Adaptive Complexity Equalization for the Downlink in WCDMA Systems

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***Abstract*— We consider the issue of terminal reconfigurability in the downlink of WCDMA. For the purpose of optimizing power consumption in mobile terminals, we propose an adaptive-complexity equalization algorithm, which adapts the equalization length to the environment. A simple approach in WCDMA systems consists of computing the equalizer coefficients in frequency domain, and carry out channel equalization in time domain. The equalizer can be easily computed in frequency domain from channel estimates, which are generally obtained through pilot symbols. In our work, we decouple the task of adaptive complexity equalization in two parallel operations: variable length equalization and equalization length control.**

We propose a practical scheme to reduce equalization complexity by prewindowing in frequency domain and performing IFFT of variable length. An element of length control monitors the equalizer coefficients, updating the equalizer length at each stage. Both simulation and experimental results in outdoor-to-indoor scenarios show good performance and significant power savings with respect to full length equalization.

I. Introduction

The increasing demand for dynamic and multiservice mobile networks have triggered research on reconfigurable radio systems, based on software-defined radio (SDR) technology. Reconfigurability is envisaged not only at the mobile terminals [1], but also at the networks and the services they provide [2] [3]. The Telecommunications community has shown great interest in this area, both from an industrial and academic point of view. Several R&D projects funded by the European Union, i.e. CAST and TRUST and more recently E^2R [4] (End-to-End Reconfigurabily), clearly show the importance of reconfiguration in mobile networks.

The evolution of wireless communication systems towards reconfigurable networks requires mobile terminals with reconfiguration capabilities. In [5], an example of the design and construction of a reconfigurable radio system is presented, which enables the terminal to switch between different network services. At the physical layer, the interest is mainly focused on adaptive antennas [6], adaptive modulation [7] and adaptive signal processing.

The work reported herein was developed within the framework of the European project E^2R, which addresses End-to-End reconfigurability. Eurécom's research is partially supported by its industrial partners: Ascom, Swisscom, Thales Communications, ST Microelectronics, CEGETEL, Motorola, France Télécom, Bouygues Telecom, Hitachi Europe Ltd. and Texas Instruments.

Power management at mobile stations can be enhanced by means of run-time management of the transceiver processing complexity, thus extending terminal autonomy. Adapting the complexity of the signal processing algorithms to the environment can decrease power consumption while preserving system performance. In our work, we propose to adapt the receiver to the channel conditions by performing adaptive channel equalization.

We present a method for adaptive complexity equalization in the downlink of WCDMA systems, decoupling the problem in two parallel operations: variable length equalization and equalization length control. The receiver computes the equalizer coefficients in frequency domain, and carries out channel equalization in time domain. An alternative approach for adaptive equalization in time domain is the NLMS (or LMS) algorithm, as proposed in [8] for HSDPA. However, since the LMS requires computation of the correlation coefficients, its complexity may be unaffordable in situations with large delay spread. On the other hand, the equalizer can be easily computed in frequency domain from channel estimates, which are generally obtained through pilot symbols.

The proposed scheme enables terminal reconfigurability by adapting the (chip-level) equalization length to the environment conditions, namely the delay spread introduced by the channel. In practical scenarios, this translates into a less complex reception algorithm and thus a more power-efficient terminal. In addition, an equalizer length adapted to the channel provides noise supression beyond the actual channel length, improving the system performance.

In order to study the complexity/performance benefits of this approach, we set the equalization length equal to the midamble size (full length equalization) regardless of the channel conditions, and compare it with the proposed equalizer with adaptive length. As we show both through simulations and outdoor-to-indoor real experiments (long delay spread), the proposed receiver with adaptive equalization provides significant power savings with respect to full length equalization, while providing good performance.

The paper is organized as follows. Section II introduces the system model and full length equalization method. The proposed adaptive complexity equalizer is described in Section III. Section IV shows MATLAB simulation results and Section V experimental results. Conclusions are drawn in Section VI.

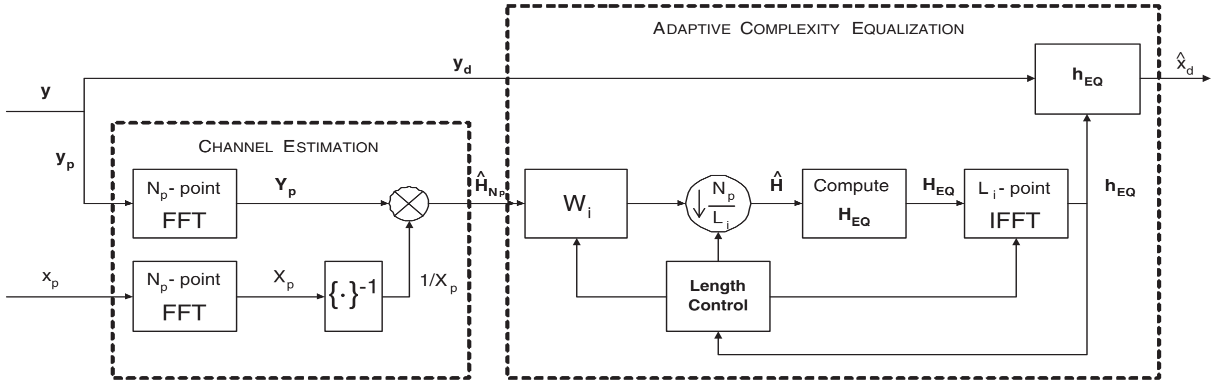

Fig. 1. Receiver block diagram - Channel estimation and adaptive complexity equalization

II. System description

In this section, we provide a system model for the received WCDMA signal. The block diagram in Fig. 1 depicts the stages of channel estimation and adaptive complexity equalization at the WCDMA receiver. The proposed approach for adaptive complexity equalization is described in detail in next section.

We consider a system with N_d data symbols and N_p pilot symbols per burst. The received WCDMA burst $\mathbf{y}$ is sampled at rate M/T_c, where T_c and M are the chip period and oversampling factor, respectively. The transmission model for the data symbols can be described as follows

$$\mathbf{y}_d[n] = x_d[n] * \mathbf{h}[n] + \mathbf{v}_d[n] \tag{1}$$

where x_d, $\mathbf{h}$ and $\mathbf{v}_d$ denote the transmitted data symbols, channel impulse response and additive noise, respectively. Due to oversampling, $\mathbf{y}_d$, $\mathbf{h}$ and $\mathbf{v}_d$ are vectors of dimension Mx1. Time index within the burst is denoted by n. The transmitted pilot sequence contains a cyclic prefix, so that the convolution with the channel becomes circular. Equation (2) describes the transmission model for the pilot symbols within a burst.

$$\mathbf{y}_p[n] = x_p[n] \odot \mathbf{h}[n] + \mathbf{v}_p[n] \tag{2}$$

The transmitted pilot-symbol sequence is represented by x_p, while $\mathbf{y}_p$ and $\mathbf{v}_p$ have dimension Mx1 and denote the received sequence and additive noise, respectively. Circular convolution is denoted by $\odot$.

In order to obtain the channel estimates in frequency domain, the receiver performs N_p-point FFT on each of the M components of $\mathbf{y}_p$. The resulting sequences in frequency domain are multiplied by the stored sequence $1/X_p$, where X_p denotes the FFT of the pilot symbols. For the i-th channel component and over the k-th frequency tone, the receiver estimates the channel as follows.

$$\widehat{H}_{N_p i}[k] = \frac{Y_{p_i}[k]}{X_p[k]} = H_i[k] + \frac{V_{p_i}[k]}{X_p[k]} \tag{3}$$

where Y_{p_i}, V_{p_i} and H_i are the FFT of the received pilot symbols, additive noise and true channel respectively. This method of channel estimation is usual in WCDMA systems, since it involves low complexity. However, in practice this technique leads to noisy channel estimates due to the second term of the addition in (3). Following channel estimation, an equalizer of variable length is computed in frequency domain and transformed to time domain through IFFT. The complexity of equalization is adaptive, since its length gets adjusted to the channel conditions, as we show in next section. In an approach with full length equalization (non adaptive), the equalizer is transformed from frequency to time domain via N_p-point IFFT, whereas the proposed approach with adaptive complexity equalization yields a time domain equalizer of length L_i. Computation of the equalizers in frequency domain $\mathbf{H}_{EQ}$ involves low complexity. The Zero-Forcig (ZF) and Minimum Mean-Squared Error (MMSE) equalizers are given by [9]

$$\mathbf{H}_{ZF}^T[k] = \frac{\widehat{\mathbf{H}}[k]^H}{\widehat{\mathbf{H}}[k]^H \widehat{\mathbf{H}}[k]} \tag{4}$$

$$\mathbf{H}_{MMSE}^T[k] = \frac{\widehat{\mathbf{H}}[k]^H}{\widehat{\mathbf{H}}[k]^H \widehat{\mathbf{H}}[k] + \sigma_{v_d}^2/\sigma_{x_d}^2} \tag{5}$$

where $\widehat{\mathbf{H}}$ is the channel estimate with dimension Mx1, $\sigma_{x_d}^2$ is the variance of the transmitted data symbols and $\sigma_{v_d}^2$ is the variance of the additive noise. Finally, the resulting estimated data symbols are computed as

$$\widehat{x}_d[n] = \mathbf{h}_{EQ}^T[n] * (\mathbf{h}[n] * x_d[n] + \mathbf{v}_d[n]) \tag{6}$$

where $\mathbf{h}_{EQ}$ denotes the equalizer in time domain, either $\mathbf{h}_{ZF}$ or $\mathbf{h}_{MMSE}$.

III. Adaptive complexity equalization

The performance of the equalizer is clearly limited by the quality of the channel estimates computed in frequency domain. Hence, in order to improve the system performance, we can either improve the quality of the channel estimates or improve directly the equalizer computed through noisy channel estimates. The proposed scheme computes the equalizer $\mathbf{H}_{EQ}$ on the basis of an improved set of channel estimates. In next section, we show by simulations that both approches exhibit similar performances, being the proposed option less complex. The channel estimates are smoothed (prefiltered) and downsampled in frequency domain. A reduced length L_i-point IFFT over the resulting frequency-domain equalizer is performed, as opposed to the N_p-point IFFT in full length equalization. Hence, complexity reduction comes from *a)* reduction in the number of coefficients $\mathbf{H}_{EQ}[k]$ to be computed b) reduced-length IFFT and c) reduction in the number of equalizer coefficients in time domain to perform the equalization operation.

This section describes a method to perform variable length equalization. However, to adapt the most appropriate length to the channel conditions, it is necessary to introduce length control. We propose to decouple the problem of variable length equalization and length control.

A. Variable length equalization

Prior to execution of the algorithm, each mobile terminal counts on a set possible window lengths $L = \{L_1, L_2, \cdots, L_K\}$. Assuming a rectangular window $w[n]$ for simplicity, the mobile stores the N_p-point FFTs of each L_i-tap window. Even though the support region of the equalizer is in the anti-causal part, the window is forced to be (circularly) symmetric, so that when transformed to frequency domain the resulting sequence is real. The time domain window has non-zero taps in time positions $[0, Li/2-1]$ and $[Np-Li/2, Np-1]$. Note that the dual operation of windowing with a rectangular window in time, corresponds to convolution with a $sinc$ in frequency domain. Hence, in frequency domain, the obtained sequence is

$$W_i\left[k\right] = FFT\left\{w\left[n\right]\right\} = L_i sinc\left[k\right] \tag{7}$$

which has real values due to the symmetry of $w[n]$ and zeroes at $k = m \cdot N_p/L_i, m = 1, \cdots, L_i - 1$. Since each W_i filter can be prestored in memory, this operation is perfomed only once, not during runtime, and does not contribute to increase complexity. By convolving the obtained channel estimates with the stored filters W_i, channel taps beyond the time domain window are neglected. In practice, in order to further decrease the complexity of convolution with W_i, only a subset of samples of W_i with amplitude above a certain threshold are taken into account. This can greatly simplify the convolution operation while introducing negligible distortion.

The algorithm executed during runtime is described as follows. At each slot, for the selected window length (set by the control element described in next subsection), the mobile computes circular convolution between the N_p-tap channel estimates $\widehat{\mathbf{H}}_{N_p}$ and W_i, subsampling by a factor N_p/L_i as depicted in Fig. 1, yielding $\widehat{\mathbf{H}}$. Actually, instead of computing the output of the convolution at N_p taps and subsample, the mobile just computes a subset of frequency positions $k = m \cdot N_p/L_i, m = 0, \cdots, L_i - 1$, which is an equivalent operation. Hence, the operation to obtain the i-th channel component of $\widehat{\mathbf{H}}$ is described as

$$\widehat{H}_i\left[m\right] = \sum_{l=0}^{N_p-1} \widehat{H}_{N_p i}\left[l\right] W_i \left[m \cdot \frac{N_p}{L_i} - l\right]_{modN_p} \quad m = 0, \ldots, L_i - 1 \tag{8}$$

This yields a smoothed version of the channel estimates in frequency domain, which now has L_i taps instead of N_p. Next, the receiver computes the equalizer from the obtained smoothed channel estimates as described in equation (2). Finally, the equalizer is transformed to time domain with an L_i-point IFFT. The first $L_i/2$ samples correspond to the causal part of the equalizer and the last $L_i/2$ samples correspond to the anti-causal part. Assuming that the support region of the equalizer is in the anti-causal part (appropiately setting the reference time in a synchronized system), only the last $L_i/2$ samples are considered for equalization, setting the remaining $N_p - Li/2$ samples to zero.

The algorithm can be extended to any type of window, not necessarily rectangular, if a certain equalizer shaping is desired. Complexity reduction with respect to full length equalization follows mainly from the fact that L_i-point IFFT is performed instead of an N_p-point IFFT. In addition, the complexity of the equalization operation is also reduced, since instead of convolving the received signal with an N_p-tap filter, it gets convolved with a sequence of $L_i/2$ taps. On the other hand, the proposed method introduces low complexity by convolving a real $sinc$ with the frequency-domain channel estimates.

B. Equalization length control

We propose a low-complexity mechanism of length control. To adapt the length of the window, we propose to evaluate the equalizer coefficients (in time domain) computed at the previous slot. If the amplitude of coefficients at the edges of the window is very small, the window length is reduced at the current slot. On the other hand, if their amplitude is significant, the window length is increased. This is in practice done by looking at the relative amplitudes, i.e. at the ratio of a given amplitude and $\max\left(h_{EQ}[n]\right)$. A hysteresi cycle can be defined to control the transitions $L_i \rightarrow L_{i+1}$ and $L_i \rightarrow L_{i-1}$, setting appropiately the cycle parameters: minimum/maximum amplitudes for transition and delimitation of equalizer edges.

As a possible extension, an additional element of control may be included to monitor the evolution of channel paths beyond the equalizer length, similarly to the path searcher in a rake receiver [10]. However, this is not necessary in practical scenarios, as we later show from experimental results.

IV. Simulation results

In this section we illustrate the performance benefits of our approach by simulation, in terms of Symbol Error Rate (SER). We compare a system with full length equalization and a system with fixed equalizer length adapted to the length of the channel. The aspect of equalization length control is evaluated in next section through experiments in a real scenario, adapting the receiver complexity needs to the time-varying multipath channel. Hence, while here we show the benefits of variable length equalization, next section incorporates length control and evaluates the proposed adaptive complexity equalization approach.

The burst structure corresponds to the type-2 in the 3.84 Mcps option, which are transmitted through the broadcast channel (BCH). As described in 3GPP/TDD [11], this corresponds to 2x1104 data symbols and a midamble of 192 pilot symbols per burst. We consider spreading factor equal to 1, oversampling factor $M = 2$ and 8-PSK symbols. In order to highlight the performance loss due to full length equalization, we transmit uncoded symbols. We consider a dispersive channel, with correlated taps and an exponential power delay profile. The simulated random channel has fixed length equal to 12 symbol periods, and the equalizer with reduced length is fixed to twice the channel length.

Fig. (2) shows a performance comparison in terms of SER for different equalizer configurations. The worst performance corresponds to the one achieved by a receiver with Matched Filtering (MF). The MF is computed in frequency domain from channel estimates and transformed to time domain via a full length IFFT, equal to the midamble length (192). A slight improvement in performance is provided by an MMSE equalizer computed in frequency domain and also transformed to time domain via full length IFFT. Clearly, the performance of the simulated system is limited by the quality of the channel estimates, since even a full length MMSE equalizer practically does not improve the system performance as Fig. (2) shows.

The system performance is significantly improved when applying the proposed method with reduced length, computed on the basis of the frequency-domain channel estimates, which after smoothing are transformed via a $2L$-point IFFT, being L the channel length. In addition, the proposed approach has lower computational complexity than full length equalization. For comparison, we also plot the performance of a system were smoothing is directly applied on the full length MMSE equalizer rather than on the frequency-domain channel estimates. The simulations show that prewindowing the channel estimates to eliminate samples beyond the actual channel length is more beneficial than smoothing an equalizer computed from noisy channel estimates.

V. EXPERIMENTAL RESULTS

A. System parameters

In a real scenario, the algorithm has to be able cope with both dispersive an non-dispersive channels, adapting its length according to the environment, both in indoor and outdoor communications. In order to test the proposed approach, we use the real-time reconfigurable radio platform of the mobile communications department at Institut Eurecom, testing the performance for boot indoor and oudoor communications. Currently, the radio equipment is designed for unpaired spectrum using Time Division Duplex (TDD) multiplexing with a 5 MHz channel bandwidth. The RF equipment operates in any of the 1900-1920 MHz IMT-2000 TDD bands. The implemented radio protocols (PHY/MAC/RLC) comply with the Release 4 Access Stratum specifications from the 3GPP, which can be found in [12]. We work with the 3.84 MChip/s mode of 3GPP/TDD operation.

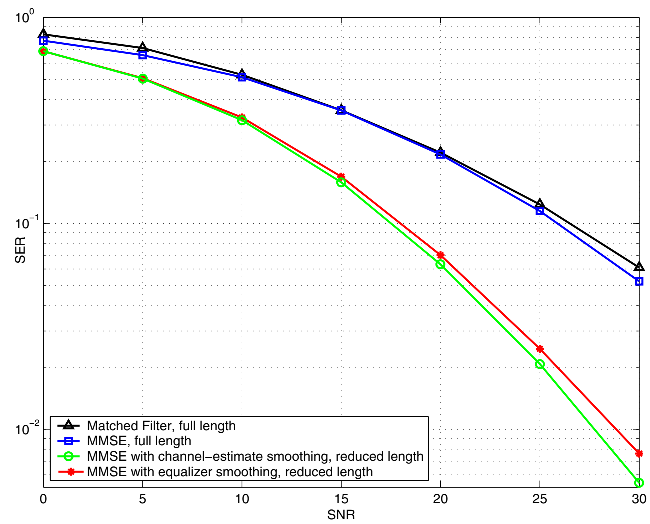

Fig. 2. Symbol Error Rate versus Signal to Noise Ratio for several equalization techniques and uncoded transmission.

At the transmitter, encoded bits are mapped into QPSK symbols and sent over the wireless channel. The spreading factor is set to 16. We implement a receiver equipped with a Zero Forcing equalizer and the following set of possible equalizer lengths: $L = \{12, 24, 48, 96, 192\}$. Hence, full length equalization corresponds to 192 chips, i.e. the midamble size. The equalized data symbols are despreaded (and unscrambled), demapped, deinterleaved, and decoded through a Viterbi decoder.

In order to carry out experiments, we consider an outdoor-to-indoor scenario, with the base station placed at the roof of a building, and the mobile inside the building with no line of sight. In this scenario the mobile receives data through a channel with delay spread of several symbol periods. In our experiments, given a certain base and mobile station setting, attenuation is added at the mobile receiver in order to decrease the average SNR and force the need for larger equalizer lengths.

B. Results

Fig. 3 shows the evolution of the average equalizer length as the SNR increases. In this scenario, the proposed adaptive complexity receiver reduces the necessary equalization length down to average values between 24 and 48 chips for relatively low SNR values (approximatealy greater than 8 dB). Hence, the equalization length can be reduced up to 8 times in average with respect to full length equalization (192 chips), for the SNR range considered. Table I provides an insight into the power savings provided by the proposed approach. The power savings correspond to the power reduction obtained by each of the different receiver configurations (for different equalizer lengths) with respect to full length equalization, taking into account all per-slot PHY-layer reception algorithms (except for synchronization). As Table I shows, equalizer lengths of 24 and 48 provide power savings of 20% and 25% respectively. Hence, the proposed adaptive receiver in this experimental

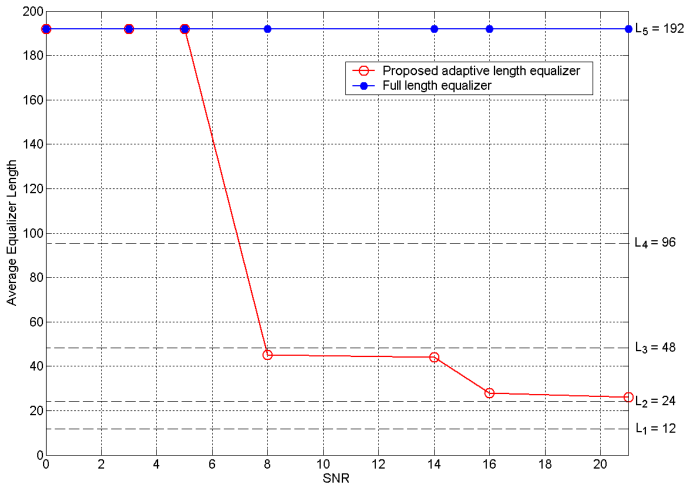

Fig. 3. Evolution of equalizer length as a function of average received SNR.

TABLE I

EXPERIMENTAL PER-SLOT EXECUTION TIMES AND POWER SAVINGS INVOLVING: EQUALIZATION, DESPREADING (AND UNSCRAMBLING), DEMAPPING, DEINTERLEAVING AND DECODING

	Equalizer Length	Execution Time			Average Power Savings
		Minimum (μs)	Average (μs)	Maximum (μs)	
L_1	12	120	120	126	29%
L_2	24	125	126	139	25%
L_3	48	135	135	141	20%
L_4	96	161	161	181	5%
L_5	192	168	169	197	0%

outdoor-to-indoor scenario provides power savings between 20% and 25% whenever the SNR is greater than 8 dB, which is a rather common situation.

On the other hand, experimental results have shown that for SNR values greater than 5dB the Block Error Rate (BLER) at the receiver is negligible both for the full length and adaptive length approaches. Hence, the system performance is preserved by the proposed scheme. For lower SNR values, both schemes converge to full length equalization, with an approximate BLER of 2% as the SNR approaches zero.

VI. CONCLUSIONS

An adaptive complexity equalizer has been presented, which adapts the equalizer length to the multipath scenario preserving the system performance. The receiver tracks the evolution of the equalizer taps through a low complexity algorithm. Experimental outdoor-to-indoor results show power savings of up to 25% with respect to full length equalization in a practical scheme. In indoor scenarios, milder conditions may allow for even further length reduction and hence power savings, thus extending terminal autonomy.

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