MIMO GENERALIZED DECORRELATING DISCRETE-TIME RAKE RECEIVER

Tunçer Baykaş, Mohamed Siala, Abbas Yongaçoḡlu

School of Information Technology and Engineering
University of Ottawa
Ottawa, Canada
tbaykas@site.uottawa.ca

SUP'COM
Cité Technologique des Communications
Ariana,Tunisia
mohamed.siala@supcom.rnu.tn

School of Information Technology and Engineering
University of Ottawa
Ottawa,Canada
yongacog@site.uottawa.ca

ABSTRACT

In this paper, we introduce a generalized decorrelating discrete-time RAKE receiver for MIMO systems (MIMO GD-DTR). The MIMO GD-DTR system is a combination of two other advanced MIMO RAKE reception methods: the jointly decoding generalized rake receiver (JD-GRAKE) and the MIMO decorrelating discrete-time RAKE reception (MIMO D-DTR). The JD-GRAKE has been proposed for correlated interference suppression in MIMO systems and it is obtained from the single antenna generalized RAKE receiver (G-RAKE). The MIMO D-DTR, which is obtained from the single antenna decorrelating discrete-time RAKE (D-DTR) system, improves the performance in the presence of channel estimation errors in diffuse channels. The MIMO GD-DTR combines the complementary advantages of both the JD-GRAKE and the MIMO D-DTR. It suppresses the interference, and takes into consideration the imperfect channel state information at the receiver. Our results show that the performance of the discrete-time version of the MIMO JD-GRAKE could be worse than a conventional RAKE receiver, when there are channel estimation errors in the system, whereas proposed MIMO GD-DTR provides gains up to 1.7 dB at a bit error rate of 10^{-2} in 3 transmit 3 receive antenna system (3×3).

1. INTRODUCTION

The conventional RAKE receivers, which are used in wireless communication to collect energy from multipath channels, have several weaknesses. This paper focuses on two of these weaknesses with the intention of reducing them for multi antenna (MIMO) systems. The first weakness is due to colored interference. Colored interference could be a major source of degradation in the system performance, since the conventional RAKE receivers assume the interference to be uncorrelated. Bottomley *et al.* proposed in [1] a generalized RAKE receiver in order to have colored (correlated) interference suppression for one transmit and one receive antenna systems. As a generalization of this principle to MIMO systems, Grant *et al.* have proposed two methods in [2]. The JD-GRAKE jointly decodes the symbols spread by the same spreading code and transmitted from different antennas. The MMSE-GRAKE is an alternative system which is less complex than the JD-GRAKE. In the MMSE-GRAKE the symbols are decoded one by one, by treating the rest of the symbols as interference. Such a decoding decreases the complexity at the price of a decrease in performance. The flaw of all theses G-RAKE systems is the nonrealistic assumption of having perfect channel state information at the receiver.

The second weakness arises from the time-varying nature of the wireless channels. Conventional RAKE receivers use an acquisition system (or searcher) to detect new paths with significant power, and a tracking system to follow the continuously time-varying delays of the paths [3, 4]. The acquisition and tracking systems have two main drawbacks. First, the assumption that the multipaths are resolvable is nonrealistic [4]. Secondly, the acquisition and tracking systems can neither distinguish nor follow paths separated by less then a chip period [5, 6]. To eliminate the need for complex acquisition and tracking devices, a discrete-time RAKE receiver (DTR) has been proposed in [7]. It is obtained by a lossless sampling of a transmit filtered version of the channel impulse response. The simulations show that the DTR is sensitive to channel estimation errors. To cope with this sensitivity, several intuitive methods have been presented in [7] and the derivation of an optimum structure, complying with the maximum *a posteriori* (MAP) criterion, has been made in [4]. This optimum structure, referred to as the decorrelating discrete-time RAKE (D-DTR), exploits the covariance matrix of the channel values at RAKE fingers to obtain robustness against channel estimation errors. In [8], the D-DTR system is extended for the MIMO case.

In this work we combine the discrete-time version of the JD-GRAKE receiver with the MIMO D-DTR receiver and obtain the MIMO generalized discrete-time RAKE receiver. The MIMO GD-DTR could suppress the interference and takes into consideration the channel estimation errors. We obtain the performance of the proposed receiver in a UMTS uplink model as well as the performances of the conventional MIMO discrete-time RAKE receiver (MIMO C-DTR), the MIMO D-DTR and the discrete-time version of the MIMO JD-GRAKE. From this point on, we will refer the discrete-time version of the MIMO JD-GRAKE as MIMO generalized discrete-time RAKE receiver (MIMO G-DTR) to ease the presentation.

The outline of the paper is as follows. In the next section we explain the system model and receivers for single antenna systems. Afterwards, we show how to extend the GD-DTR to the MIMO GD-DTR. Section 4 presents the simulation results. We finish the paper with the conclusions.

2. SYSTEM MODEL AND DISCRETE-TIME RAKE RECEIVERS

In this section we explain the single antenna versions of the RAKE receivers using a basic DS-CDMA transmitter model. The modulated symbol s_k is spread using a long spreading code $c_k = (c_{k0}, c_{k1} ..., c_{kQ-1})$. The spreading factor of the code is denoted by Q and c_{kq} is the qth chip of the spreading code c_k. We assume that the code c_k has the perfect correlation property to eliminate the interchip interference. This assumption is asymptotically met with large spreading factors. The spread symbol is convolved with a pulse shaping filter with impulse response $g(t)$ and transmitted over a quasi-static diffuse multipath Rayleigh fading channel. Throughout this paper we assume that the modulation scheme is BPSK. The channel impulse response changes independently from one time slot to another, but the multipath intensity profile of the channel does not change. Assuming the impulse response of the channel for the time slot comprising s_k to be $h(t)$, the received signal $r(t)$ for the symbol s_k can be written as

$$r(t) = h(t) \otimes g(t) \otimes \Big(\frac{s_k}{\sqrt{Q}} \sum_{q=0}^{Q-1} c_{kq} \delta(t - qT_c) \Big) + n(t),$$

where $\otimes$ and $\delta(t)$ denote the convolution operator and Dirac delta function, respectively. In the above equation, T_c denotes the chip period and $n(t)$ denotes the noise plus interference. All of the receiver types we mention in this paper are discrete-time receivers. They sample the received symbol at a rate of $T_c/2$, i.e. at twice the chip rate, and passes it to the correlators. The sampling is lossless for square-root raised cosine filters with rolloff less than 1. In the correlators, the symbol s_k is despread using the oversampled version of c_k. Due to the assumption that the spreading sequence is perfect in nature, the output vector of the correlators, $\mathbf{y}$, for the symbol s_k can be written as

$$\mathbf{y} = \mathbf{f} s_k + \mathbf{u},$$

where $\mathbf{u}$ is a zero-mean complex gaussian additive noise plus interference vector with covariance matrix $\mathbf{R_u}$, and the vector $\mathbf{f}$ consists of elements from the sampled version of the filtered channel response $h(t) \otimes g(t)$. Although $\mathbf{y}, \mathbf{f}$ and $\mathbf{u}$ depend on the time index k, to ease the presentation we only keep the index in s_k.

If the estimate of $\mathbf{f}$, denoted by $\hat{\mathbf{f}}$, is available, the C-DTR obtains the decision variable ψ by forming $\psi = \hat{\mathbf{f}}^H \mathbf{y}$, where the superscript H denotes hermitian transpose. For BPSK modulation, a hard decision on the symbol is obtained from sgn(Re(ψ)), where Re is the real part operator. As we mentioned above, the C-DTR does not take into consideration the correlation matrix $\mathbf{R_u}$ and also the erroneous nature of the estimated channel. The G-DTR uses $\mathbf{R_u}$ to obtain a decision variable for s_k, which is equal to

$$\psi = (\mathbf{R_u}^{-1} \hat{\mathbf{f}})^H \mathbf{y}.$$

If $\hat{\mathbf{f}}$ is equal to $\mathbf{f}$, i.e. the channel is perfectly known at the receiver, this method provides the maximum likelihood detection of the symbols [1]. The covariance matrix $\mathbf{R_u}$ depends on thermal noise and interference. The noise in $\mathbf{u}$ is assumed to be AWGN with variance $\Sigma^2 = 2N_0/T_c$, due to sampling, where N_0 is the power spectral density of the complex noise. In the UMTS uplink, the (x,y)th element of the $\mathbf{R_u}$ which corresponds the covariance between the interference values of the correlators with discrete delays d_x and d_y can be approximated as:

$$R_u(x,y) = \Sigma^2 \delta_{x,y} + E_I \int_{-\infty}^{\infty} g(d_x - \tau) g^*(d_y - \tau) d\tau,$$

where E_I denotes the power spectral density of the interference and $\delta_{x,y}$ denotes the Kronecker delta function. The number of interferers and the effectiveness of the power control could affect the values of E_I. In our simulations, to set the value of E_I, we define a noise rise coefficient η, which is equal to

$$\eta = (E_I + N_0)/N_0.$$

The noise rise coefficient is also related to the system load x through $\eta = 1/(1-x)$ [9]. The system load shows the percentage of energy coming from interference sources to the noise plus interference in the system.

The D-DTR takes into consideration that only an estimate $\hat{\mathbf{f}}$ of the channel vector is available, such that

$$\hat{\mathbf{f}} = \mathbf{f} + \mathbf{e},$$

where $\mathbf{e}$ is the channel estimation error vector. We assume in this work that $\mathbf{f}$ is estimated with N_p pilot symbols and the estimated channel vector is equal to:

$$\hat{\mathbf{f}} = \frac{1}{N_p E_p} \sum_{m=0}^{N_p-1} \mathbf{y}_m s_m^*,$$

where E_p is the energy per pilot symbol, $\mathbf{y}_m$ is the received vector of the mth pilot symbol and the s_m's are the pilot symbols within a time slot. The D-DTR uses the channel covariance matrix $\mathbf{R_f}$, which is defined as

$$\mathbf{R_f} = E[\mathbf{f}\mathbf{f}^H],$$

to construct the decision variable. The channel covariance matrix $\mathbf{R_f}$ can be estimated using the estimated channel values or calculated if the delay profile of the channel and the impulse response of the pulse shaping filter are known. The decision variable of the D-DTR is equal to:

$$\psi = (\mathbf{U}\mathbf{W}\mathbf{U}^H \hat{\mathbf{f}})^H \mathbf{y},$$

where $\mathbf{U}$ is the matrix consisting of the eigenvectors of the channel covariance matrix $\mathbf{R_f}$. The weighting matrix $\mathbf{W}$ is a diagonal matrix with the lth diagonal entry being a weighting factor w_l. If the data and pilot symbols have common energy E_s, then w_l is equal to

$$w_l = \frac{1}{1 + \sigma^2 (\frac{E_s}{\Sigma^2} + \frac{1}{\Gamma_l})},$$

where σ^2=$\Sigma^2/(N_p E_s)$ and Γ_l is the eigenvalue corresponding to the lth eigenvector of $\mathbf{R_f}$. In [4], it is shown that the above weights are optimum under the assumptions of estimated channels and white noise.

2.1 The GD-DTR System

The GD-DTR uses both covariance matrices $\mathbf{R_f}$ and $\mathbf{R_u}$ to combine complimentary advantages of the D-DTR and the G-DTR. The GD-RAKE first decorrelates the colored noise $\mathbf{u}$ in $\mathbf{y}$ by using the eigenvectors and eigenvalues of $\mathbf{R_u}^{-1}$ such that,

$$\mathbf{y}' = \Lambda \Omega^H \mathbf{y} = \Lambda \Omega^H (\mathbf{f} s_k + \mathbf{u}),$$

where Λ is the diagonal matrix of the square roots of the eigenvalues of $\mathbf{R}_u^{-1}$ and Ω is the eigenvector matrix of $\mathbf{R_u}^{-1}$.

On the other hand, the channel estimate $\hat{\mathbf{f}}$ is also multiplied with $\Lambda\ \Omega^H$ such that,

$$\hat{\mathbf{f}}' = \Lambda\Omega^H\hat{\mathbf{f}}.$$

In the above equation $\hat{\mathbf{f}}'$ should be understood as the estimate of the equivalent channel vector of a received vector with an equivalent white noise vector. The decorrelation of noise changes the covariance matrix $\mathbf{R_f}$ to $\mathbf{R}'_\mathbf{f}$ as well,

$$\mathbf{R}'_\mathbf{f} = \Lambda\Omega^H\mathbf{R_f}\Omega\Lambda.$$

At this point the system has white gaussian noise samples with unit energy (Σ'^2=1), since we used the eigenvalues of the $\mathbf{R_u}^{-1}$ in the noise decorrelating process. Additionally, the channel vector has the new covariance matrix $\mathbf{R}'_\mathbf{f}$. Since this situation matches the assumptions of the D-DTR, that method can be applied directly. Then the decision variable is equal to:

$$\psi = (\mathbf{U}'\hat{\mathbf{f}}')^H\mathbf{W}'\mathbf{U}'\mathbf{y}',$$

where $\mathbf{U}'$ consists of the eigenvectors of $\mathbf{R}'_\mathbf{f}$ and $\mathbf{W}'$ is a diagonal matrix with the lth diagonal entry being the weighting factor w'_l, which is equal to

$$w'_l = \frac{1}{1+\frac{1}{N_pE_s}(E_s+\frac{1}{\Gamma'_l})},$$

where Γ'_l is the lth eigenvalue of $\mathbf{R}'_\mathbf{f}$.

3. THE MIMO GD-DTR SYSTEM

In this section we explain the MIMO generalized decorrelating discrete-time RAKE receiver. For the MIMO GD-DTR, we assume a system with M transmit and N receive antennas. For each receive antenna there are L correlators. The spreading code is the same for the transmit antennas of a user to reduce the consumption of orthogonal codes and preserve orthogonality between different users in the system. The signals from different transmit antennas are synchronized. The channels between any transmit and receive antennas are independent and have the same delay profile and therefore the same covariance matrix $\mathbf{R_f}$. The received signal, after sampling and despreading at the nth receive antenna, is equal to

$$\mathbf{y}_n = \mathbf{F}_n\mathbf{s}_k + \mathbf{u}_n,$$

where $\mathbf{F}_n$ is the channel matrix between the transmit antennas and the nth receive antenna, $\mathbf{s}_k$ consists of the symbols sent simultaneously from all the transmit antennas. The noise plus interference vector $\mathbf{u}_n$ has the covariance matrix $\mathbf{R_u}$. The MIMO GD-DTR follows the steps of the GD-DTR. The MIMO GD-DTR starts with the decorrelation process of the noise plus interference using the matrices Λ and Ω resulting from the decomposition of the covariance matrix $\mathbf{R_u}$, such that,

$$\mathbf{y'}_n = \Lambda\Omega^H\mathbf{y}_n = \Lambda\Omega^H(\mathbf{F}_n\mathbf{s}_k + \mathbf{u}_n),$$

and $\hat{\mathbf{F}}'_n$=$\Lambda\Omega^H\hat{\mathbf{F}}_n$, where $\hat{\mathbf{F}}$ is the estimation of $\mathbf{F}$. Afterwards the decorrelating process of the channels are applied to $\mathbf{y'}_n$, as:

$$\mathbf{z}_n = \mathbf{U'}^H\mathbf{y'}_n,$$

and to $\hat{\mathbf{F}}'_n$, such that

$$\hat{\mathbf{E}}_n = \mathbf{U'}^H\hat{\mathbf{F}'}_n.$$

In [8], the probability of receiving $\mathbf{z}_n$, given the transmitted symbol vector $\mathbf{s}$, $\hat{\mathbf{E}}_n$ and Σ^2, is given as:

$$p(\mathbf{y}_n|\mathbf{s},\hat{\mathbf{E}}_n) = \prod_{l=0}^{L-1}\frac{1}{\Sigma^2\pi}\Big(\frac{1+\frac{\sigma^2}{\Gamma_l}}{w_l}\Big)^M \exp(-X_{nl})$$

where X_{nl} is equal to

$$X_{nl} = \frac{1}{\Sigma^2}\Big(\frac{1+\frac{\sigma^2}{\Gamma_l}}{w_l}\Big)^M\Big|\mathbf{y}_{nl} - \sum_{m=0}^{M-1}\frac{\hat{e}_{mnl}s_m}{1+\frac{\sigma^2}{\Gamma_l}}\Big|^2 \qquad (1)$$

where $\hat{e}_{mnl}$ is the lth estimated channel coefficient from the mth transmit antenna to the receive antenna n. In (1), s_m and y_{nl} denote the symbol from the mth transmit antenna and the lth element of $\mathbf{y}_n$ respectively. For the GD-DTR system, due to the noise decorrelating process, Σ^2=1 and σ^2=$1/N_pE_s$. The MIMO GD-DTR chooses the symbol vector $\hat{\mathbf{s}}$ which maximizes the following quantity:

$$\prod_{n=0}^{N-1} p(\mathbf{y}_n|\mathbf{s}=\hat{\mathbf{s}},\hat{\mathbf{E}}_n),$$

with respect to all possible $\mathbf{s}$ vectors. The soft output of the bit from the mth transmit antenna, Λ_m, can be obtained from the above equation, such that:

$$\Lambda_m = \log\frac{\prod_{n=0}^{N-1}p(\mathbf{y}_n|s_m=-1,\hat{\mathbf{E}}_n)}{\prod_{n=0}^{N-1}p(\mathbf{y}_n|s_m=1,\hat{\mathbf{E}}_n)}.$$

The MIMO D-DTR system does not apply the noise decorrelating process, whereas the MIMO G-DTR system does not apply the channel decorrelating process and therefore does not have the weight matrix $\mathbf{W}$. The MIMO C-DTR system applies neither the noise decorrelating process nor the channel decorrelating process. In the next section we present the simulation results for 2x2 and 3x3 systems.

4. SIMULATION RESULTS

To show the performance of the MIMO GD-DTR, we first choose a 2 transmit and 2 receive antenna (2×2) system. The channel has an exponential delay profile with RMS delay spread of T_m=$0.25T_c$, which is observed in UMTS channels in [10] by Foo *et al*. The pulse shaping filter is a root raised cosine filter with roll off $\alpha = 0.22$. The number of correlators L is equal to 9. This system collects %99 of the channel energy according to [4]. The number of pilot symbols per channel realization, N_p, is equal to 2. To simulate a highly loaded system we set η=10. The results are shown in Figure 1. The curves from top to bottom follow the same order of the legend. The gain of the MIMO GD-DTR compared to MIMO C-DTR is about 1.4 dB at a bit error rate of 10^{-2} and employing the MIMO G-DTR actually causes a performance loss compared to the MIMO C-DTR. Since the performance of the MIMO D-DTR is close to the performance of the MIMO GD-DTR, we can conclude that most of the gain of the MIMO GD-DTR comes from suppressing

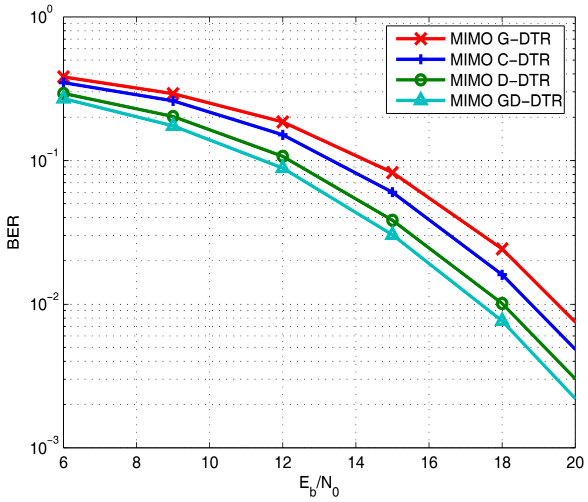

Figure 1: The BER performances of (from top to bottom, in that order) the MIMO G-DTR, the MIMO C-DTR, the MIMO D-DTR and the MIMO GD-DTR for a 2×2 system

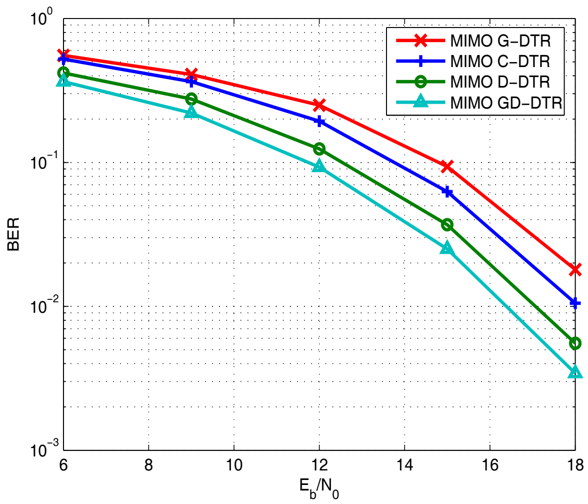

Figure 2: The BER performances of (from top to bottom, in that order) the MIMO G-DTR, the MIMO C-DTR, the MIMO D-DTR and the MIMO GD-DTR for a 3×3 system

noisy channel estimates.
We did another set of simulations with a 3×3 system. We observe a slight increase in the gain of the MIMO GD-DTR system with respect to the MIMO C-DTR. For example, in Figure 2, which shows the performance of 3×3 system with N_p=2 and L=9, the gain of the MIMO GD-DTR is around 1.7 dB at a 10^{-2} bit error rate compared to the MIMO C-DTR.
Although it is not shown, we did other simulations to see the effect of different parameters on the system performances. We can conclude that the gain of the MIMO GD-DTR with respect to the MIMO C-DTR increases with a decrease of the delay spread and with an increase of the number of correlators and the number of antennas. The MIMO G-DTR system gives the worst results in all simulations with channel estimation errors. Even if the perfect channel state information is available, the gain of the MIMO G-DTR is very limited. The performance of the MIMO D-DTR receiver is always between the MIMO GD-DTR and the C-DTR.

5. CONCLUSION

In this paper, we have introduced a generalized decorrelating discrete-time RAKE receiver for MIMO systems to improve the overall performance in the presence of channel estimation errors and colored noise plus interference. The system exploits the covariance matrix of the channel and the covariance matrix of the noise plus interference to produce combining weights.
Our results showed that gains up to 1.7 dB with respect to the conventional discrete-time RAKE receiver are available for 3×3 systems. We observed that low number of pilot symbols, short delay spreads and high number of correlators and antennas increase the gain of the MIMO GD-DTR with respect to the conventional MIMO RAKE receiver. We simulated the MIMO G-DTR and the MIMO D-DTR systems as well. The MIMO G-DTR system is not suitable since its performance is worse than the MIMO C-DTR if the channel coefficients are estimated. The MIMO D-DTR performs better than the MIMO C-DTR but worse than the MIMO GD-DTR.

REFERENCES

[1] G. E. Bottomley, T. Ottosson, and Y. E.Wang, “A generalized RAKE receiver for interference suppression,” *IEEE J. Select. Areas Commun.*, vol. 18, pp. 1536–1545, Aug. 2000.

[2] S. Grant, K. Molnar, and G. Bottomley, “Generalized RAKE receivers for MIMO systems,” in *Proc. IEEE 54th Vehicular Technology Conference, VTC Fall 2003*, vol. 1, Oct. 2003, pp. 424–428.

[3] A. Viterbi, *CDMA: Principles of Spread Spectrum Communication*. Addison-Wesley, 1995.

[4] M. Siala, “Maximum *a posteriori* decorrelating discrete-time RAKE receiver,” *Annals of Telecommunications*, vol. 59, no. 3-4, pp. 374–411, Apr./Mar. 2004.

[5] Q. Zhang, J. Huang, and Y. Xie, “High resolution delay estimation,” in *TENCON’93*, 1993, pp. 579–583.

[6] R. V. Nee, “The multipath estimation delay lock loop,” in *Int. Sym. on Spread Spectrum Tech. and App.*, Dec. 1992, pp. 39–42.

[7] H. Boujemaa, O. Fratu, M. Siala, and P. Loubaton, “On the performance of a discrete-time RAKE,” in *The 11th IEEE Int. Sym. on Personal, Indoor and Mobile Radio Com.*, vol. 2, Sept. 2000, pp. 949–953.

[8] T. Baykas, M. Siala, and A. Yongacoglu, “MIMO decorrelating discrete time RAKE receiver,” in *Proc. IEEE 57th Vehicular Technology Conference, VTC Spring 2005*, May 2005.

[9] H. Holma and A. Toskala, *WCDMA for UMTS: Radio Access for Third Generation Mobile Communications*. Wiley, 2004.

[10] S. Foo *et al.*, “Spatio-temporal investigation of UTRA-FDD channels,” in *Third Int. Conf. on 3G Mobile Com. Tech.*, May 2002, pp. 175–179.