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Multichannel Estimation Algorithms for Wideband OFDM-Based Networks

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Abstract

This paper evaluates the performance of combining space-time transmit diversity (STTD) and an efficient multichannel estimator for wideband orthogonal frequency division multiplexing (WOFDM)-based multiantenna systems in a multipath channel. Using decision variable, we also derive the analytic bit error rate (BER) for OFDM systems applying multichannel estimator for STTD schemes. Our results show that, even with imperfect channel estimation, the multichannel estimator algorithm is effective in improving the output signal-to-noise ratio (SNR) and significantly reducing the error floor.

Keywords: wideband OFDM, multichannel estimation, pilot diversity, space-time transmit diversity, bit error probability

1 Introduction

Future mobile systems should be able to fulfill the stringent requirements for quality of service (QoS), mainly in terms of throughput, delay, and the error rate. This situation necessitates advanced transmission techniques that not only guarantee QoS, but also manage resources efficiently. The space-time codes are considered one of the most promising techniques to meet the stringent requirements of WOFDM-based systems [1][2]. In the future ubiquitous (or pervasive) communication age, all networks and services may be integrated, implying that keeping commonalities of technologies between networks is crucial.

This commonality requirement calls for evaluation of important technologies in various networks and channels.

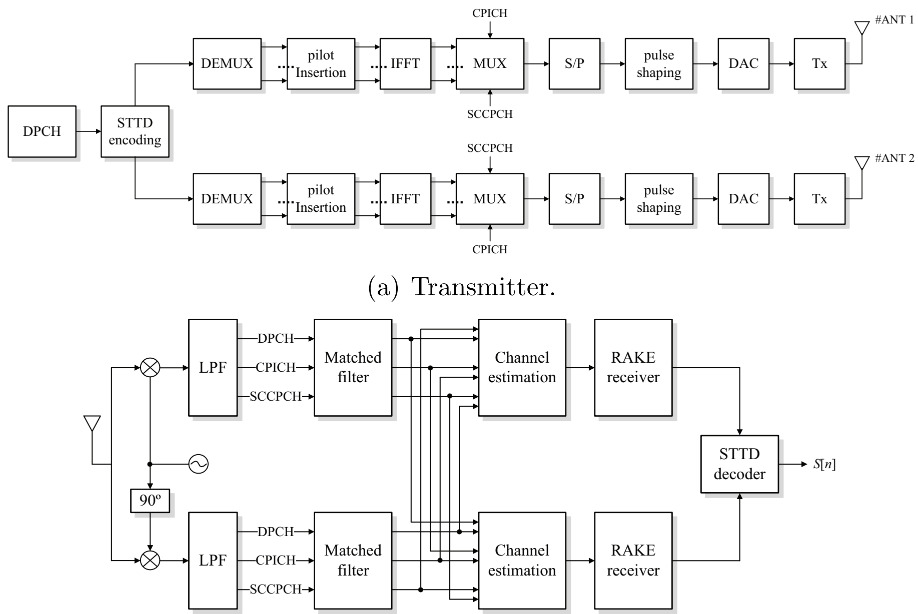
Therefore, the purpose of this paper is to evaluate the performance when the proposed multichannel estimation and space-time transmit diversity (STTD) scheme are combined in various mobile communication systems. The work presented in this paper is motivated by the previous work in [3]. This multichannel estimator estimates the channel gain by combining the common pilot symbols in the common control physical channel (CPICH) symbols and dedicated pilot symbols, which are inserted into the dedicated physical control channel (DPCH). The additional pilot diversity, the secondary common control physical channel (SCCPCH), is used for our proposed channel estimator.

2 Multichannel Estimator for Space-Time Transmit Diversity in WOFDM multiantenna Systems

Figure 1 shows the block diagram of a transmitter and receiver in the OFDM system using a STTD scheme. We propose a multichannel estimator based jointly on common, dedicated, and SCCPCH pilots. In other words, to estimate the complex amplitudes of the propagation paths, the multichannel estimator combining dedicated pilot symbols that are time-multiplexed into each slot of the DPCCH, pilot symbols of the SCCPCH and the continuous CPICH can be used for multichannel estimation algorithms. The main differences between the method in [3] and the proposed estimator is that pilot symbols of the SCCPCH are also used in the method in [3].

Figure 2 shows the relative timing relationship for the proposed estimator, where α, β, and γ are weighting factors corresponding to the received signal-to-noise ratios (SNRs) of the DPCCH, CPICH, and SCCPCH, respectively. To normalize a weighting factor, we assume that $\alpha + \beta + \gamma = 1$. Also, n, k and l denote the n-th slot, k-th symbol, and l-th resoluble multi-path, respectively. In addition $P'_{CPICH}(n,k,l)$, $P'_{DPCCH}(n,k,l)$, and $P'_{SCCPCH}(n,k,l)$ refer to the estimates of the CPICH, DPCCH, and SCCPCH, respectively. Since the Cramer-Rao lower bound (CRLB) is useful to compare the efficiencies of two different estimators, in the following we prove that combining the CPICH symbols, pilot symbols of the DPCCH, and pilot symbols of the SCCPCH can lower the CRLB if we know the power ratio $\gamma(i)$ perfectly. Using the conclusion in [3], we have

$$\frac{1}{N/\sigma_n^2 + M/\mu\sigma_m^2 + K/\mu\mu_1\sigma_k^2} \leq \frac{\sigma_m^2}{N} \quad , \tag{1}$$



(a) Transmitter.

(b) Receiver.

Figure 1: Block diagram of OFDM system using a space-time transmit diversity scheme.

where N, M, and K are the number of pilot symbols within one slot in CPICH, DPCCH, and SCCPCH, respectively. In addition, $\lambda(j) = \sqrt{1/\mu(j)}$ denotes the ratio of the power of the pilot symbols of the DPCCH to the power of the CPICH. Furthermore, $\lambda_1(k) = \sqrt{1/\mu_1(k)}$ stands for the ratio of the power of the pilot symbols of the SCCPCH to power of the CPICH. Moreover, σ_n^2, σ_m^2, and σ_k^2 are the AWGN variances when only the CPICH is used for the multichannel estimation; the DPCCH and CPICH are used in combination for the multichannel estimation; and a combination of the DPCCH, CPICH, and SCCPCH is used, respectively. Therefore, the CRLB of the multichannel estimator combining the CPICH, DPCCH, and SCCPCH has a lower CRLB than that of the estimator using only the CPICH symbols or combined CPICH and DPCCH.

In most of the previous works on STTD schemes, perfect channel estimation was assumed [1][2]. It was also assumed that channel gain was always a constant, defined as quasi-static Rayleigh fading when the transmitter transmits two symbols. Accordingly, these schemes could achieve dual diversity, obtaining about 3 dB of gain with identical performance to maximum ratio combining (MRC) in the downlink of the satellite/terrestrial OFDM multi-antenna system using space-time codes with perfect channel estimation. However, in practice, we can obtain neither the ideal channel estimation in the downlink nor quasi-static Rayleigh fading.

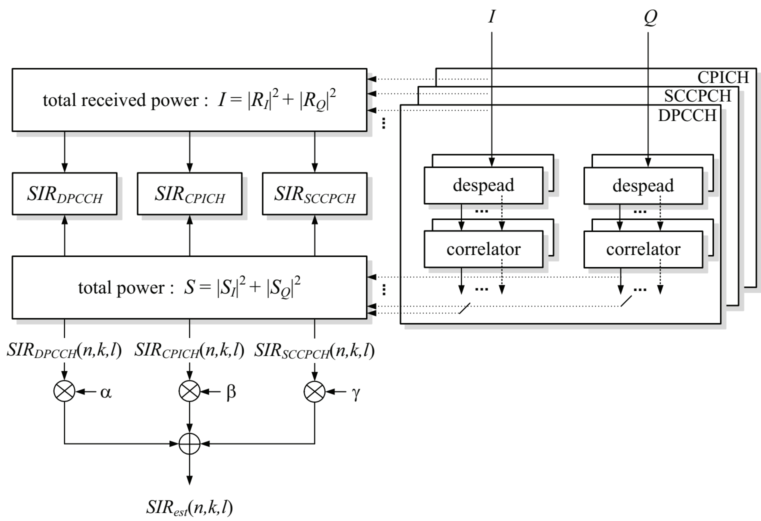


Figure 2: Timing diagram for a channel estimator.

In this paper, we evaluate the performance of the STTD schemes with imperfect channel estimation. This is because, in practice, we cannot implement an ideal channel estimator in downlink OFDM systems. We use the channel estimation error expressed in (2):

$$W_i = H_i + E_i,\ i = 0, 1 \tag{2}$$

where W_i is the estimate for channel gain H_i which is equivalent to $P'(n, k, l)$. Moreover, E_i is the channel estimation error according to an independent zero-mean complex Gaussian random variable with a variance of σ_z^2. The correlation coefficients of H_i and W_i are approximately σ_H/σ_W [2]. It is also assumed that W_i is a zero-mean complex Gaussian random variable that only depends on H_i, with correlation coefficients ρ'. In general, correlation coefficients ρ' is given by

$$\rho' = 1/\sqrt{1 + \sigma_z^2}\ . \tag{3}$$

The cross-correlation of H_i and W_i is easily seen to be $E\{H_i W_i^*\} = \rho'$, where * is the complex conjugate of W_i [4].

3 Bit Error Probability

To evaluate the bit error probability, we use the decision variable-based approach employed in [5]. Here, detection is carried out codeword by codeword;

for known channel matrix $\mathbf{H}$, the maximum likelihood (ML) detector is given by

$$[S_1' S_2'] = \max \mathrm{Re} \left((\mathbf{S}_k^*)^T (\mathbf{W}^*)^T \mathbf{A}_k \right) \quad , \tag{4}$$

where W_i is the estimate for channel gain, T signifies the transpose operation, and $\mathbf{A_k}$ denotes the equation derived from (4). Note that the ML detector performs an MRC operation on the received signal vector. Unlike the conventional MRC, here, there are two weight vectors resulting in a 2×1 output. Let

$$\mathbf{D} = [Z_1' Z_2']^T = (\mathbf{W}^*)^T \mathbf{A}_k \quad , \tag{5}$$

where Z_t^1 and Z_t^2 are the output values of the received signal through each finger of the first and second RAKE receivers at time t, respectively. Note that D_1 is dependent only on the symbol S_1, and

$$Z_1' = W_{1,t}^* \left\{ h_1(b_0 + jb_1) + N_1 \right\} + W_2 \left\{ h_2^*(-b_2 + jb_3) + N_3 \right\} \quad , \tag{6}$$

where b_i stands for the transmission of bits. The bit error probability of S_1' can be obtained from the probability density function (PDF) of Z_1'. The random variables are defined as follows: $J_1 = W_1^*$, $J_2 = W_2$, $K_1 = h_1(b_0 + jb_1) + N_1$, and $K_2 = h_2^*(-b_2 + jb_3) + N_3$; then, Z_1' can be expressed as

$$Z_1' = \sum_{i=1}^{2} J_i K_i^* \quad . \tag{7}$$

Conditional on the symbol $(b_0 + jb_1)$, the sets (X_i, Y_i), $i = 1, 2$, are two pairs of correlated, complex-valued, zero-mean Gaussian random variables. The two pairs, however, are mutually statistically independent and identically distributed. We define $Z_r = Re(Z_1')$ and $Z_i = Im(Z_1')$. The joint characteristic function $\phi(jv_1, jv_2)$ of the random variables Z_r and Z_i can be obtained using [5]. An alternative interpretation to (7) is that the phase of Z_1' is the decision variable for the detection of $(b_0 + jb_1)$. We define $R = \sqrt{Z_r^2 + Z_i^2}$, and $\phi = \tan^{-1}(Z_i/Z_r)$. Our goal is to obtain the PDF $p(\phi)$, where lower case notations for realization of the corresponding upper case denote random variables. This is achieved as follows: We compute the joint PDF of Z_r and Z_i, $p(z_r, z_i)$, from the Fourier transform of the joint characteristic function $\phi(jv_1, jv_2)$. From $p(z_r, z_i)$, we obtain $p(r, \theta)$, the joint PDF of the envelope R and phase θ. By integrating $p(r, \theta)$ over the variable r, we can obtain the PDF $p(\theta)$. The result can be found in [4]. To obtain normalized covariance in case of Rayleigh multipath fading, we perform the following calculations:

$$\begin{aligned} m_{xx} &= E\{|J_1|^2\} = 1 + \sigma_z^2 + 2(\rho' - 1) \quad , \\ m_{yy} &= E\{|K_1|^2\} = 1/2 + N_0 \quad , \end{aligned} \tag{8}$$

where $E\{|h_1|^2\} = 1$ and $E\{|(b_0 + jb_1)|^2\} = 1/2$, as normalized power is employed for the two transmit antennas on OFDM using STTD. Thus,

$$m_{xy} = E\{J_i K_i^*\} = \frac{1}{\sqrt{2}}\rho' \quad . \tag{9}$$

Therefore, we can achieve the normalized cross-correlation μ as follows:

$$\mu = \frac{m_{xy}}{\sqrt{m_{xx} m_{yy}}} = \frac{1/\sqrt{2}\rho'}{\left(\sqrt{1 + \sigma_z^2 + 2(\rho' - 1)}\sqrt{1/2 + N_0}\right)} \quad . \tag{10}$$

The error probability of the received signal can be obtained by integrating $p(\theta)$ over the angle interval complementary to the correct decision. According to [5][6], we can obtain the bit error probability for OFDM using STTD over the Rayleigh fading channel in QPSK modulation while integrating $p(\theta)$ from $\pi/4$ to $5\pi/4$. The symbol error probability is expressed as

$$P_b' = 2\int_{\pi/4}^{3\pi/4} p(\theta)d\theta + 4\int_{3\pi/4}^{\pi} p(\theta)d\theta \quad . \tag{11}$$

Thus, the bit error probability is given by

$$P_b = \frac{1}{2}\left(1 - \frac{3}{2}\rho + \frac{1}{2}\rho^2\right) \quad , \tag{12}$$

where $\rho = \mu/\sqrt{(2 - \mu^2)}$. The analytic bit error probability for the OFDM using STTD according to imperfect channel estimation is shown in Figure 3.

4 Simulation Results and Discussion

Figure 4 shows the BER performance of OFDM combining the efficient multichannel estimator and STTD schemes when the incomplete multichannel estimation is 0.0 dB and 0.3 dB. It turns out that the multichannel estimator combining the CPICH, DPCCH, and SCCPCH has a lower BER than

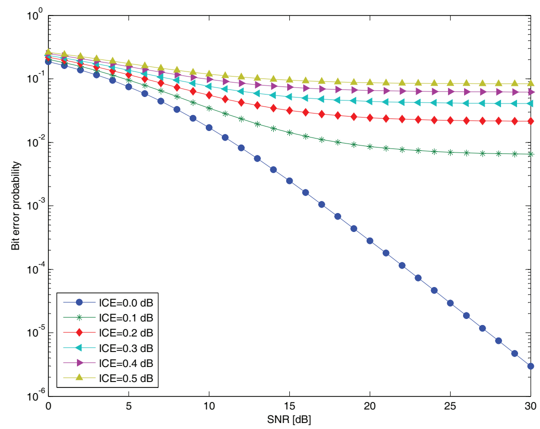


Figure 3: BER performance for the OFDM using STTD according to imperfect channel estimation.

the estimator that uses only the CPICH symbols or combined CPICH and DPCCH does, despite the use of imperfect channel estimation. A drawback of the conventional channel estimation method is that it incurs more errors when a channel related to the channel estimation is under deep fading. Therefore, it is one of the objectives of this paper to provide a multichannel estimation method that exhibits better performance than the multichannel estimation in a receiver of a terminal with the conventional pilot symbols of a CPICH or DPCCH by combining the pilot symbols of the CPICH, DPCCH, and SC-CPCH and estimating the channel.

5 Conclusions

In this paper, we analyzed the performance of the OFDM multiantenna system combining STTD and a multichannel estimator over various mobile communications channels. Our results show that, even with imperfect channel estimation, the efficient multichannel estimator algorithm is effective in improving the output SNR and significantly reducing the error floor. Furthermore, the simulation results investigated in this paper also reveal that the OFDM combining the efficient multichannel estimator and STTD scheme could provide appreciable performance improvements in the presence of imperfect channel estimation over Rayleigh multipath fading channels. The performance improvement was confirmed by the simulation results. For further improvement in performance and efficiency, additional advanced techniques are required. These include powerful error correction coding, smart antennas, advanced multichannel esti-

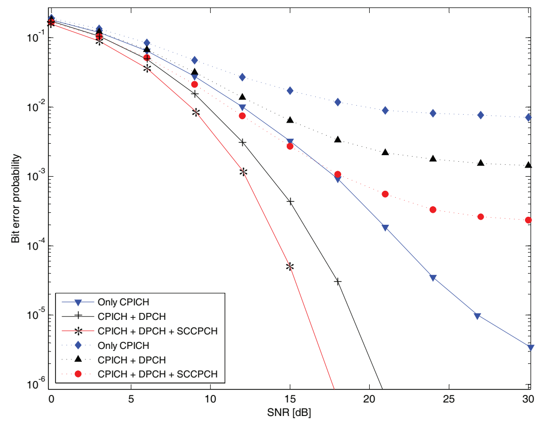


Figure 4: BER performance for OFDM combining efficient multichannel estimator and STTD.

mation, and the optimal combination of these elements.

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