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Adaptive antennas exploit the inherent spatial diversity of the mobile radio channel and are, therefore, seen as an important technology to meet the high spectral efficiency and quality requirements of third generation mobile radio systems. The proposed concepts for the third generation [4] allow an easy and flexible implementation of new and more sophisticated services. Recently, ETSI SMG adopted the TD-CDMA concept as the radio access scheme for time-division duplex systems and the WCDMA concept for frequency-division duplex systems. This solution has been contributed to the International Telecommunication Union - as the European proposal for IMT-2000 transmision technology.

The main advantage of the conventional space-time rake [5] for single-user data detection is its simplicity. However, since co-channel interferers are not taken into account, the conventional space-time rake is not near-far resistant. A space-frequency rake [6, 1, 2] based on space-frequency covariance matrices combines each multipath across space and frequency, performs multi-user interference suppression in addition to diversity combining, and, therefore, mitigates the stringent power control requirements common to DS-CDMA systems. Hence, a direct capacity increase is achieved by transmitting more codes with lower spreading factors. Due to the reduced transmission rate for uplink power control due to mitigated power control requirements, the downlink capacity for data transmission is also increased.

In this paper, we examine a low-complexity space-frequency rake receiver whose character is determined by decoupled spatial and frequency processing with respect to interfering users and spatial interference suppression. In this case, the IN space and frequency covariance matrices can be estimated very efficiently by using the outputs of the antenna elements before correlation. The IN space-frequency covariance matrix is approximated with the outputs of the rake fingers of a conventional rake receiver. Thus, there are only as many correlations required as rake fingers exist, which is a significant reduction compared to the space-frequency rake described in [2]. Moreover, the optimum weight vector can be estimated very efficiently due to the structure of the approximated IN space-frequency covariance matrix. Notice, however, that the performance of the low-complexity space-frequency rake is degraded, since joint space-frequency processing is not done with respect to interfering users in contrast to, e.g., the space-frequency rake described in [2].

This paper is organized as follows. Taking into account the uplink channel structure of WCDMA, which is described in Section 2, we explain the estimation of the SIN space-frequency covariance matrix along with the IN space-frequency covariance matrix and the estimation of the optimum weight vector in Section 3. The conventional space-time rake receiver, the space-frequency rake receiver, and the low-complexity space-frequency rake receiver are explained in Sections 4. Section 5 compares the different rake receivers by means of Monte-Carlo simulations with respect to performance and computational complexity. In the sequel, adaptive antennas are only assumed at base stations. The space-frequency rake, however, also applies to mobiles with adaptive antennas.

2. WCDMA UPLINK CHANNEL STRUCTURE

WCDMA has two types of dedicated physical channels on the uplink (and the downlink), the PCCH and the PDCH. In case of low and medium data rates, one connection consists of one PCCH and one PDCH [4]. The PDCH baseband signal of the user of interest may be expressed as

\[ s(t) = \sum_{i=0}^{N_t-1} a_i s(t - T_{ch}), \]

where \( s(t) \) is the chip rate in WCDMA is \( 1/T_{ch} = 6.096 \text{Mchip/Sec} \).

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Figure 1: Structure of the low-complexity space-frequency rake receiver. The $N_p$ pilot symbols at the beginning of each PCCH slot are used to generate the space-frequency covariance matrices $R_r$ and $R_s$. The “largest” generalized eigenvector $w$, which is the optimum weight vector, is applied to the PCCH and PDCH to obtain the remaining control symbols and the data symbols, respectively.

Moreover, the spreading sequence of the PDCH, $s(t)$, is of length $T_s = nDT$, and is composed of $nD$ chips $d_m \in \{-1, 1\}$, $1 \leq m \leq nD$. The symbols, $s(t) \in \{-1, 1\}$, are BPSK modulated. WCDMA uses a chip waveform, $p(t) \in \mathbb{R}$, characterized by a square-root raised cosine spectrum with a rolloff factor of $\alpha = 0.22$.

The PCCH baseband signal, $x_c(t)$, can be expressed in the same way. A combination of code and IQ multiplexing is used on the uplink, where the PDCH and the PCCH are spread by different spreading codes (4) and mapped to the I and Q branches, respectively, according to $s(t) = s(t) + j \cdot s(t)$. Next, the complex I + Q signal is either scrambled by a complex short code (256 chips) or by a complex long code (40960 chips) (4). For simplicity, we do not include scrambling in our notation.

3. GENERATION OF THE COVARIANCE MATRICES

We assume that the receiver is synchronized to the beginning of a slot. The searcher of a conventional rake [1] is applied to the correlator output of each antenna to estimate $N_f$ correlation peaks per antenna element. The corresponding correlation sequence is given by $x_{c,n}(T) = \sum_{m=-\infty}^{\infty} x(t-mT_s) H_m(t+mT_s)$, where $N_f$ denotes the number of pilot symbols which are transmitted at the beginning of each PCCH slot. Recall that complex scrambling is required but not considered in our notation. Each finger is characterized by its delay $\tau_m$ and complex amplitude $w_{m,n}$, where $1 \leq n \leq N_f$ and $1 \leq m \leq M$ denote the rake finger and the antenna element, respectively. Let $X_f$ contain the correlation outputs corresponding to the rake fingers according to

\[ X_f(i,n) = x_{c,i}(T) w_{m,n} \]

where $X_f(t,n)$ denotes the $n$-th element in the $i$-th column of the matrix $X_f$. Each row and each column of $X_f \in \mathbb{C}^{M \times N_f}$ corresponds to an antenna element and a delay, respectively. The length of the delay spread of the user of interest in samples is denoted by $N_s = M_s T_s$, where $M_s = T_s/T_c$ is the oversampling factor. Next, a space-frequency transformation is performed by post-multiplying the selected correlation output with the $N_s$-point DFT matrix $W_s$ according to

\[ y_n = \text{vec}(W_s X_f N_f) \in \mathbb{C}^{M_s N_f} \]

The $\text{vec}$-operator maps an $m \times n$ matrix into an $mn$-dimensional column vector by stacking the columns of the matrix. Each column of $W_s$ is of the form $w_{m,n} = \begin{bmatrix} 1 & e^{-j2\pi m/n} & \cdots & e^{-j2\pi (M_s-1) m/n} \end{bmatrix}$. Here, the columns of $W_s$ compute $M_s \leq N_s$ frequency bins centered at DC according to $W_s = \begin{bmatrix} w_{0,0} & w_{0,1} & \cdots & w_{0,(M_s-1)/2} & \cdots & w_{0,((M_s-1)/2) N_f} \end{bmatrix}$, where we have invoked the wrap-around property of the DFT matrix. Assuming that the channel stays approximately constant for at least one slot, the space-frequency SIN covariance matrix

\[ R_{ss} = y_n y_n^H \]
is estimated by using only one space-frequency snapshot that contains the multipaths of the user of interest, cf. Figure 1. If we decouple spatial and frequency processing with respect to the interferers, the IN space-frequency covariance matrix can be approximated by the Kronecker product of the frequency and the spatial covariance matrices according to

\[ R_{\text{IN}} = R_f \otimes R_s \]

where \( R_f \) and \( R_s \) denote estimates of the frequency and the spatial covariance matrices, respectively. Moreover, \( \otimes \) denotes the Kronecker product. Note that spatial interference suppression is reduced to spatial interference suppression by approximating the frequency interference and noise ratio.(4) is then improved noise suppression is attained. Note that the space-frequency SINR covariance matrix \( R_{\text{IN}} \) estimated in (3) has rank one. Therefore, a scaled version of \( u \) is given by \( R_{\text{IN}}u = y_2 \). In (2), a Cholesky factorization (3) of \( R_{\text{IN}} \) was performed in order to obtain \( u \); by solving two triangular systems, exploiting that \( R_{\text{IN}} \) is positive definite and symmetric. Due to the block diagonal structure of \( R_{\text{IN}} \), cf. (4), it is sufficient to invert only one (positive definite and symmetric) block of dimension \( M \times M \) instead of \( R_{\text{IN}} \) of dimension \( M \times M \). Each block is equal to the spatial covariance matrix \( R_s \). The snapshot corresponding to the 6th bit of the PDCH slot is obtained according to (2) as \( y_2^{(i)} = \text{vec}(X^{(i)}W) \) for operation in the space-frequency domain, where \( X^{(i)} \) is obtained as in Section 4.1. With the optimum weight vector \( w \) defined in (6), the optimum decision statistic (6) may then be computed as

\[ \hat{y}_2^{(i)} = w^* \cdot y_2^{(i)}. \]

5. SIMULATION RESULTS

A simple four-ray multipath channel is assumed for the user of interest as well as for the interfering users. The time-invariant channel parameters are given in Table 1. Weaker interfering users are modeled by adding white Gaussian noise to the channel. We assume a maximum delay spread of \( \tau_{\text{max}} = 10\mu s \) which corresponds to 40 chips. Adjacent sensors of the base station ULA are spaced by half a wavelength. Moreover, long scrambling codes were used. Further simulation parameters are given in Figure 2.
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