PERFORMANCE OF MULTIPLE ANTENNA COMBINING ALGORITHMS IN THE UPLINK OF UTRA/FDD

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***Abstract* – The aim of our work was to rate the suitability of different array combining algorithms to the special requirements of the UTRA/FDD uplink. After formulating the algorithms, we assess their bit error performance in a modified vehicular B channel that is spatially inhomogeneous. The bit error ratios in different situations are presented together with some considerations about the computational complexity of the used methods. From bit error performance and required effort, we conclude that the minimum mean squared error (MMSE) algorithm with reduced complexity is the most suitable method for a real operating system.**

I. INTRODUCTION

The dominant limiting factor of transmission quality in cellular CDMA-systems is co-channel interference caused by echoes and multiple access. Therefore, every receiver architecture that is able to effectively suppress these spurious contributions directly enhances capacity. A temporal RAKE receiver remedies the effect of multipath propagation since it is able to sum up the echoes coherently. Additionally using smart antennas at base stations exploits the directional nature of the mobile radio channel against multiple access interference. In both cases, a combining algorithm is needed to compute weights to properly merge the different multipath components collected at multiple antenna elements to a single decision variable.

Generally, two different strategies for weight computation can be distinguished. Maximising the signal to noise ratio at the decision device (maximum ratio combining) is optimal in scenarios where interference can be modelled as being spatially and temporally white. On the other hand, optimum combining also accounts for spatially coloured interference by maximising the signal to noise and interference ratio.

The aim of this paper is to investigate various smart antenna combining algorithms with respect to performance and computational complexity in different propagation scenarios.

II. ALGORITHMS

We consider combining algorithms in the uplink of the UTRA/FDD (UMTS terrestrial radio access / frequency division duplex) mode [1], where we transmit one data channel on the I-branch and the control channel containing the pilot bits on the Q-branch. These symbol streams are denoted as s_I and s_Q. Together with the spreading codes c_I and c_Q, the complex valued scrambling code c, and the amplitudes A_I and A_Q, the transmitted signal of the desired user is given as

$$d(n) = \left(A_I \, s_I(k_I) c_I(n) + \mathrm{j} \cdot A_Q \, s_Q(k_Q) c_Q(n) \right) \cdot c(n) \,, \tag{1}$$

where n and $k_{I/Q}$ denote the chip and symbol index, respectively. We used a spreading factor of SF_I=128 for data and SF_Q=256 for control information, corresponding to 20 and 10 bits per slot, respectively. The complex valued scrambling code has always the fixed length of 256 chips.

In our single-user detection case, the equivalent base band signal received at the M antenna elements is a discrete time convolution sum of the transmitted data with the channel impulse response

$$\mathbf{h}(n) = \sum_{l=0}^{L-1} \mathbf{a}_l \cdot \delta(n - \tau_l) \,. \tag{2}$$

It consists of L dominant paths, where the integer τ_l denotes the delay of the l-th path in multiples of the chip duration. The Mx1vector $\mathbf{a}_l$ represents the array response vector of the l-th path including path loss, fading, and pulse shaping. The whole received signal including spatially and temporally white Gaussian noise $\mathbf{n}$, and interference $\mathbf{s}_{INT}$ is then given as

$$\mathbf{x}(n) = \sum_{l=0}^{L-1} d(n - \tau_l)\mathbf{h}(\tau_l) + \mathbf{s}_{INT}(n) + \mathbf{n}(n) \,. \tag{3}$$

In the despreading process we can distinguish between the signals of different paths by choosing an according offset. Correlating $\mathbf{x}(n)$ with the spreading codes yields the signals $\mathbf{y}_{I,l}(k_I)$ and $\mathbf{y}_{Q,l}(k_Q)$, where the index l indicates the dependency on the l-th path and the symbol indices k_I and k_Q represent the different bit rates. With these quantities we are able to define the covariance matrices

$$\mathbf{R}_{\mathbf{xx}} = \mathrm{E}\{\mathbf{x}\mathbf{x}^H\}, \tag{4}$$

$$\mathbf{R}_{\mathbf{yy},l} = \mathrm{E}\{(\mathbf{y}_{I,l} + \mathbf{y}_{Q,l})(\mathbf{y}_{I,l} + \mathbf{y}_{Q,l})^{\mathrm{H}}\}, \tag{5}$$

that we will need in the sequel. In practical implementations, the expectation value $\mathrm{E}\{\cdot\}$ has to be approximated by the time average.

The task of the combining algorithms is now to find proper weights to combine all the multipath components impinging on multiple antenna elements to one decision variable.

These weight computation algorithms can be roughly divided into two classes. Maximum ratio combining (MRC) algorithms use an estimate of the channel impulse response as weight vector and are optimal only in the presence of spatially and temporally white noise. If, on the other hand, significant coloured interference is present, we have to use interference suppression methods to achieve optimal performance. Here, the impulse response vector is pre-multiplied by the inverse of a suitably chosen covariance matrix to decrease the influence of interference.

A. Pilot Based MRC

Since UTRA employs pilot symbols in the control channel, the simplest way to obtain an estimate of the channel impulse response is to correlate the de-spread control information with the known pilot symbols [2]

$$\mathbf{w}_l = \mathrm{E}\{\mathbf{y}_{Q,l} \cdot s_{Q,pilot}\}. \tag{6}$$

As mentioned above, we will approximate the expectation value by a sample mean. This robust and simple way to obtain the $\mathbf{w}_l$ inherently yields the absolute channel phase, enabling coherent detection.

B. Principal Components (PC) MRC

In the pilot based method, antenna weights are computed separately and independently of each other, while the averaging process is only extended over the number of pilot symbols in the control channel. Possibly, principal components methods offer an improvement in performance since they are able to use *all* symbols of data *and* control channel and consider the correlations between the antenna elements. On the other hand, the required eigenvalue decomposition will make principal components methods more sensitive to noise.

With (3), (4) and under the assumptions [3] that noise is temporally and spatially white, the chip sequence of the desired user is white, the transmitted signal is independent of noise and interference, and all channels are LTI with a finite duration, we can write the chip-covariance matrix as

$$\mathbf{R}_{\mathbf{xx}} = 2(A_I^2 + A_Q^2)\sum_{l=0}^{L-1}\mathbf{a}_l\mathbf{a}_l^H + \mathbf{R}_{\mathbf{nn}}. \tag{7}$$

The sum contains all the multipath components of the desired signal, whereas $\mathbf{R}_{\mathbf{nn}}$ denotes the noise and interference covariance matrix.

If we additionally assume that noise and interference in $\mathbf{y}_{I,l}$ and $\mathbf{y}_{Q,l}$ are mutually independent, and that the code chips are binary i.i.d. random variables, we can write the de-spread covariance matrix as

$$\mathbf{R}_{\mathbf{yy},l} = 2(A_I^2 + A_Q^2)\mathbf{a}_l\mathbf{a}_l^H + \frac{1}{C}\left(\sum_{\substack{k=0,\\ k\neq l}}^{L-1} 2(A_I^2 + A_Q^2)\mathbf{a}_k\mathbf{a}_k^H + \mathbf{R}_{\mathbf{nn}}\right) \tag{8}$$

where the constant

$$C = \frac{SF_I \cdot SF_Q}{SF_I + SF_Q} \tag{9}$$

describes the attenuation of noise, interference and multipath components. For the considered spreading factors we can expect that this attenuation is strong enough to let the matrix $\mathbf{R}_{\mathbf{yy},l}$ be dominated by the term $\mathbf{a}_l\mathbf{a}_l^H$. Thus, the eigenvector belonging to the dominant eigenvalue of $\mathbf{R}_{\mathbf{yy},l}$ provides an estimate for the desired array response vector of the l-th path [2]. But to obtain the absolute channel phase for coherent detection, we have to additionally correlate the eigenvector with the known pilot symbols.

In the presence of spatially coloured interference we could also take the dominant eigenvector of the matrix $C\cdot \mathbf{R}_{\mathbf{yy},l} - \mathbf{R}_{\mathbf{xx}}$ [2] as a weight vector, which is a better estimate of the channel impulse response (coloured noise (cn) principal components). But even this approach cannot totally remedy the inherent sub-optimality of MRC algorithms in the presence of coloured interference.

C. Minimum Mean Squared Error (MMSE)

The Wiener filtering interference suppression approach minimises the mean squared error between the symbol decision and the transmitted data, and is hence superior to MRC in the presence of coloured interference. The desired weight vector is given as [4]

$$\mathbf{w}_l = \mathbf{R}_{\mathbf{xx}}^{-1} \cdot \hat{\mathbf{a}}_l, \tag{10}$$

where $\hat{\mathbf{a}}_l$ is the estimated array response vector. The pre-multiplication by the inverse of the chip covariance matrix suppresses coloured noise and interference. To estimate $\mathbf{R}_{\mathbf{xx}}$ accurately enough, it might not be necessary to use all chips for computing the sample mean. By taking just a fraction of the available chips, we can reduce the computational complexity significantly (reduced complexity MMSE).

D. Optimum Combining (OC)

A second interference suppression algorithm is called optimum combining [5]. It maximises the signal to noise *and* interference ratio (SNIR) at the combiner output. With (7) and (8), we find that

$$\mathbf{R_{xx}} - \mathbf{R}_{\mathbf{yy},l} = \left(1 - \frac{1}{C}\right)\left(\sum_{\substack{k=0,\\ k\neq l}}^{L-1} 2(A_I^2 + A_Q^2)\mathbf{a}_k \mathbf{a}_k^H + \mathbf{R_{nn}}\right) \tag{11}$$

contains only the unwanted multipath components, noise, and interference. The weights given as

$$\mathbf{w}_l = \left(\mathbf{R_{xx}} - \mathbf{R}_{\mathbf{yy},l}\right)^{-1} \cdot \hat{\mathbf{a}}_l\,, \tag{12}$$

thus optimally enhance the SNIR [5].

III. SIMULATIONS

To evaluate the performance of the presented methods in various scenarios we used a simulation environment according to the standard of UTRA/FDD uplink [1].

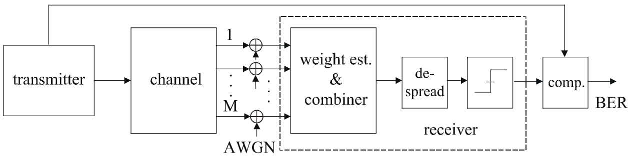

Figure 1: Simulation environment

The transmitter generates data channel and control channel from random data and known pilot symbols, performs spreading, scrambling, and pulse-shaping.

The signal is then fed into the spatial channel, which is modelled as a separate vehicular B [6] tapped delay line from the transmitter to every receiving antenna. Each delay tap represents a distinct path with independent fading. In the spatial domain, we can select two different fading scenarios. Independent fading among the antennas models diversity reception, whereas spatially correlated fading corresponds to closely spaced antennas. In this case the Rayleigh-coefficients at the antennas differ only by a constant phase factor determined by the path's direction of arrival (DOA). Interference is modelled as an additional mobile station with constant DOA and variable power. It uses a different scrambling code and all paths have a constant angle offset and a random time offset. After summing up all paths, independent white Gaussian noise is added at each antenna element.

After pulse-shaping at the receiver, the multiple antenna signals are fed into the three blocks shown in Fig. 2.

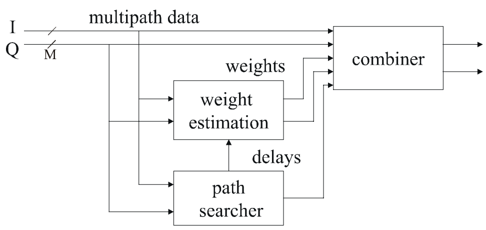

Figure 2: Weight estimation, path searcher, and combiner

The *path searcher* uses all incoming antenna signals to identify the delay times of the dominating paths and delivers them to weight estimator and combiner. Using the methods described in Section II, the *weight estimator* computes a weight vector for every detected path. Finally, the *space-time combiner* multiplies the incoming antenna signals with the complex conjugated weight vector, removes the delay time differences and adds up all contributions to obtain a single data stream for succeeding despreading. After passing the decision device, the received data is compared to the transmitted to obtain the bit error ratio.

Fig. 3 shows results with variable white noise and an interferer having the constant power of 32 times the desired user (SF_{int}=4). Fading is un-correlated among the six antennas, mobile velocity is set to 50km/h and we simulated 10^6 data symbols.

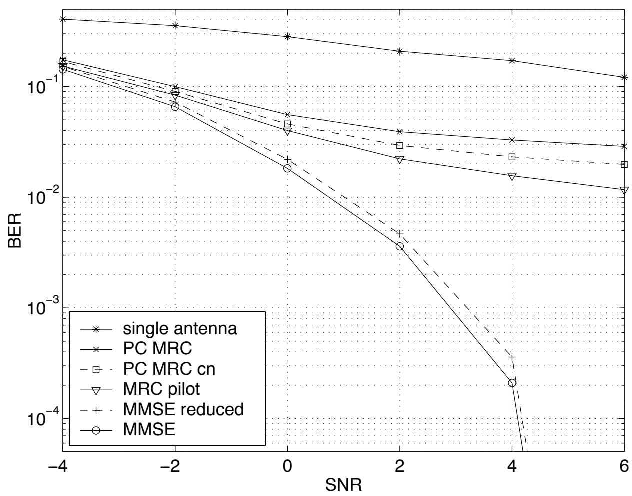

Figure 3: White noise and one interferer with SF_{int}=4

In the region of low SNR, noise is the dominating source of error, hence MRC and interference suppression algorithms perform nearly equally. Above about -2dB SNR (after despreading), the BER of MRC methods saturates. Both MMSE algorithms still show a performance gain and just need about 3dB SNR for a BER of 0.1%.

To keep Fig. 3 more readable, we neglected optimum combining. As you can see in Fig. 4 (correlated fading, variable powered interference, no noise),

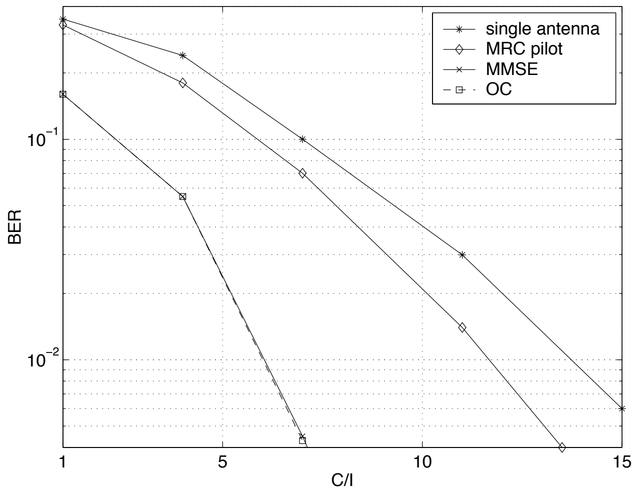

Figure 4: Variable interferer power and negligible noise

the increased computational effort of optimum combining does not lead to a performance gain against MMSE.

A second consequence of Fig. 4 is that MRC is very sensitive to strong interference. This reflects the fact that an MRC algorithm is optimal for spatially white noise only. Correlated fading additionally reduces the gain of MRC against the single antenna receiver in this scenario.

IV. COMPUTATIONAL COMPLEXITY

Besides the BER performance, the required computational effort of the algorithms is important for practical implementation. TABLE I shows the number of floating point operations[1] per slot for the parameter selection and weight computation methods of Fig. 3.

Table I: Required computations for used algorithms

algorithm	**operations**
single antenna	430
MRC with pilot	2600
reduced complexity MMSE	$20 \cdot 10^3$
MMSE	$375 \cdot 10^3$
OC	$113 \cdot 10^4$
principal components MRC	$114 \cdot 10^4$
principal components MRC (cn)	$151 \cdot 10^4$

As input for the different algorithms we assumed the de-spread control channel, the spread data stream (chips), the estimated multipath delays, and the wanted user's codes. All the other quantities had to be derived from this input data. Therefore, the list is dependent on the actual implementation of the receiver and you should understand it as a rather coarse classification of the presented algorithms.

Pilot based maximum ratio combining requires about six times as many operations as the single antenna receiver, while all the other methods are clearly more complex (~factor 2000).

The dominant source of effort is neither eigenvalue decomposition nor matrix inversion, but all processing that has to be performed on chip-level. For instance, the creation of the chip covariance matrix is the most complex part of the MMSE algorithm, since we have to average over 2560 data vectors. Here we can save nearly a factor of 20 by just using every 32nd sample for averaging (reduced complexity MMSE). By far the most expensive methods are OC and all principal components algorithms, because the data channel has to be de-spread to obtain $\mathbf{R}_{yy,l}$. The coloured noise (cn) principal components algorithm additionally needs the chip-matrix, making it the most complex of all considered methods.

[1] One floating point operation is either addition, multiplication, subtraction or division.

V. CONCLUSIONS

In simulations, we find the MMSE algorithm performing best with strong interference and just like the other methods with spatially white noise. Its computational complexity is higher by more than a factor of 100 compared to the pilot based MRC algorithm, but if we use MMSE with reduced complexity, we can save nearly a factor of 20 in effort with a performance loss of only about 0.25dB. With our parameter selection, optimum combining and principal components MRC do not provide sufficient performance benefits to justify their increased complexity

Therefore, we conclude that the reduced complexity MMSE algorithm is the most suitable one for practical implementation. If however, the system is not able to handle its computational demands, we suggest to use the simple and robust pilot based MRC algorithm.

VI. ACKNOWLEDGEMENT

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VII. REFERENCES

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