Space-Time Combining in the Uplink of UTRA/FDD

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***Abstract* - We investigate various array combining algorithms considering the special requirements of the UTRA/FDD uplink. After formulating the algorithms, we assess their 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

Co-channel interference in cellular CDMA-systems is the main limiting factor of transmission quality and capacity. The origin of this spurious power is mainly twofold. Echoes received via multiple paths with distinct propagation delays lead to inter-chip interference (ICI). Secondly, incompletely vanishing cross-correlation between different user's spreading-codes leads to unwanted contributions, called multiple access interference (MAI). Employing a temporal RAKE receiver remedies ICI, since multipath components can be separated and coherently summed up to gain maximum signal power. If we additionally use smart antennas at base stations, it becomes possible to exploit the directional nature of the mobile radio channel against MAI. By coherently combining all the echoes at different antenna outputs, we are able to accumulate received signal power, while simultaneously suppressing interference. The aim of this paper is to rate the performance and computational effort of various combining algorithms in different propagation scenarios.

II. SYSTEM LAYOUT

For bit error performance assessment, we used a baseband link-level simulation chain without error correction coding and power control. The implementation is according to the W-CDMA standard of the UTRA/FDD uplink [1]. An existing single antenna simulation chain was equipped with a spatial channel model and up to six antennas.

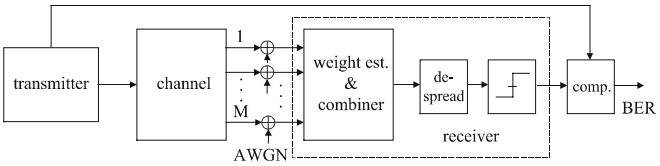

Fig. 1: Simulation environment

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

The signal is then fed into the spatial channel, which is modeled as a separate vehicular B [2] 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 modeled 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

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paths, independent white Gaussian noise is added at each antenna element.

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

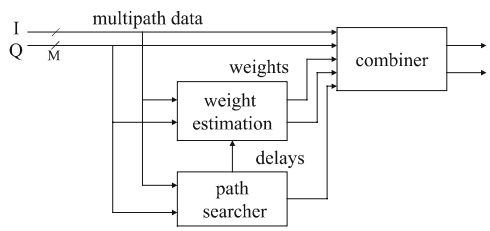

Fig. 2: Weight estimation, path searcher, and combiner

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 4, 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.

III. SIGNAL MODEL

We transmit one data channel (DPDCH) on the I-branch and the control channel (DPCCH) on the Q-branch. These symbol streams are denoted as s_I and s_Q. Together with the channelization 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

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

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 base band signal received at the M antenna elements is a discrete time convolution sum of the transmitted data with the channel impulse response

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

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

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

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

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

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

that we will need in the following section. In practical implementations, the expectation value has to be approximated by the time average.

IV. ALGORITHMS

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 colored 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 DPCCH, 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 [3]

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

As mentioned above, we will approximate the expectation value by a sample mean over one slot. This robust

and simple way to obtain the $\mathbf{w}_l$ inherently yields the absolute channel phase, enabling coherent detection. Averaging data over more than one slot is beneficial only if the channel stays constant during this time.

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 [4] 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 presume 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 despread 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 that the matrix $\mathbf{R}_{\mathbf{yy},l}$ is 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 [3]. 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 colored interference we could also take the dominant eigenvector of the matrix $C\cdot \mathbf{R}_{\mathbf{yy},l} - \mathbf{R}_{\mathbf{xx}}$ [3] as a weight vector, which is a better estimate of the channel impulse response (colored noise (cn) principal components). But even this approach cannot totally remedy the inherent sub-optimality of MRC algorithms in the presence of colored interference.

C. Minimum Mean Squared Error (MMSE)

The Wiener filtering interference suppression approach minimizes the mean squared error between the symbol decision and the transmitted data and is hence superior to MRC in the presence of colored interference. The desired weight vector is given as [5]

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

where $\hat{\mathbf{a}}_l$ is the estimated array response vector. The pre-multiplication by the inverse of the chip covariance matrix suppresses colored 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 [6]. It maximizes the signal to noise *and* interference ratio (SNIR) at the combiner output. With (7) and (8), we find that

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

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

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

thus optimally enhance the SNIR [6].

E. 2D Algorithms

Up to now, we computed the weights separately for each delay. Although the correlation receiver is able to separate the different paths, we may nevertheless expect

an additional benefit by considering all multipath components jointly. Using the vectors [4]

$$\mathbf{x}_L(n) = \begin{bmatrix} \mathbf{x}^T(n+\tau_0) & \mathbf{x}^T(n+\tau_1) & \cdots & \mathbf{x}^T(n+\tau_{L-1}) \end{bmatrix}^T,$$
$$\mathbf{y}_{I,L}(k_I) = \begin{bmatrix} \mathbf{y}_{I,0}^T(k_I) & \mathbf{y}_{I,1}^T(k_I) & \cdots & \mathbf{y}_{I,L-1}^T(k_I) \end{bmatrix}^T,$$
$$\mathbf{y}_{Q,L}(k_Q) = \begin{bmatrix} \mathbf{y}_{Q,0}^T(k_Q) & \mathbf{y}_{Q,1}^T(k_Q) & \cdots & \mathbf{y}_{Q,L-1}^T(k_Q) \end{bmatrix}^T, \tag{13}$$

to create the above mentioned covariance matrices, we can employ the same weight computation algorithms as in the sequential space-time case.

V. SIMULATIONS

In simulations with various parameter settings we assessed the performance of the presented algorithms. As a reference, we further included a single antenna receiver. Fig. 3 shows results with variable white noise and one intra-cell interferer with spreading factor SF_{int}=16. To compensate for the lack in spreading gain, the interferer is transmitting with a constant power of 8 times the desired user. Fading was uncorrelated among the six antennas, mobile velocity was set to 50km/h and we simulated 10^6 data symbols.

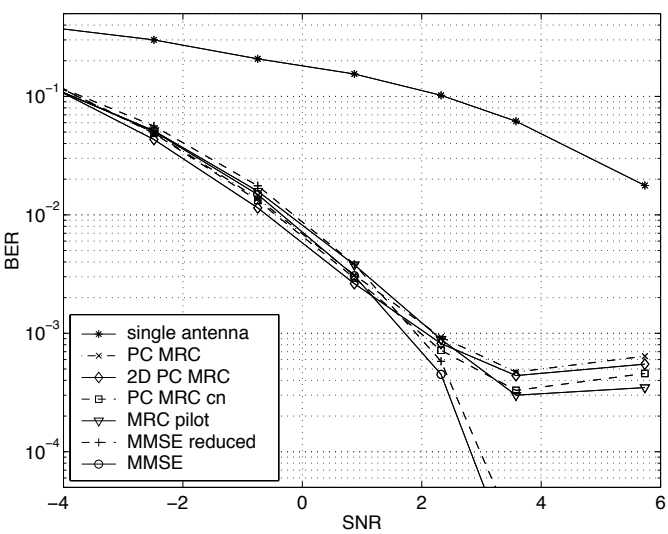

Fig. 3: White noise and one interferer with SF_{int}=16

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 influence of interference becomes obvious. The BER of MRC methods saturates, while interference suppression methods still offer improvement as noise goes down. In Fig. 4, the advantage of interference suppression algorithms becomes even more clear, as a 32 times stronger interferer is present (SF_{int}=4), while all other parameters are the same as in Fig. 3. Saturation of the MRC algorithms starts at about –2dB SNR and they can hardly obtain a BER below 1%. Both MMSE algorithms loose about 1.5dB compared to Fig. 3, but they still just need about 3dB SNR for a BER of 0.1%.

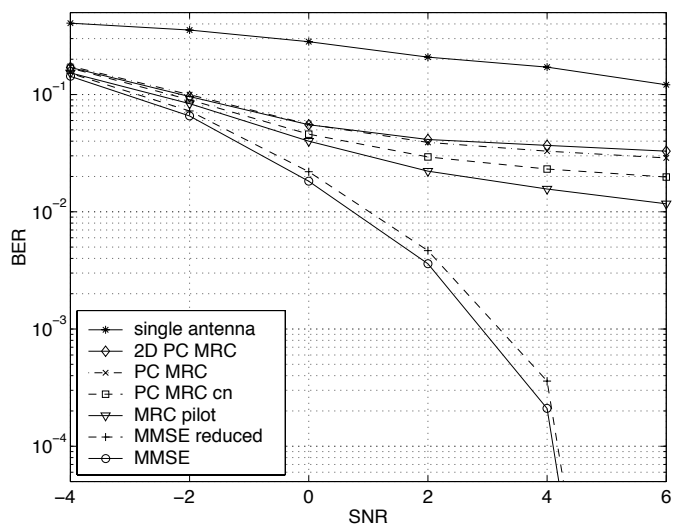

Fig. 4: White noise and one interferer with SF_{int}=4

To keep Fig. 3 and 4 more readable, we presented only one 2D algorithm and neglected optimum combining. As you can see in Fig. 5 (correlated fading, variable powered interference, no noise, mobile speed 50km/h), the increased computational effort of optimum combining does not lead to a performance gain against MMSE.

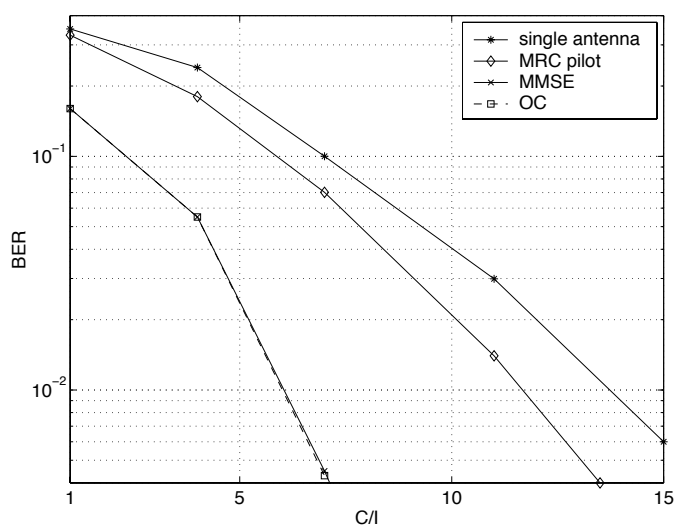

Fig. 5: Variable interferer power and negligible noise

A second consequence of Fig. 5 is that MRC is very sensitive to strong interference. This reflects the fact that an MRC algorithm is designed for spatially white noise. Correlated fading additionally reduces the gain of MRC against the single antenna receiver in this scenario.

From Fig. 3, 4, and simulations including other 2D methods we saw that 2D algorithms generally show no significant performance improvement. This may be due to the high spreading factors we used, enabling good suppression of multi-path components.

VI. 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 and 4.

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	$114 \cdot 10^4$
2D principal components MRC	$146 \cdot 10^4$
principal components MRC (cn)	$151 \cdot 10^4$

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. 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 $\mathbf{R}_{\mathbf{xx}}$ 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 all principal components algorithms, because the data channel has to be despread to obtain $\mathbf{R}_{\mathbf{yy},l}$. The colored 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.

VII. 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 more than a factor of 10 in effort with a performance loss of only about 0.25dB. With our parameter selection, optimum combining, principal components MRC and all 2D algorithms 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.

ACKNOWLEDGMENT

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REFERENCES

[1] 3GPP, "UE radio transmission and reception (FDD) – 3G TS 25.101 version 3.0.0," http://www.3gpp.org, October 1999.

[2] ETSI, "Selection procedures for the choice of radio transmission technologies of the UMTS – TR 101 112 version 3.2.0," Sophia Antipolis, France, April 1998.

[3] J. S. Thompson, P. M. Grant, B. Mulgrew, "Performance of antenna array receiver algorithms for CDMA," Signal Processing, vol. 68, no. 1, pp. 23-41, July 1998.

[4] H. Liu, M. D. Zoltowski, "Blind equalization in antenna array CDMA systems," IEEE Transactions on Signal Processing, vol. 45, no. 1, pp. 161-172, January 1997.

[5] S. Haykin, "Adaptive filter theory," Prentice-Hall, 590 p., 1986.

[6] J. H. Winters, "Optimum combining in digital mobile radio with cochannel interference," IEEE Journal on Selected Areas in Communications, vol. 2, no. 4, pp. 528-539, July 1984.