A ROBUST PILOT-ASSISTED EQUALISER FOR THE PARTIALLY LOADED DOWNLINK UMTS TDD

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Abstract: In this paper, we propose a robust pilot-assisted equalisation strategy for the partially loaded time-division duplex (TDD) component of the universal mobile telecommunications system (UMTS). In addition to training-based equalisation performed using the midamble of a data packet, some of the unused spreading codes are exploited to upload pilots in order to perform an additional semi-blind adaptation over the payload of a packet. The latter ensures continuous adaptation and better tracking performance. The affine projection concept along with the concurrent constant modulus algorithm (CMA) and decision-directed (DD) mode are implemented to update the equaliser weights. Computer simulations are used to assess the performance of the proposed adaptation strategy over various UMTS TDD time bursts.

Keywords: concurrent adaptation, affine projection scheme, constant modulus, decision directed, downlink UMTS TDD, spectrum efficiency.

1. INTRODUCTION

Frequescy Division Duplex (FDD) is the most commonly used component of the universal mobile telecommunications system (UMTS). However, the time division duplex (TDD) mode provides a high transmission rate, a more efficient use of the spectrum and flexible capacity allocation. It has previously become the basis for the third generation (3G) standard, and is possible to be selected as the main duplex mode operation for fourth generation (4G) systems [1]. The UMTS TDD mode provides uplink and downlink services within the same frequency bandwidth which are separated in time through the use of different time slots as shown in Fig. 1(a). In each time slot the contribution of each user, a so-called burst, is a combination of two data fields ,a midamble and a guard period as depicted in Fig. 1(b). The midamble is a training sequence used particularly for channel equalisation. The injection of the midample into the transmitted payload reduces the spectral efficiency of the whole UMTS TDD physical channel by up to 20%. Furthermore, hostile and fast fading channels might significantly degrade the system's performance, whereby a continuous adaptation over the whole time slot would be desirable to ensure acceptable tracking performance of an equaliser.

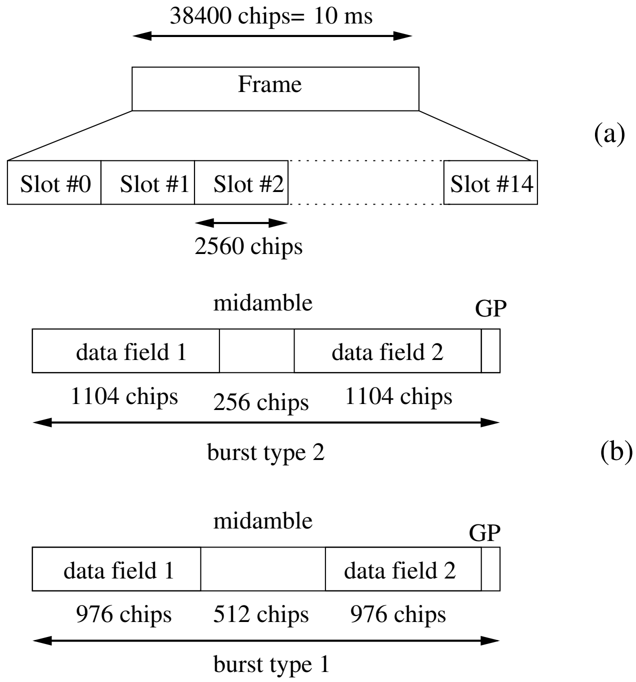

Fig. 1. Time structure in UMTS TDD: (a) basic frame structure, and (b) burst structure.

Blind approaches, which could ensure adaptation over data fields, have been performed using a constant modulus (CM) criterion [3, 4]. However, the typical slow convergence, the relatively high mean square error (MSE) level, and the phase ambiguity of such approaches limits the tracking performance of the receiver. In this paper, we aim i) to lower the MSE of the algorithm proposed in [4] by operating two equalisers — one driven by a CMA algorithm, the other by a decision directed criterion — concurrently similar to [6], ii) to speed up the convergence of [4] by employing the affine projection (AP) concept [7], and iii) to overcome the phase ambiguity of a CMA-based equaliser. We further want to achieve aims i) and ii) by exploiting a number of inactive users to upload pilots in case of a partially loaded system.

The paper is organised as follows. In Sec. 2 a description of the UMTS TDD physical channel is given. Based on the definition of a signal model in Sec. 3, a suitable cost function for the semi-blind adaptation is given in Sec. 4. In Sec. 5 we derive the concurrent affine projection adaptation scheme proposed for the downlink UMTS TDD. Simulations of the proposed algorithm are presented in Sec. 6, and conclusions drawn in Sec. 7.

2. UMTS TDD PHYSICAL CHANNEL

In the UMTS TDD physical channel, every 15 time slots form one frame, each frame has a duration of 10 ms [2] as shown in Fig. 1(a). Within every time slot a maximum of $N = 16$ users can transmit their signals simultaneously by means of different spreading codes. The contribution of each user is called a burst, which is a combination of two data fields, a midamble and a guard period as depicted in Fig. 1(b). There are two burst types proposed in [2], namely burst type 1 and burst type 2. As illustrated in Fig 1(b), both types have the same length of 2560 chips, concluded by guard period of 96 chips in order to avoid overlapping of consecutive time slots. Burst type 1 has a longer midamble (512 chips), suitable for channel conditions where long training periods are required for adaptation and tracking of an equaliser.

3. SIGNAL MODEL

We consider the UMTS-TDD downlink model in Fig. 2 with a maximum of N symbol-synchronous active users, which for simplicity are assumed to have the same rate. In the case of a partially loaded system with $K \leq N-1$, we assume the first K users with signals $u_l[n]$, $l = 0(1)K-1$, to be active, and the next $N_p \leq N-K$ to be pilots with signals $p_l[n]$, $l = 0(1)N_p-1$ while for the remaining $N-K-N_p$ user signals $z_l[n]$ are assumed to be zero. The signals $u_l[n]$ and $p_l[n]$ are code multiplexed using Walsh sequences of length N extracted from a Hadamard matrix $\mathbf{H}$. The resulting chip rate signal, running at N times the symbol rate, is further scrambled by $c[m]$ prior to transmission over a channel with dispersive impulse response $g[m]$ and corruption by additive white Gaussian noise $v[m]$, which is assumed to be independent of the transmitted signal $s[m]$.

The dispersive channel $g[m]$ destroys the orthogonality of the Walsh codes, such that direct decoding of the received signal $r[m]$ with descrambling by $c^*[m]$ and code-matched filtering by $\mathbf{H}^{\mathrm{T}}$ will lead to both multiple access interference and inter-symbol interference of the decoded user signals $\hat{u}_l[n]$, $l = 0(1)K-1$. In order to re-establish

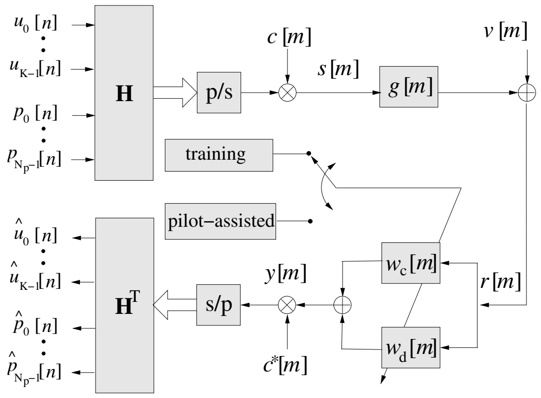

Fig. 2. Signal model.

orthogonality of the codes, a chip level equaliser $w[m]$ can be utilised. The equalisation is performed in both midamble period and data fields. In the former by means of the training sequence at the chip rate in the minimum mean-squared error (MMSE) sense [9], in the latter by using a blind or semi-blind scheme [4]. The equaliser $\mathbf{w}$ consists of an equaliser with a CM component $\mathbf{w_c}$ and a DD component $\mathbf{w_d}$ operated in parallel, such that $\mathbf{w} = \mathbf{w_c} + \mathbf{w_d}$. In the following, we are concerned with concurrently updating $\mathbf{w}$ by implementing the affine projection scheme.

4. SEMI-BLIND EQUALISATION CRITERIA

We first derive the detected user signals $\hat{u}_l[n]$ and the pilot signals $\hat{p}_l[n]$ as a function of the equaliser $w[m]$. Based on this, we state a suitable cost function based on which the equaliser can be adapted.

4.1. Demultiplexed User and pilot Signals

For the decoding, Walsh sequences are used as matched filters. The sequence for decoding the lth user, contained in a vector $\mathbf{h}_l$, can be taken from an $N \times N$ Hadamard matrix,

$$\mathbf{H}^{\mathrm{T}} = [\mathbf{h}_0 \;\; \mathbf{h}_1 \;\; \cdots \;\; \mathbf{h}_{N-1}]^{\mathrm{T}} . \tag{1}$$

The lth user is thus decoded as

$$\begin{aligned}\hat{u}_l[n] &= \mathbf{h}_l^{\mathrm{T}} \cdot \begin{bmatrix} c^*[nN] & & & \mathbf{0} \\ & c^*[nN-1] & & \\ & & \ddots & \\ \mathbf{0} & & & c^*[nN-N+1] \end{bmatrix} \cdot \begin{bmatrix} y[nN] \\ y[nN-1] \\ \vdots \\ y[nN-N+1] \end{bmatrix} \\ &= \tilde{\mathbf{h}}_l^{\mathrm{T}}[nN] \cdot \begin{bmatrix} \mathbf{w}^{\mathrm{H}} & & & \mathbf{0} \\ & \mathbf{w}^{\mathrm{H}} & & \\ & & \ddots & \\ \mathbf{0} & & & \mathbf{w}^{\mathrm{H}} \end{bmatrix} \cdot \begin{bmatrix} r[nN] \\ r[nN-1] \\ \vdots \\ r[nN-L-N+2] \end{bmatrix}\end{aligned}$$

whereby the descrambling code $c^*[m]$ has been absorbed into a modified and now time-varying code vector $\tilde{\mathbf{h}}_l[nN]$, and $\mathbf{w} \in \mathbb{C}^L$ contains the equaliser's L chip-spaced complex conjugate weights. Rearranging $\mathbf{w}$ and $\tilde{\mathbf{h}}_l[nN]$ yields

$$\begin{aligned}\hat{u}_l[n] &= \mathbf{w}^{\mathrm{H}} \cdot \begin{bmatrix} \tilde{\mathbf{h}}_l^{\mathrm{T}}[nN] & & \mathbf{0} \\ & \tilde{\mathbf{h}}_l^{\mathrm{T}}[nN] & \\ & & \ddots \\ \mathbf{0} & & \tilde{\mathbf{h}}_l^{\mathrm{T}}[nN] \end{bmatrix} \cdot \begin{bmatrix} r[nN] \\ r[nN-1] \\ \vdots \\ r[nN-L-N+2] \end{bmatrix} \\ &= \mathbf{w}^{\mathrm{H}}\, \mathbf{H}_l[nN]\, \mathbf{r}_{nN}, \qquad (2)\end{aligned}$$

and with similar analysis, the lth pilot's demultiplexed signal can be given as

$$\hat{p}_l[n] = \mathbf{w}^{\mathrm{H}}\, \mathbf{H}_l[nN]\, \mathbf{r}_{nN} \qquad (3)$$

with $\mathbf{H}_l[nN] \in \mathbb{C}^{L\times(N+L-1)}$ being a convolutional matrix comprising of the lth either user's or pilot's modified code vector $\tilde{\mathbf{h}}^{\mathrm{T}}[n]$ and $\mathbf{r}_{nN} \in \mathbb{C}^{N+L-1}$.

4.2. Cost Functions

Since the modulation scheme used for downlink UMTS-TDD is mainly the quaternary phase-shift keying (QPSK) (with some exceptions the 8PSK) [2], the K active user signals $u_l[n]$ consist of symbols with a constant modulus γ. By forcing all received user symbols $\hat{u}_l[n]$ onto γ and the received pilot symbols $\hat{p}_l[n]$ onto the known transmitted sequences $p_l[n]$, a semi-blind cost function ξ_c is proposed to adapt w_c weights. Note that the remaining $N-K-N_p$ inactive users $\hat{z}_l[n]$ should be taken into consideration, otherwise the equalisation criterion are under-determined. Accordingly, the signals $\hat{z}_l[n]$ are forced to zeros in MSE sense to ensure that the overall system is fully determined. Therefore, the proposed cost function ξ_c consists of three terms and is formulated as

$$\xi_c = \mathcal{E}\left\{\sum_{l=0}^{K-1}(\gamma^2-|\hat{u}_l[n]|^2)^2\right\}+\mathcal{E}\left\{\sum_{l=0}^{N_p-1}|p_l[n]-\hat{p}_l[n]|^2\right\}+ +\mathcal{E}\left\{\sum_{l=0}^{N-K-N_p-1}|\hat{z}_l[n]|^2\right\}, \qquad (4)$$

where $\mathcal{E}\{\cdot\}$ denotes the expectation operator. The optimum equaliser coefficient vector w_c in the CM sense is obtained from

$$\mathbf{w}_{c,\mathrm{opt}} = \arg\min_{\mathbf{w}_c} \xi_c \quad . \qquad (5)$$

Similarly, by employing a non-linearity $q(\cdot)$ that maps its input onto the the closest constellation point, the multiuser decision directed cost function ξ_d for the DD part can be formulated as

$$\xi_d = \mathcal{E}\left\{\sum_{l=0}^{K-1}|q(\hat{u}_l[n])-\hat{u}_l[n]|^2\right\}+\mathcal{E}\left\{\sum_{l=0}^{N_p-1}|p_l[n]-\hat{p}_l[n]|^2\right\}+ +\mathcal{E}\left\{\sum_{l=0}^{N-K-N_p-1}|\hat{z}_l[n]|^2\right\} \qquad (6)$$

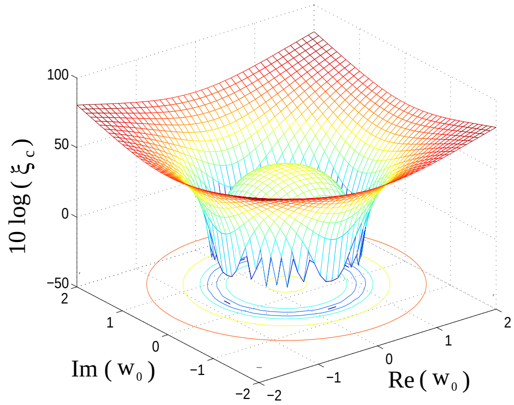

Fig. 3. Cost function ξ_c in dependency of a single complex valued coefficient w_0.

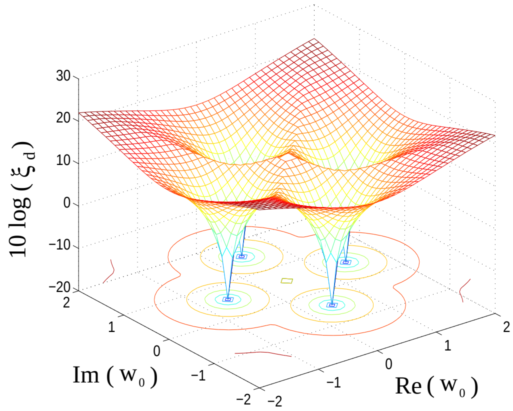

Fig. 4. Cost function ξ_d in dependency of a single complex valued coefficient w_0.

The optimum equaliser coefficient vector w_d in the mean square error sense based on the assumption of correct decisions is obtained from

$$\mathbf{w}_{d,\mathrm{opt}} = \arg\min_{\mathbf{w}_d} \xi_d \quad . \qquad (7)$$

In case where no pilot is loaded there are no unique solutions to either (5) or (7), since minimising (4) or (6) is ambiguous due to an indeterminism in phase rotation. Also, note that erroneous decisions are possible in (6) and therefore affect (7).

Example. In this example the two cost functions ξ_c and ξ_d are plotted in Figs. 3 and 4 respectively, in dependency of an equaliser with a single complex coefficient w_0. The system adopted here is a fully loaded UMTS TDD system with $N = 16$ users transmitting their signals over a distortionless and delayless channel with SNR=30dB. The modulation scheme employed here is QPSK with $\gamma = 1$. Fig. 3 shows that ξ_c exhibits a manifold of optimum solutions satisfying $|w_0| = \gamma$. Yet, only four solutions can be seen in ξ_d due to the four possible QPSK decisions, as depicted in Fig. 4.

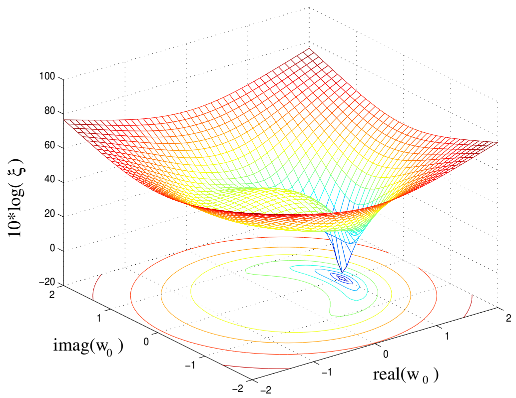

Fig. 5. Cost function ξ_c in dependency of a single complex valued coefficient w_0, for a partially loaded system with 10 active users and 6 pilots.

4.3. Phase ambiguity

Since an ambiguity with respect to a complex rotation $e^{j\varphi}$ ($\varphi \in [0; 2\pi]$) cannot be resolved by a CM criterion, this rotation invariance could be overcome by the use of the inactive codes to load pilot signals.

Example. To show how pilots overcome the phase ambiguity, the following example is presented. We assume a system with $K = 10$ active users and $N - K = 6$ pilots, over a distortion-less and delayless channel $g[n] = \delta[n]$. Thus, as shown in Fig. 5, the cost function ξ_c has one unique optimum solution $w_0 = 1$. Hence the phase ambiguity does not manifest itself any more.

4.4. Modified Cost Function

The CM term in (4) can be further reformulated as [7]

$$\mathcal{E}\left\{\sum_{l=0}^{K-1} |d_{l,c}[n]-\hat{u}_l[n]|^2\right\} \tag{8}$$

$$with \qquad d_{l,c}[n] = \gamma\frac{\hat{u}_l[n]}{|\hat{u}_l[n]|}. \tag{9}$$

This alternative CM philosophy suggests to enforce the detected symbol $\hat{u}_l[n]$ to its nearest symbol $d_{l,c}[n]$ from the circle which has the radius γ and the centre at the origin, as illustrated in Fig. 6. The new form has a structure similar to MSE and DD criteria, whereby the only difference between them is the value of the desired symbol $d_l[n]$. Consequently, both ξ_c and ξ_d can simply be written as

$$\xi_m = \mathcal{E}\left\{\sum_{l=0}^{N-1} |d_{l,m}[n]-b_l[n]|^2\right\}, \tag{10}$$

with $m \in \{c, d\}$ indicating the operational mode, CM or DD. The index $l = 0(1)N-1$ represents either active users for $l \le K-1$, pilots for $K \le l \le K+N_p-1$, or inactive users for $K+N_p \le l \le N-1$. Tab. 1 shows the various parameter values of the modified cost function ξ_m.

Next, we are concerned with minimising both ξ_c and ξ_d concurrently based on the affine projection scheme.

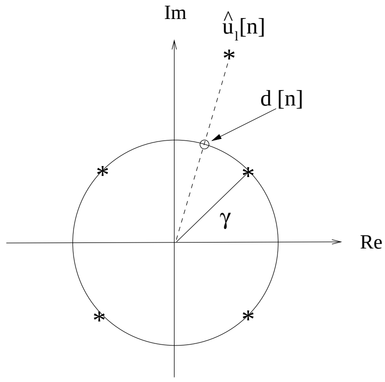

Fig. 6. Configuration of the desired response for the CM criterion, assuming a QPSK constellation

	Active user	Pilot	Inactive user
$b_l[n]$	$\hat{u}_l[n]$	$\hat{p}_l[n]$	$\hat{z}_l[n]$
$d_{l,c}[n]$	$\gamma\frac{\hat{u}_l[n]}{\lvert\hat{u}_l[n]\rvert}$	$p_l[n]$	0
$d_{l,d}[n]$	$\mathrm{q}(\hat{u}_l[n])$	$p_l[n]$	0

Table 1. Parameter values of the generalised cost function ξ_m.

5. CONCURRENT AFFINE PROJECTION ADAPTATION

In this section we consider the Affine Projection Algorithm (APA) being a popular algorithm within the acoustic echo cancellation schemes [10]. In the pth order of the APA algorithm, the current and last p data vectors are explicitly taken into account for updating. Therefore, it is convenient to define: $\mathbf{x}_l[n] = \mathbf{H}_l[nN]\mathbf{r}_{nN}$ and

$$\mathbf{X}_l[n]=[\mathbf{x}_l[n]\ \mathbf{x}_l[n-1]\ \cdots\ \mathbf{x}_l[n-p+1]], \tag{11}$$

$$\mathbf{d}_{l,m}[n]=[d_{l,m}[n]\ d_{l,m}[n-1]\ \cdots\ d_{l,m}[n-p+1]]^{\mathrm{T}}. \tag{12}$$

Hence, the implementation of the pth order algorithm could be summarised as shown in Tab. 2, where μ_c and μ_d are the relaxation factors and α is a small number used for weighting the identity matrix $\mathbf{I}$. The indicator $\delta(.)$ is a vectorial decision function. Therefore, $\mathbf{\Lambda}_l[n]$ disables the DD adaptation step for a specific user if the CMA adaptation step leads to an alteration in the decision.

The convergence of this concurrent scheme is governed by the step sizes. In practice, the DD step size μ_d can often be chosen much larger than the CMA step size μ_c. However, choosing too large values can cause serious error propagation due to incorrect decisions [6].

pth order concurrent affine projection algorithm	
1:	update $,\mathbf{X}_l[n],\ \mathbf{d}_{l,c}[n]$ and $\mathbf{d}_{l,d}[n]$ for $l = 0(1)N-1$
2:	$\mathbf{R}_l^{-1}[n] = (\mathbf{X}_l[n]^{\mathrm{H}}\mathbf{X}_l[n] + \alpha\mathbf{I})^{-1}$
3:	$\mathbf{e}_{l,c}[n] = \mathbf{d}_{l,c}[n] - \mathbf{X}_l^{\mathrm{T}}[n]\mathbf{w}^*[n]$
4:	$\mathbf{w}_c[n+1] = \mathbf{w}_c[n] + \mu_c \sum_{l=0}^{N-1} \mathbf{X}_l[n]\mathbf{R}_l^{-1}[n]\mathbf{e}_{l,c}^*[n]$
5:	$\tilde{\mathbf{b}}_l[n] = \mathbf{X}_l^{\mathrm{T}}[n]\mathbf{w}_c^*[n+1] + \mathbf{X}_l^{\mathrm{T}}[n]\mathbf{w}_d^*[n]$
6:	$\mathbf{\Lambda}_l[n] = \mathrm{diag}(\delta\{\mathbf{q}(\tilde{\mathbf{b}}_l[n]) - \mathbf{d}_{l,d}[n]\})$
7	$\mathbf{e}_{l,d}[n] = \mathbf{d}_{l,d}[n] - \mathbf{X}_l^{\mathrm{T}}[n]\mathbf{w}^*[n]$
8	$\mathbf{w}_d[n+1] = \mathbf{w}_d[n] + \mu_d \sum_{l=0}^{N-1} \mathbf{X}_l[n]\mathbf{R}_l^{-1}[n]\mathbf{\Lambda}_l[n]\mathbf{e}_{l,d}^*[n]$
9	$\mathbf{w}[n+1] = \mathbf{w}_c[n+1] + \mathbf{w}_d[n+1]$

Table 2. Concurrent affine projection algorithm for pilot-assisted multiuser equalisation.

The potential drawback of DD adaptation is the probability of error propagation occurring in case of a wrong hard decision, which subsequently degrades the performance. It was shown that if the equaliser's hard decision before and after the CM adaptation are the same then the decision is likely to be correct [6].

6. SIMULATION RESULTS

In order to demonstrate the convergence behaviour of the proposed algorithm, we transmit $K = 14$ QPSK active user signals and $N - K = 2$ pilots over a noise-free and a dispersive channel $g[m]$, represented by its transfer function $G(z) = 0.84 + (0.42 - 0.34j)z^{-1} + 0.09z^{-2}$. The length of the equaliser is $L = 20$, and the relaxation factor $\mu_c = 0.05$. The adaptation is initialised with the first coefficients in both weight vectors w_c and w_d set to 0.5. The MSE curves of the proposed algorithm on different scenarios over three UMTS TDD bursts are shown in Fig. 7.

As evident from Fig. 7, by operating the proposed algorithm with $\mu_d = 0$ (only CM branch is active) over bursts of type 2 (short training period), a better MSE performance is reached as compared to case where only training is performed in type 1 (larger midamble). The shortening of the midamble at no performance gain is equivalent to an increase in data throughput of 13%. Furthermore, faster convergence is obtained by either activating the DD equaliser ($\mu_d = 0.1$) or increasing the algorithm's order to $p = 2$ and $p = 5$.

7. CONCLUSIONS

A concurrent affine projection algorithm for pilot-assisted multiuser equalisation, suitable for UMTS TDD downlink scenario, has been derived. The algorithm provides continuous channel tracking and presents better convergence behaviour over the basic training equalisation even with longer midambles, whereby advantages in terms of data rate and spectrum efficiency can be achieved. The convergence can be accelerated by either activating the DD equaliser or increasing the affine projection algorithm's order.

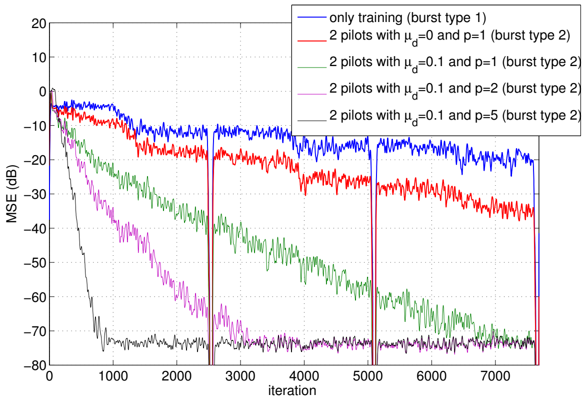

Fig. 7. MSE curves.

8. REFERENCES

[1] R. Esmailzadeh, M. Nakagawa, A. Jones, "TDD-CDMA for the 4th Generation of Wireless Communication," *IEEE Wireless Communications*, **10**(4):8–15, Aug 2003.

[2] TS. 25.233, 3GPP, "Spreading and Modulation (TDD)," *Third Generation Partnership Project*, vol.5.3.0, April 2003.

[3] C. Papadias and A. Paulraj, "A Constant Modulus Algorithm for Multiuser Signal Separation in Presence of Delay Spread Using Antenna Arrays," *IEEE SP Let.*, **4**(6):178–181, 1997.

[4] M. Hadef, S. Weiss and M. Rupp, "Adaptive Blind Multiuser DS-CDMA Downlink Equaliser," in *Electronics Letters.*, **41**(21): 1184 - 1185 , 2005.

[5] M. Hadef and S. Weiss, "A Pilot-Assisted Equalisation Scheme for the UMTS-TDD Downlink with Partial Loading ," in *IEEE VTC spring conference* , May 2005.

[6] F.C.C. De Castro, M.C.F. De Castro and D.S. Arantes "Concurrent Blind Deconvolution for Channel Equalisation ," *IEEE Conference ICC 2001*, **2**: 366–371, 2001.

[7] C.B. Papadias and D.T.M. Slock, "Normalized Sliding Window Constant Modulus and Decision-Directed Algorithms: a Link Between Blind Equalization and Classical Adaptive Filtering," *IEEE Trans.-Sig. Proc.*, **45**(1):231–235, 1997.

[8] C. R. Johnson, P. Schniter, T. J. Endres, J. D. Behm, D. R. Brown, and R. A. Casas, "Blind Equalization Using the Constant Modulus Criterion: A Review," *Proc. IEEE*, **86**(10): 1927–1950, 1998.

[9] P. Schniter and A.R. Margetts, "Adaptive Chip-Rate Equalization of Downlink Multirate Wideband CDMA," *Asilomar Conference on Signals, Systems, and Computers*, pp. 1228–1232, November 2002.

[10] S. Makino, J. Noebauer, Y. Haneda, and A. Nakagawa "SSB Subband Echo Canceller Using Low-Order Projection Algorithm," *Proc. IEEE International Conference on Acoustics, Speech, and Signal Processing*, **2**: 945–948, May 1996.