PIPELINED IMPLEMENTATION OF ADAPTIVE MULTIPLE-ANTENNA CDMA MOBILE RECEIVERS

Ramin Baghaie, Stefan Werner and Timo Laakso

Helsinki University of Technology
Laboratory of Telecommunications Technology
P.O. BOX 3000, 02015 HUT, Finland
ramin.baghaie@hut.fi

ABSTRACT

Pipelined implementation of an adaptive Direct-Sequence Code Division Multiple Access (DS-CDMA) receiver is proposed when multiple antennas are utilized for mobile communications. Adaptive multiple-antenna receivers can provide insensitivity to the interfering powers and room for more users or require smaller number of antennas than the matched filter solution.

In this paper, a number of approximation techniques are utilized to pipeline the adaptive algorithm used for the proposed multiple-antenna receiver. The resulting pipelined receiver requires minimal hardware increase and achieves a higher throughput or requires lower power as compared to the receiver using the serial algorithm. Simulation results illustrate the signal-to-interference (SIR) versus the relative interfering power for different number of antennas and different levels of pipelining.

1. INTRODUCTION

In the area of digital communications, there is a growing need for high speed circuits having a low power consumption. Two popular approaches for achieving high processing speed are pipelining and parallel processing. From a single-chip implementation point of view, the pipeline approach could be an advantage due to its lower hardware cost [1].

In general, algorithm transformation techniques have been employed to introduce concurrency in serial algorithms [1]-[3]. Pipelined DSP algorithms allow us to tradeoff speed, power and area during the course of VLSI implementation. Pipelining is simply accomplished by placing latches at any feed-forward cutsets of the data flow graph representation of the algorithm. However, pipelining of DSP algorithms having a feedback loop is not a trivial task [1].

Different algorithm transformation techniques such as the Look-Ahead (LA) and the Relaxed Look-Ahead (RLA) have already been proposed for the pipelining of recursive DSP algorithms [1]-[4].

These transformations introduce additional concurrency in a serial algorithm at the expense of hardware overhead. The look-ahead technique has been successfully applied to a number of such algorithms [1]. The LA technique, however, results in a large hardware overhead as it transforms a serial algorithm into an equivalent pipelined algorithm. This equivalency is in terms of the input-output behavior [1]-[3]. RLA technique involves approximating the algorithms obtained via the look-ahead technique. Through these approximations, the technique maintains functionality of the algorithm rather than the input-output behavior. A number of approximations such as sum relaxation and product relaxation are possible and each result in a different algorithm. Depending on the sophistication of the approximation there may or may not be a performance degradation. In the context of adaptive filtering, the approximation can be quite crude and yet result in minimal performance loss [1].

Unlike the LA technique, the application of the RLA technique modifies the original algorithm and therefore a convergence analysis is necessary. This could be considered as one drawback when using the relaxed look-ahead technique, since this analysis can be cumbersome. However, despite of this, in all cases, the resulting pipelined algorithm requires minimal hardware increase and achieves a higher throughput compared to the serial algorithm [1].

This increased of throughput as a result of pipelining can be exchanged for either reducing power or reducing area on the chip. Reducing power can be done in combination with power supply scaling [5,6]. Area reduction, however, can be achieved in combination with folding transformation [1]. Reductions in power or area is of great importance when implementing mobile communication systems.

This paper is organized as follows. In Section 2, multiple-antenna receivers are discussed. Section 3 deals with the pipelined implementation of the adaptive receiver. In Section 4, simulation results are reported. Finally, in the last section conclusions are given.

2. MULTIPLE-ANTENNA RECEIVER FOR CDMA MOBILE RECEPTION

In [7], a stochastic gradient algorithm was proposed which only requires knowledge of the desired user's spreading code. In [8,9], the idea in [7] was generalized by including multiple antennas and also employing adaptive algorithms.

The structure of the receiver equipped with N antennas is shown in Figure 1 [8,9]. Each of the N antenna branches contains a linear filter whose coefficients are to be optimized. The filtered signals from each antenna are then added together to form a decision variable.

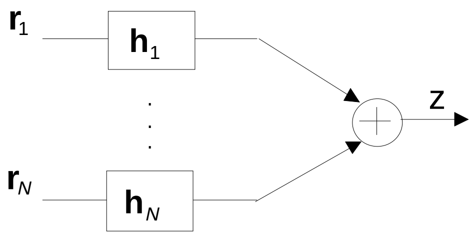

Figure 1. Structure of linear receiver

In Figure 1, $\mathbf{r}_i$ denotes the received signal after chip-matched filtering at antenna i, $\mathbf{h}_i$ contains the complex filter coefficients for the i^{th} antenna, and z is the decision variable formed by adding the filtered outputs from each antenna.

In order to get a compact notation, let us collect the filter coefficients and the received sequences from the antennas in vectors as

$$\mathbf{h} = \left[\mathbf{h}_1^{\mathrm{T}} \ldots \mathbf{h}_N^{\mathrm{T}}\right]^{\mathrm{T}} \tag{1}$$

$$\mathbf{r} = \left[\mathbf{r}_1^{\mathrm{T}}\ \mathbf{r}_2^{\mathrm{T}} \ldots \mathbf{r}_N^{\mathrm{T}}\right]^{\mathrm{T}} \tag{2}$$

Using the above notation, the output from the receiver can be written as

$$z = \mathbf{h}^{\mathrm{H}}\mathbf{r} \tag{3}$$

The variance of the output, i.e., the output power, is

$$E\left\{|z|^2\right\} = E\left\{\mathbf{h}^{\mathrm{H}}\mathbf{r}\mathbf{r}^{\mathrm{H}}\mathbf{h}\right\} = \mathbf{h}^{\mathrm{H}}\mathbf{R}\mathbf{h} \tag{4}$$

where $\mathbf{R}$ is the correlation matrix with elements

$$\mathbf{R}_{ij} = E\left\{\mathbf{r}_i\mathbf{r}_j^{\mathrm{H}}\right\} = \mathbf{S}_i E\left\{\mathbf{A}_i\mathbf{A}_j^{\mathrm{H}}\right\}\mathbf{S}_j^{\mathrm{H}} + \sigma_i \mathbf{I}_G \delta(i-j) \tag{5}$$

Let us now state the optimization problem as follows. We want to find the filter $\mathbf{h}$ such that the output variance of Equation (4) is minimized under the constraints that the desired user's code sequence in every antenna can pass undistorted. By introducing the $GN\times N$ matrix $\mathbf{C}$ and the $N\times 1$ vector $\mathbf{u}$ as

$$\mathbf{C} = \begin{bmatrix} a_1\mathbf{s}_{1,1} & 0 & \cdots & 0 \\ 0 & a_2\mathbf{s}_{1,2} & 0 & \vdots \\ \vdots & 0 & \ddots & 0 \\ 0 & \cdots & 0 & a_N\mathbf{s}_{1,N} \end{bmatrix} \tag{6}$$

$$\mathbf{u} = \left[|a_1|^2\ |a_2|^2 \ldots |a_N|^2\right]^{\mathrm{T}} \tag{7}$$

where $\mathbf{s}_{1,i}$ is the code sequence and a_i is the complex phase factor of the desired user at the antenna element, the minimization problem can now be formulated as

$$\hat{\mathbf{h}} = \arg\min_{\mathbf{h}} E\left\{|z|^2\right\} \tag{8}$$

$$\text{subject to:} \quad \mathbf{C}^{\mathrm{H}}\mathbf{h} = \mathbf{u}$$

The formulation in Equation (8) is general in that sense that, if the interference environment changes Equation (8) remains the same. The solution to this problem is found by the method of Lagrange multipliers, see, e.g., [10]

$$\mathbf{h}_{\text{opt}} = \mathbf{R}^{-1}\mathbf{C}\left[\mathbf{C}^{\mathrm{H}}\mathbf{R}^{-1}\mathbf{C}\right]^{-1}\mathbf{u} \tag{9}$$

The minimum output variance is obtained by substituting (9) into (4):

$$E\left\{|z|^2\right\} = \mathbf{u}^{\mathrm{H}}\left[\mathbf{C}^{\mathrm{H}}\mathbf{R}^{-1}\mathbf{C}\right]^{-1}\mathbf{u} \tag{10}$$

The closed-form solution is not suitable for practice, as we need to estimate the correlation matrix and perform an inversion. Thus, an adaptive implementation of the detector is considered. In [9], the use of the Frost algorithm [12] was proposed.

We will here use the structure of the generalized sidelobe canceler (GSC) [10,14], which transforms a constrained problem into an unconstrained problem by means of an orthogonal decomposition of $\mathbf{h}$. The main reason for doing this is that simpler algorithms can be applied. The idea is to divide the weight vector $\mathbf{h}$ into two parts as

$$\mathbf{h} = \mathbf{h}_q - \mathbf{C}_a\mathbf{h}_a \tag{11}$$

where $\mathbf{h}_q$ is a fixed vector satisfying the constraint equations, $\mathbf{C}_a$ is a $GN\times(GN\text{-}N)$ matrix which is the orthogonal complement to the constraint matrix, i.e., $\mathbf{C}_a^{\mathbf{H}}\mathbf{C}=\mathbf{0}$, and $\mathbf{h}_a$ is a $(GN\text{-}N)\times 1$ vector unaffected by the constraints and freely to adapt.

By choosing $\mathbf{h}_q = \mathbf{C}(\mathbf{C}^{\mathbf{H}}\mathbf{C})^{-1}\mathbf{u}$ and defining $\mathbf{x} = \mathbf{C}_a^{\mathbf{H}}\mathrm{r}$ and $d=\mathbf{h}_q^{\mathbf{H}}\mathrm{r}$, we can apply the LMS adaptive implementation for the update of the vector $\mathbf{h}_a$ [10].

3. PIPELINED IMPLEMENTATION OF THE ADAPTIVE RECEIVER

Consider the LMS algorithm of Equations (12) and (13) [10,11]:

$$\mathbf{h}_a(k+1) = \mathbf{h}_a(k) + \mu\mathbf{x}(k)z^*(k) \tag{12}$$

$$z(k) = d(k) - \mathbf{h}_a^{\mathrm{H}}(k-1)\mathbf{r}(k) \tag{13}$$

where μ is the step size parameter.

By applying the M-step look-ahead to Eq. (12) we have:

$$\mathbf{h}_a(k) = \mathbf{h}_a(k-M) + \mu\sum_{i=0}^{M-1}\mathbf{x}(k-i)z^*(k-i) \tag{14}$$

For $M = 1$, Eq. (14) represents the serial LMS algorithm.

Substituting Eq. (14) in Eq. (13) leads to:

$$z(k) = d(k) - \Big[\mathbf{h}_a^{\mathrm{H}}(k-M-1) + \mu\sum_{i=0}^{M-1}\mathbf{x}^{\mathrm{H}}(k-i-1)z^*(k-i-1)\Big]\mathbf{x}(k) \tag{15}$$

The above technique, however, results in a large hardware overhead since it transforms the serial LMS algorithm into an equivalent pipelined algorithm. This equivalency is in terms of the input-output behavior. It is obvious that the above hardware overhead can not be tolerated specially for large M when implementation of mobile receivers are of interest. Thus, some approximations should be utilized.

Assuming that μ is sufficiently small, the third term on the right hand side of Eq. (15) can be approximated as zero. Finally, by replacing $\mathbf{h}_a^{\mathrm{H}}(k-M-1)$ by $\mathbf{h}_a^{\mathrm{H}}(k-M)$ [1], Equation (15) can be approximated as:

$$z(k) = d(k) - \mathbf{h}_a^{\mathrm{H}}(k-M)\mathbf{x}(k) \tag{16}$$

Equations (14) and (16) describe the pipelined-LMS algorithm. In Section 4, with the aid of simulations it is demonstrated that these approximations are reasonable. Through these approximations, the functionality of the algorithm has been maintained. However, the input-output behavior of the Equations (12) and (13) has been altered. One could still apply more approximation to Eq. (14) by using techniques such as the sum and delay relaxation and reduce the hardware overhead of Equation (14) [1].

As a result of these approximations and relaxation techniques, the convergence condition should be checked. This problem has been addressed in [1], and it is shown that the upper bound on μ to guarantee the convergence is found to be tighter than that of the serial LMS algorithm.

As a result of these approximations, the performance may degrade. Usually, for non-stationary signals, this would mean slight increase in the mean-squared error and slower convergence speed. In the context of our application, in Section 4, the simulations results illustrate that for a moderate M, these approximations result in minimal performance loss. As it can been seen from Equation (14), by applying the LA technique, M delays have been introduced in the recursive loops, which must be redistributed to pipeline the multiply-add operation.

By proper distribution of these extra delays, the pipelined architecture will operate M times faster [1]. This increase, however, could be traded for either reducing power or reducing the chip area [1].

4. SIMULATION RESULTS

A number of simulations have been conducted to compare the performance of the serial LMS and the pipelined LMS algorithms for different speedup factors M and number of antennas N.

In these simulations, antennas are structured as a uniform linear array (ULA) with half the wavelength spacing. The direction of arrival is set to 15°. The spreading sequences are Gold codes of length 7 [13]. The system consists of five users. The signal-to-noise ratio at the antennas for the desired user is fixed to 8 dB.

Figures 2 and 3 illustrate the average signal-to-interference ratio (SIR) as a function of the relative power of the interfering users when using one and two antennas respectively. In these simulations, the interfering power of all users varies from 0 to 10 dB.

In these figures, as M -i.e., the number of pipelining stages- increases, the SIR will decrease. This is due to the higher misadjustment as a result of the approximations.

By comparing Figures 2 and 3, one could notice that the SIR loss for the two-antenna case when introducing pipelining is more as compared to the single antenna case. For example, for the relative interference power of 10 dB and M = 10, the SIR loss for a two-antenna receiver was 5.6 dB. However, in the single-antenna case it was only 4.3 dB.

Further more, we can observe that the level of pipelining M should be carefully selected when more antennas are introduced. As an example, consider again the case where the relative interference power was 10 dB. For a single-antenna receiver with M=5, or with two antennas having M=10, roughly the same SIR can be achieved.

Figure 4 shows the SIR as a function of the number of the iterations for different levels of pipelining. The convergence speed is seen to be dependent on the level of pipelining as expected.

5. CONCLUSIONS

In this paper, pipelined implementation of a DS-CDMA receiver was proposed when multiple antennas are utilized in mobile receivers. A number of approximation techniques were utilized to introduce pipelining and achieve a higher throughput as compared to the receiver using the serial algorithm.

This increase, however, can be traded for either reducing power or reducing the chip area. Simulations were carried out to illustrate the SIR versus the relative interfering power for different number of antennas and different levels of pipelining. Also, the convergence speed for different levels of pipelining was compared. It is important to note that in the multiple-antenna case, the drop of SIR versus the relative interference power is more for a large M as compared to receivers using only one antenna.

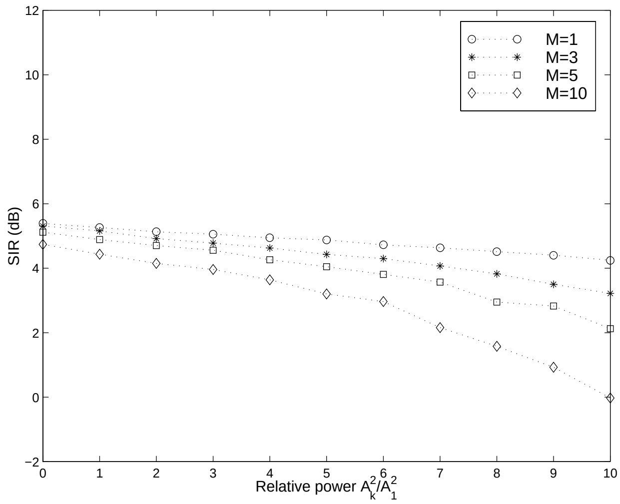

Figure 2. SIR versus the relative powers of the interfering users when using one antenna (N=1) for M=1, 3, 5, and 10.

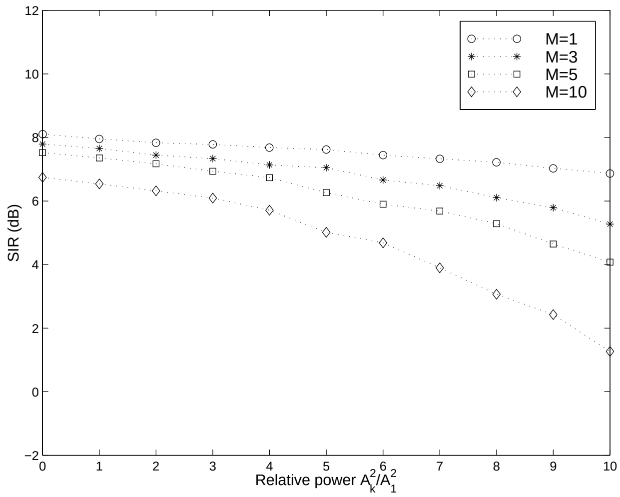

Figure 3. SIR versus the relative powers of the interfering users when using two antennas (*N*=2) for *M*=1, 3, 5, and 10.

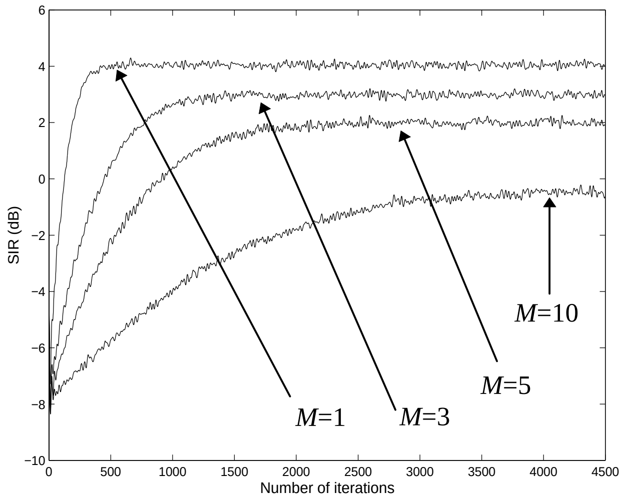

Figure 4. SIR as a function of the number of iterations (300 runs smoothed) when using one antenna and *M*=1, 3, 5, and 10.

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