MUI experienced by the high data rate user is low and the Rake receiver performance is essentially limited by ISI in the downlink. Most of the conventional equalization techniques, used in the TDMA (Time Division Multiple Access) systems, were adapted to the UMTS downlink. Two interesting solutions were previously proposed [3]-[4] and [5]. The use of a Linear Minimum Mean Square Error (LMMSE) chip equalizer was introduced [3]-[4] in order to reduce both the ISI and the MUI. Note that in the downlink, the MUI is due to the multipath channel since the spreading sequences of the different users are orthogonal (Orthogonal Variable Spreading Factor (OVSF) sequences).
This solution showed a very interesting MUI reduction compared to the Rake receiver. However, this approach does not reduce much of the ISI because of the linear characteristic of the equalizer. It is well known that a close to optimal approach (optimal for sequence detection) for ISI reduction is the Maximum Likelihood Sequence Estimator (MLSE). The use of this kind of equalizer following the classical Rake receiver was recently proposed in [5], the whole receiver structure was called Rake-MLSE receiver. The obtained performance in the case of perfect channel knowledge was very near to the Matched Filter Bound (MFB) [6]-[8]. However, the complexity of this approach increases exponentially with the channel memory and the constellation size. Furthermore, the performance degradation due to channel estimation noise and bias was not taken into account in [5].

In this paper, we use a Decision Feedback Sequence Estimation (DFSE) equalizer [9] at the Rake receiver output (Rake-DFSE). This approach consists in reducing the trellis state complexity thanks to an embedded Decision Feedback per survivor structure [10], [9], often denominated “Per Survivor Processing” [11]. For TDMA systems, this approach suffers from error propagation and requires a pre-filter beforehand to turn the equivalent discrete time channel into minimum phase [12]. In CDMA contexts using long spreading sequences, such an approach is impossible as the equivalent channel model at the RAKE output varies from symbol to symbol. Fortunately, during the simulations, we observed that the equivalent channels originating from a Rake receiver output do not involve too much of error...
propagation in the DFSE. Thus, our Rake-DFSE gives close performance to that of the Rake-MLSE with much less computational complexity.

Mobile radio interfaces include channel coding and their performance should be assessed after the decoding. It is also well known that the performance of the decoder is improved if fed with soft value. We extend our receiver to accept a priori probability ratios and to deliver a posteriori probability (APP) ratios on symbol digits (Rake Soft-In-Soft-Out DFSE) [13]-[15]. Moreover, some sub-optimality results from performing equalization and decoding separately. A way to recover from it is to perform equalization and decoding iteratively [13]-[15]. An iterative receiver based on the Rake Soft-In Soft-Out (SISO) DFSE is finally considered.

We then focus on improving channel estimation quality. First, we noticed that conventional channel estimates suffer from ISI [22] so that the performance of the receiver degrades significantly. Hence, we propose to use the knowledge of the ISI structure in order to construct a MMSE (Minimum Mean Square Error) channel estimator. Moreover, we suggest to insert a data aided channel estimation process [2] within the turbo-detection loop. This iterative process consists in making a first decoding based on channel estimates obtained from the pilot symbols. Then, we use the estimated data at the channel decoder output to improve channel estimation quality for the next iteration. It is to be stressed that the re-estimated channel coefficients are used both by the Rake and the DFSE receivers via the equivalent channel model [16].
The paper is organized as follows. Section 2 gives the system model. Section 3 describes the equivalent channel model at the Rake receiver output. Section 4 focuses on the structure of the proposed low-complexity iterative receiver. The Rake SISO DFSE structure is presented, which is as a particular case of the Rake SISO MLSE. Section 5 describes the MMSE channel estimator and the data aided approach. Section 6 gives some simulation results. Finally, section 7 draws some conclusions.

2 System model

In this section, we present the system model used to study the UMTS downlink. For this link, the receiver performance is essentially limited by ISI therefore the presented model contains a single user as depicted in Figure 1. A data sequence \( \mathbf{u}_1^{\tau_0} = (\mathbf{u}_1, ..., \mathbf{u}_{\tau_0})^\top \) of \( \tau_0 \) symbols enters an encoder \( C_0 \) which outputs a coded sequence \( \mathbf{c}_1^{\tau_0} = (\mathbf{c}_1, ..., \mathbf{c}_{\tau_0})^\top \). Each data symbol \( \mathbf{u}_k = (u_{k,1}, ..., u_{k,k_0})^\top \) contains \( k_0 \) bits, whereas each coded symbol \( \mathbf{c}_k = (c_{k,1}, ..., c_{k,n_0})^\top \) contains \( n_0 \) bits. Coded bits are interleaved by an interleaver \( \Pi \) and punctured to match data rates to a transmitted format. Resulting bits are grouped into \( Q \)-ary symbols \( \mathbf{a}_k = (a_{k,1}, \cdots, a_{k,q})^\top \) containing \( q = \log_2(Q) \) bits. Each symbol \( \mathbf{a}_k \) finally pilots a \( Q \)-ary modulator that transmit the corresponding modulated symbol \( s_k \). According to the UMTS specifications [1], the symbols \( s_k \) are then spread using the product of a Walsh Hadamard sequence for channelization and a Gold sequence for scrambling. The so obtained chip sequence, \( d_k \), is decomposed in slots of
2560 chips. Note that $P$ spread pilot symbols are placed at the beginning of each slot in order to estimate the channel.

In the presence of multipath propagation, the received signal at the input of a spread spectrum receiver at time $t$ can be written as

$$r(t) = \sum_{l=1}^{L} f_l(t) \sum_{k} s_k \ e_k(t - kT_s - \tau_l(t)) + w(t),$$  \hspace{1cm} (1)

where $e_k(t) = \sum_{q=0}^{N-1} e_{kN+q} \ g(t - qT_c)$ is the spreading waveform for the $Q$-ary modulated symbol $s_k$, $\{e_q\}_q$ is the spreading sequence, $N$ is the spreading factor, $T_c$ and $T_s$ are respectively the chip and symbol periods, $g(t)$ is a square root raised cosine filter with roll-off 0.22, $L$ is the number of paths, $f_l(t)$ and $\tau_l(t)$ are respectively the complex amplitude and the delay of the $l$-th path and $w(t)$ is a white Gaussian noise with one-sided power spectral density $N_0$.

The system model (1) is used in the next section to derive the equivalent channel model at the Rake receiver output.

3 Equivalent channel model at the Rake receiver output

The Rake receiver consists in combining the output of correlators locked on the most significant paths. The Rake receiver decision variable used to estimate $s_k$ is therefore given by [23]

$$\hat{s}_k = \sum_{i=1}^{\hat{L}} \int_{-\frac{\Delta}{2}}^{\frac{\Delta}{2}} r(t)e_k^*(t - kT_s - \hat{\tau}_i)dt.$$  \hspace{1cm} (2)
where $\hat{L}$, $\hat{f}_i$ and $\hat{\tau}_i$ are estimates respectively of the number of the most significant paths, the complex amplitude and the delay of the $i$-th path.

Note that the estimates of path complex amplitudes are updated slot by slot using the pilot symbols whereas the number of paths and their delays are refreshed frame by frame since they vary slowly. Channel estimation procedures for the Rake receiver are discussed in [23]. For small spreading factors, the spreading sequence has bad autocorrelation properties causing the appearance of ISI. Since the spreading waveform can be approximated by a time-limited impulse response, the ISI is limited to some symbols before and after the symbol of interest. More precisely, the Rake receiver output can be written as

$$\hat{O}_k = \sum_{j=-L'}^{L'} g_j(k) s_{k-j} + w_k, \quad (3)$$

where

$$L' = 1 + \max_{i,j} \left( \frac{\tau_i - \tau_j}{T_c} \right),$$

$$g_j(k) = \sum_{1 \leq i \leq L} \hat{f}_i^* f_i \int e_{k-j}(t)e_k^*(t - jT_s + \tau_i - \hat{\tau}_i) dt,$$

and

$$w_k = \sum_{i=1}^{\hat{L}} \hat{f}_i^* \int w(t)e_k^*(t - kT_s - \hat{\tau}_i) dt.$$
symbol due to the use of a long scrambling sequence. Finally, note that
the channel coefficients, \( g_j(k) \), of the equivalent model are estimated using
estimates of path delays and complex path gains. The equivalent channel
modeling of the Rake receiver output (3) is at the basis of the DFSE
implementation which is described in the next section.

4 Low-complexity iterative receiver

Here, the equalizer is placed after the Rake receiver using the equivalent
channel model at its output which is described in section 3. A delay was
introduced after the Rake receiver in order to make the equivalent channel
model causal. Let \( y_k \) be the delayed samples

\[
y_k = \hat{y}_{K-L'} = \sum_{l=0}^{\mu} h_l(k)s_{k-l} + w_{k-L'}
\]

where \( h_l(k) = g_{n-L'}(k-L') \) and \( \mu = 2L' \). For convenience, we define
\( \mathbf{h}(k) \) as the vector of channel coefficients \([h_0(k), \ldots, h_{\mu}(k)]^T\). The rest of
this section is organised as follows: subsection 4.1 briefly recalls the turbo
detection principle whereas subsection 4.2 describes the SISO DFSE.

4.1 Turbo detection principle

In this work, the channel code is a (terminated) Recursive Systematic Con-
volutional (RSC) code, which, as well known, can be optimally decoded
(i.e., symbol-by-symbol) using the BCJR algorithm [17]. Inputs are log a
priori probability ratios on bits while outputs are log a posteriori probabil-
ity (APP) ratios on bits. From delivered APP on each bit of a sequence,
an extra knowledge, called extrinsic information, is drawn, which basically consists of the incremental information about that particular bit brought by information available from all other bits through the decoding process. The basic SISO module has been extensively described (see [18] [19] [20]).

We now recall the turbo detection principle (figure 2). The SISO ISI decoder delivers log a posteriori probability (log-APP) ratios on bits $a_{k,j}$ of symbols $a_k$ composing burst $a_1^T = (a_1, ..., a_r)^T$, aided with log a priori probability ratios on them coming from the decoder (null at the beginning) and given the received burst $y_1^T = (y_1, ..., y_r)^T$ and an estimate (or a re-estimate) $\hat{h}(k)$ of the (equivalent) channel coefficient vector at time instant $k$. As we will show thereafter, those log-APP ratios on bits can be divided into two parts according to the following formula:

$$\lambda_{\text{det}}(a_{k,j}) = \lambda_{\text{det}}^p(a_{k,j}) + \lambda_{\text{det}}^{\text{extr}}(a_{k,j})$$ \hspace{0.5cm} (5)

where $\lambda_{\text{det}}^p(a_{k,j})$ and $\lambda_{\text{det}}^{\text{extr}}(a_{k,j})$ are respectively the log a priori probability ratio and the log extrinsic probability ratio. After de-interleaving $\Pi^{-1}$, the overall sequence of log extrinsic probability ratios becomes a sequence of log probability ratios on bits of coded symbols for the channel decoder.

Similarly at the output of the SISO channel decoder, each log-APP ratio on coded bit $\lambda_{\text{dec}}(c_{k,j})$ can be split into an a priori part and an extrinsic part. The latter can be computed by subtracting bitwise the log a priori ratio $\lambda_{\text{dec}}^p(c_{k,j})$ at the input of the decoder from the corresponding log-APP ratio $\lambda_{\text{dec}}(c_{k,j})$ at the output, so that:
\[ \lambda_{\text{ext}}^{c_{k,j}} = \lambda_{\text{dec}}^{c_{k,j}} - \lambda_{\text{dec}}^{p_{k,j}} \]  

Sequence of log extrinsic probability ratios on coded bits is re-interleaved and passed to the SISO ISI decoder as new sequences (one per burst) of log a priori probability ratios on bits of bit-labeled symbols for a next detection attempt. Iterating the procedure a few times leads to a large improvement of the final Bit Error Rate (BER) and Frame Error Rate (FER) on data bit sequence.

### 4.2 Soft-input soft-output DFSE detector

To reduce the overall turbo detection complexity, sub-optimal trellis-based detectors must be introduced. Among the set of trellis-based reduced-states sequence estimators [10], the DFSE appears as the most suitable candidate because it is high regular, Viterbi-like structure, and the good performance it provides regarding its moderate complexity [9]. The SISO DFSE operates on a reduced-state trellis, where only \( \nu \) most recent symbols form the trellis state space \( S \) and the earlier \( \mu - \nu \) symbols are used through the embedded decision-feedback structure. Similar developments have been published in the literature in different contexts [13]-[15]. Thus, the following derivation of the SISO DFSE is made concise on purpose. The conditioning by \( \mathbf{h}(k) \) (or its estimate \( \hat{\mathbf{h}}(k) \)) is implicit and omitted for the ease of expressions. Let \( B \) denote the branch (or transition) space. Any trellis branch \( b = (s', s, e) \in B \) with \( s', s \in S^2 \), carries a label \( e \) which, at time \( k \), represents an input symbol
$a_k$ formed by bits $\{a_{k,1}, a_{k,2}, ..., a_{k,q}\}$. Each trellis path is a set of transitions starting from state $\zeta$ at $k = 0$ and terminating at state $\eta$ at $k = \tau$. An optimal bitwise detector would compute for $\forall k \in [1, \tau]$ and $\forall j \in [1, q]$ the log a posteriori probability ratio (LAPR) on a symbol bit $a_{k,j}$, defined as

$$\lambda_{\text{det}}(a_{k,j}) = \ln \frac{\Pr(a_{k,j}=1|y)}{\Pr(a_{k,j}=0|y)}.$$ 

The “max-log” approximation on soft decision outputs rather considers

$$\lambda_{\text{det}}(a_{k,j}) = \max_{a_i^t|a_{k,j}=1} \{ \ln p(\mathbf{a}, \mathbf{y}) \} - \max_{a_i^t|a_{k,j}=0} \{ \ln p(\mathbf{a}, \mathbf{y}) \}$$  \hspace{1cm} (7)$$

where $\{ -\ln p(\mathbf{a}, \mathbf{y}) \}$ is the trellis path metric associated with symbol sequence $a_i^t$.

(7) can also be written

$$\lambda_{\text{det}}(a_{k,j}) = \min_{b \in B|a_{k,j}=0} \left\{ \alpha_{k-1}(s') + \gamma_k(b) + \beta_k(s) \right\}$$

$$- \min_{b \in B|a_{k,j}=1} \left\{ \alpha_{k-1}(s') + \gamma_k(b) + \beta_k(s) \right\}$$

where $\gamma_k(b)$ is the DFSE branch metric,

$$\gamma_k(b) = \frac{1}{N_0} \bigg| y_k - \sum_{I=0}^{\nu} h_I(k)z_{k-I} - \sum_{I=\nu+1}^{\mu} h_I(k)\hat{z}_{k-I}^{(s',k-1)} \bigg|^2 ;$$  \hspace{1cm} (8)$$

$\alpha_{k-1}(s')$ is the accumulated metric of the best subpath starting from state $\zeta$ at time $k = 0$ and terminating in $s'$ at $k - 1$, $\beta_k(s)$ is the backward accumulated metric of the best subpath starting from state $\eta$ at time $k = \tau$ and terminating in $s$ at time $k$, and $\hat{a}_{k,j}^{(s',k-1)}$ denotes the estimated symbol bit $a_{k,j}$ from the survivor path attached to $s'$ at $k - 1$. $\alpha_k(s)$ and $\beta_{k-1}(s')$ can be recursively computed using the forward and backward recursion similar
to [17]:

\[
\alpha_k(s) = \min_{b \in B(s)} \{ \alpha_{k-1}(s') + \gamma_k(b) \} \quad (9)
\]

\[
\beta_{k-1}(s') = \min_{b \in B(s')} \{ \beta_k(s) + \gamma_k(b) \} \quad (10)
\]

where \( B(s) \) (respectively \( B(s') \)) is a set of all branches \((s', s)\) terminating in state \( s \) (respectively starting at state \( s' \)), \( B(s) \subset B \) (respectively \( B(s') \subset B \)); boundary conditions are (i) \( \alpha_0(\zeta) = 0, \alpha_0(s \neq \zeta) = \infty \) and (ii) \( \beta_\tau(\eta) = 0, \beta_\tau(s \neq \eta) = \infty \). The expression (8) involves the sequence \( \{a_{k-1}, \ldots, a_{k-\nu}\} \) forming state \( s' \) and the sequence \( \{a'_{k-\nu-1}, \ldots, a'_{k-\mu-1}\} \) obtained by reading off the survivor path terminating in \( s' \) at \( k-1 \).

The branch metric in (9) (8) in the presence of independent a priori probabilistic information on the transmitted symbol from decoder is modified as \( \gamma_k(b) - \gamma_k^b(b) \). Assuming perfect decorrelation between symbol bits after interleaving \( \Pi \), we have

\[
\gamma_k^b(b) = \ln \Pr(a_{k,j} = e_j) + \sum_{i=1, i \neq j}^q \ln \Pr(a_{k,i} = e_i) \quad (11)
\]

where all \( \Pr(a_{k,j}) \) are calculated from interleaved extrinsic probability ratios delivered by the decoder. At the output of the SISO DFSE, the bitwise soft output \( \lambda_{\text{det}}(a_{k,j}) \) is usually split into two parts

\[
\lambda_{\text{det}}(a_{k,j}) = \lambda_{\text{dec}}^b(a_{k,j}) + \lambda_{\text{det}}^{\text{extr}}(a_{k,j})
\]

where \( \lambda_{\text{dec}}^b(a_{k,j}) = \ln \frac{\Pr(a_{k,j} = 1)}{\Pr(a_{k,j} = 0)} \) is the log a priori probability ratio on symbol bit \( a_{k,j} \) provided by the SISO decoder; \( \lambda_{\text{det}}^{\text{extr}}(a_{k,j}) \) is the log extrinsic probability ratio (LEPR) that, in case of the SISO DFSE with forward-backward
recursion, may be expressed as

\[
\lambda_{\text{det}}^{\text{ext}}(a_{k,j}) = \min_{b \in B | a_{k,j} = 0} \left\{ \alpha_{k-1}(s') + \gamma_{k}^{\text{ext}}(b) + \beta_{k}(s) \right\} \\
- \min_{b \in B | a_{k,j} = 1} \left\{ \alpha_{k-1}(s') + \gamma_{k}^{\text{ext}}(b) + \beta_{k}(s) \right\}
\]

with

\[
\gamma_{k}^{\text{ext}}(b) = \gamma_{k}(b) - \sum_{i=1, i \neq j}^{q} \ln \Pr(a_{k,i} = e_{i}) .
\]  \hspace{1cm} (12)

Note that the effect of prior probabilistic information in turbo-detection is twofold. First, it is accumulated during forward recursion as shown in (9) (8) (11). Second, it is explicitly present in (12). It must be emphasized that in case \( \nu = \mu \), the SISO DFSE becomes formally equivalent to the Min-Log-BCJR algorithm (or SISO MLSE) applied on the full ISI channel trellis.

When considering processing on a reduced-state trellis, estimated sequences taken from the path history and involved in branch metric derivations introduce a degradation in performance, due to a possible error propagation effect. However, as noted in the introduction, it appears by simulation that equivalent channels at the output of the Rake receiver does not induce significant error propagation into the DFSE structure. As a result, the choice of \( \nu = 1 \) seems to be sufficient in the case of \( \mu = 2 \) which is proved by simulation in the following.
5 Channel estimation improvement

5.1 MMSE channel estimates

The conventional channel estimates, which are obtained by correlating and averaging over pilot symbols [23], become biased when the spreading factor is low [22]. In this section, we propose to improve channel estimation quality by using the knowledge of the ISI structure. The conventional channel estimate of the $k$-th path complex gain over the $l$-th slot is obtained by correlating and averaging over the pilot symbols as follows

$$\hat{f}_k(l) = \frac{1}{E_{\text{pilot}}} \frac{1}{LM+P-1} \sum_{i=LM}^{LM+P-1} s_i^* \int r(t) e_i^* (t - iT_s - \hat{\tau}_k) \, dt,$$

where $M$ is the number of symbols per slot and $E_{\text{pilot}}$ is the total energy of the pilot symbols of the slot.

After some developments, we deduce

$$\hat{f}(l) = \left( \hat{f}_1(l), \cdots, \hat{f}_L(l) \right)^T = M(l) f(l) + \mathbf{n},$$

where $f(l)$ is a vector containing the complex amplitude of paths over the $l$-th slot:

$$f(l) = (f_1(l), \cdots, f_L(l))^T,$$

$$M_{k,j}(l) = \frac{1}{E_{\text{pilot}}} \frac{1}{LM+P-1} \sum_{i=LM}^{LM+P-1} s_i^* \sum_{q=-L'}^{i+L'} \int e_{i-q}(t) e_i^* (t - qT_s + \tau_j - \hat{\tau}_k) \, dt,$$
and \( \mathbf{n} \) is the channel estimation noise which has a variance equal to \( N_0/E_{\text{pilot}} \).

Note that \( s_{i-q} \) is assumed to be null if it is not a pilot symbol when computing \( \mathbf{M} \). Hence, the conventional channel estimates are biased if path delays are separated by less than one chip period or if the spreading factor is low (less than sixteen). In order to improve channel estimation quality, we propose to use the following MMSE channel estimates for the \( l \)-th slot:

\[
\hat{\mathbf{f}}^{MMSE}(l) = \mathbf{L}^H_{MMSE}(l)\hat{\mathbf{f}}(l),
\]

where

\[
\mathbf{L}_{MMSE}(l) = \arg\min_{\mathbf{L}} \| \mathbf{L}^H \hat{\mathbf{f}}(l) - \mathbf{f}(l) \|.
\]

By using (14), we deduce

\[
\hat{\mathbf{f}}^{MMSE}(l) = \mathbf{M}^H(l) \left( \mathbf{M}(l)\mathbf{M}(l)^H + \frac{N_0}{E_{\text{pilot}}} \mathbf{I}_L \right)^{-1} \hat{\mathbf{f}}(l).
\]

A Least Square estimate can also be used, which is

\[
\hat{\mathbf{f}}^{LS}(l) = \left( \mathbf{M}(l)^H \mathbf{M}(l) \right)^{-1} \mathbf{M}(l)^H \hat{\mathbf{f}}(l).
\]

5.2 Iterative (turbo) channel estimation

Once an iterative receiver has been built for joint equalization and decoding, it is very tempting to add channel re-estimation in the turbo detection loop. This iterative channel estimation consists in making a first decoding based on channel estimates obtained from the pilot symbols. Then, we use the estimated data from the channel decoder soft output to improve channel estimation quality for the next iterations. Thus, the data aided channel estimation benefits from time diversity brought by interleaving and from
channel decoding efficiency. The full turbo receiver including channel re-
estimation and turbo detection is described in figure 3.

6 Simulation results

Simulations were conducted for a single user whose spreading factor is equal
to \( N = 4 \) and for a EQ-4 wireless channel which consists of four taps of equal
average powers and delays separated by the chip period. Each tap is a circu-
larly complex gaussian random variable. The maximum Doppler frequency
is set to 19 Hz corresponding to a mobile speed of 10 km/h and a carrier
frequency of 2 Ghz. Hence, the channel is almost constant over each slot
whose duration is 2/3 ms. The spreading sequence results from the superpo-
sition of a Walsh Hadamard sequence for channelization and a Gold sequence
for scrambling as defined in the UMTS standard [1]. The outer code is a
16-state rate-1/2 RSC code with generator polynomials \( (1, \frac{1+D+D^4}{1+D+D^2}) \)
generating a coded sequence \( \mathbf{c} \) whose size is the number of data bits per frame,
i.e., \( 15 \times 636 \times q \) bits (including tail), 15 is the number of slots per frame,
636 is the number of data symbols per slot, \( q \) is the number of bits per
symbol. The coded sequence \( \mathbf{c} \) is sent to a pseudo-random interleaver \( \Pi \),
whose output is divided into 15 slots of length 640 symbols including four
pilot symbols at their start. The reduced trellis for the DFSE has only 4
states.

Fig. 4 compares the performance of the different equalizers for a QPSK
modulation and a perfectly estimated channel. The Rake DFSE performance
is given for the first iteration. We see that the Rake DFSE (with four states) outperforms by far the LMMSE chip equalizer and keeps close to the Rake MLSE performance. The obtained performance is also close to the MFB.

These results are confirmed in figures 5-6 which compare the Rake DFSE first iteration performance and the LMMSE chip equalizer for QAM-16 and QAM-64 modulations, respectively. The choice of such modulations is primarily motivated by the fact that they are used in the UMTS high data rate packet access mode called HSDPA (High Speed Downlink Packet Access) [21]. In these simulations, the channel is estimated using four pilot symbols having the same symbol energy of the data symbols. For these modulations, we have not simulated the Rake MLSE approach because of its prohibitive complexity. We see that the Rake-DFSE receiver gives better performance than the LMMSE chip equalizer. However, the obtained performance is far from the MFB because of channel estimation noise and ISI bias.

Finally, figure 7 shows the performance of the complete receiver at the channel decoder output for a QPSK modulation. MMSE and iterative channel estimates were used to improve channel estimation quality. We observe that the MMSE channel estimator improves the receiver performance by 2 dB at high SNR (Signal to Noise Ratio). Gains come from turbo detection and turbo channel estimation. Moreover, we note that the proposed receiver performance is quite close to the coded MFB after two iterations. This means that the obtained performance is close to the receiver perfor-
7 Conclusion

We proposed a low-complexity iterative receiver for CDMA systems with small spreading factors. We showed that the SISO DFSE introduced to fight back the residual ISI after the Rake receiver gives close to optimal performance for spreading factor as low as four. The channel estimation issue was also treated. We observed that we must remove the channel estimation bias due to ISI by using MMSE or LS channel estimates. Finally, the turbo equalization principle was transposed onto that structure by using data aided channel estimates and turbo-detection between the SISO DFSE and the channel decoder.

References


Figure 1: Transmitted signal.

Signal émis.
Figure 2: Turbo-Detection using a SISO Rake-DFSE receiver.

Turbo détecteur utilisant le récepteur SISO Rake-DFSE.
Figure 3: Turbo-Detection using a SISO Rake-DFSE receiver and iterative channel estimates.

Turbo détecteur utilisant le récepteur SISO Rake-DFSE et une estimation de canal itérative.
Figure 4: Performance at the equalizer output for QPSK and perfect channel estimates.

Performance à la sortie de l’égaliseur pour MDP 4 et pour une estimation de canal parfaite.
Figure 5: Performance at the equalizer output for QAM 16

Performance à la sortie de l’égaliseur pour MAQ 16.
Figure 6: Performance at the equalizer output for QAM 64

Performance à la sortie de l’égaliseur pour MAQ 64.
Figure 7: Performance at the channel decoder output for QPSK.

Performances à la sortie du décodeur canal pour MDP4.
Biography

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