Performance improvement of HIPERLAN/2 by adaptive receiver antennas

Torsten Bing [1], Egon Schulz [2], Bernhard Wegmann [3], Edgar Bolinth [4]
[1] Research Group for RF Communications, University of Kaiserslautern
P.O. Box 3049, D-67653 Kaiserslautern, Germany
e-mail: bing@rhrk.uni-kl.de, phone: +49 631 205 3880, fax: +49 631 205 3612
[2,3] Siemens AG, ICM CA MR MA, Hofmannstraße 51, D-81359 München, Germany
[4] Siemens AG, ICM CD TI, Frankenstraße 2, D-46393 Bocholt, Germany

Abstract—**The IST (Information Society Technologies) Project BRAIN (Broadband Radio Access for IP based Networks) is working on a broadband radio access system with the goal to integrate 2G, 3G and broadband WLAN systems through a common IP based network platform [1]. The BRAIN radio access system is based on HIPERLAN/2 (HIgh PErformance Radio LAN type 2). With respect to spectrum efficiency, and the support of IP, enhancements on all protocol layers of the HIPERLAN/2 standard are proposed and evaluated in the BRAIN Project [3-5]. Since adaptive antennas are considered to be one of the most important techniques to increase spectrum efficiency their application to HIPERLAN/2 is almost mandatory. In this paper a simple adaptive antenna concept, consisting of multiple independent omnidirectional receiver antennas and the ZF (Zero Forcing) algorithm for the adaptive signal processing, is applied to the HIPERLAN/2 air interface and its performance is evaluated at link level. The achieved gain is significant.**

A. INTRODUCTION

Multimedia applications, Internet access and high data rate services make great demands on wireless communications systems. Third generation (3G) cellular mobile radio networks like UMTS (Universal Mobile Telecommunications System) are in the position to provide bit rates in the order of 2 Mbps. Their initial deployment will start in 2001 in order to supplement the worldwide successfully operating second generation (2G) systems by new and more powerful services. However, since 3G systems are constrained by wide area coverage and scarce spectrum, their performance is insufficient for many applications. High speed wireless local area networks (WLANs) are specifically designed to provide high bit rates for local and short range communications. Therefore, broadband WLANs are ideally suited to complement 3G systems in hot spots.

HIPERLAN/2 is a new European standard specified by the ETSI (European Telecommunications Standards Institute) Project BRAN (Broadband Radio Access Networks) operating in the 5 GHz band and providing bit rates of at least 20 Mbps [2]. The evaluation and implementation of the HIPERLAN/2 standard incorporating powerful performance improving techniques is currently the topic of various research activities. The IST Project BRAIN, which is partly funded by the European Commission, is working on a broadband radio access system with the goal to integrate 2G, 3G and broadband WLAN systems through a common IP based network platform [1]. The BRAIN radio access system is based on HIPERLAN/2. With respect to spectrum efficiency and the support of IP, enhancements on all protocol layers of the HIPERLAN/2 standard are proposed and evaluated in the BRAIN Project [3-5].

The utilization of adaptive antennas is considered to be one of the most important measures to increase capacity in cellular mobile radio systems and is presently studied worldwide [5-11]. Adaptive antenna techniques are capable of reducing the required transmission power and combating interference. Therefore, the application of adaptive antennas to HIPERLAN/2 within the BRAIN Project seems to be unavoidable and very promising with respect to system efficiency. Various adaptive antenna concepts exist and their applicability is manifold [6]. The benefits of adaptive antenna techniques applied to the receiver is the improvement of the receiver's performance by exploiting the diversity information contained in the received signals of the multiple antennas. Depending on the antenna characteristics and their arrangement different basic diversity effects like the separation, the reuse or the introduction of new signal paths can be exploited [6]. This paper presents a basic adaptive receiver antenna concept, which consist of multiple omnidirectional receiver antennas and the ZF algorithm [12] applied to the HIPERLAN/2 air interface. The antennas are arranged in a macro structure, i.e. they are so far apart, at least several wavelengths, that at each of the antenna locations different wave fronts impinge. This macro structure based concept enables the reception of signals over additional paths and, therefore, provides spatial macro diversity [6].

In Section B the underlying OFDM system model is mathematically presented. Based on this mathematical representation the ZF detector which carries out the adaptive processing of the multiple antenna signals is derived in Section C. The content of Section D is the presentation and discussion of the simulation results. The

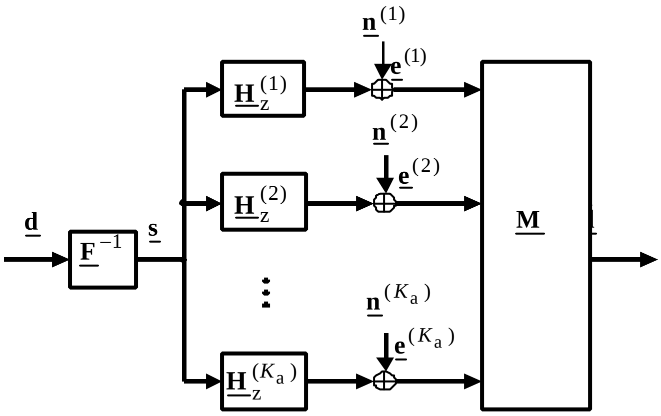

Fig. 1: OFDM multi antenna system model

conclusions are drawn in Section E.

B. System Model

Fig 1. shows the underlying discrete time low pass OFDM system model for multiple receiver antennas. The data vector $\underline{\mathbf{d}} = \left(\underline{d}_1, \underline{d}_2 \ldots \underline{d}_N\right)^{\mathrm{T}}$ to be transmitted consists of N complex modulation symbols $\underline{d}_i$ which are processed in parallel by an N-point Inverse Discrete Fourier Transformer (IDFT), represented by the IDFT matrix $\underline{\mathbf{F}}^{-1}$. The output $\underline{\mathbf{s}} = \left(\underline{s}_1, \underline{s}_2 \ldots \underline{s}_N\right)^{\mathrm{T}}$ of the IDFT is transmitted through the K_{a} mobile radio channels between the single antenna of the transmitter and the K_{a} antennas of the receiver. Each of the K_{a} channels is characterized by its particular impulse response

$$\underline{\mathbf{h}}^{(k_{\mathrm{a}})} = \left(\underline{h}_1^{(k_{\mathrm{a}})}, \underline{h}_2^{(k_{\mathrm{a}})} \ldots \underline{h}_W^{(k_{\mathrm{a}})}\right)^{\mathrm{T}},\; k_{\mathrm{a}} = 1 \ldots K_{\mathrm{a}}\,. \tag{1}$$

In HIPERLAN/2 the IDFT output $\underline{\mathbf{s}}$ is cyclically extended in order to maintain the orthogonality of the subcarriers [2, 10]. Here, the cyclic extension of $\underline{\mathbf{s}}$ is modeled by using the cyclic channel matrices $\underline{\mathbf{H}}_{\mathrm{z}}^{(k_{\mathrm{a}})}$, $k_{\mathrm{a}} = 1 \ldots K_{\mathrm{a}}$, which represent the cyclic convolution with the channel impulse responses of (1). This model is valid if the length of the cyclic extension is smaller than or equal to W, the length of the channel impulse responses, see (1). With the interference vector

$$\underline{\mathbf{n}}^{(k_{\mathrm{a}})} = \left(\underline{n}_1^{(k_{\mathrm{a}})}, \underline{n}_2^{(k_{\mathrm{a}})} \ldots \underline{n}_N^{(k_{\mathrm{a}})}\right)^{\mathrm{T}},\; k_{\mathrm{a}} = 1 \ldots K_{\mathrm{a}}\,, \tag{2}$$

we obtain the received signal vector

$$\underline{\mathbf{e}}^{(k_{\mathrm{a}})} = \left(\underline{e}_1^{(k_{\mathrm{a}})}, \underline{e}_2^{(k_{\mathrm{a}})} \ldots \underline{e}_N^{(k_{\mathrm{a}})}\right)^{\mathrm{T}},\; k_{\mathrm{a}} = 1 \ldots K_{\mathrm{a}}\,, \tag{3}$$

at the k_{a}-th antenna element. The cyclic channel matrices $\underline{\mathbf{H}}_{\mathrm{z}}^{(k_{\mathrm{a}})}$, $k_{\mathrm{a}} = 1 \ldots K_{\mathrm{a}}$, of all channels are arranged in the $K_{\mathrm{a}} N \times N$ total cyclic channel matrix

$$\underline{\mathbf{H}}_{\mathrm{z}} = \left(\underline{\mathbf{H}}_{\mathrm{z}}^{(1)\mathrm{T}}, \underline{\mathbf{H}}_{\mathrm{z}}^{(2)\mathrm{T}} \ldots \underline{\mathbf{H}}_{\mathrm{z}}^{(K_{\mathrm{a}})\mathrm{T}}\right)^{\mathrm{T}}. \tag{4}$$

Furthermore, the noise vectors $\underline{\mathbf{n}}^{(k_{\mathrm{a}})}$, $k_{\mathrm{a}} = 1 \ldots K_{\mathrm{a}}$, are combined to the total noise vector

$$\underline{\mathbf{n}} = \left(\underline{\mathbf{n}}^{(1)\mathrm{T}}, \underline{\mathbf{n}}^{(2)\mathrm{T}} \ldots \underline{\mathbf{n}}^{(K_{\mathrm{a}})\mathrm{T}}\right)^{\mathrm{T}} \tag{5}$$

of length $K_{\mathrm{a}} N$, and the received signal vectors $\underline{\mathbf{e}}^{(k_{\mathrm{a}})}$, $k_{\mathrm{a}} = 1 \ldots K_{\mathrm{a}}$, are arranged in the total received signal vector

$$\underline{\mathbf{e}} = \left(\underline{\mathbf{e}}^{(1)\mathrm{T}}, \underline{\mathbf{e}}^{(2)\mathrm{T}} \ldots \underline{\mathbf{e}}^{(K_{\mathrm{a}})\mathrm{T}}\right)^{\mathrm{T}} \tag{6}$$

of length $K_{\mathrm{a}} N$. Using (4), (5) and (6) the OFDM transmission can be expressed in matrix vector notation as follows

$$\underline{\mathbf{e}} = \underline{\mathbf{H}}_{\mathrm{z}} \underline{\mathbf{F}}^{-1} \underline{\mathbf{d}} + \underline{\mathbf{n}}\,. \tag{7}$$

C. Adaptive Signal Processing

Based on the linear transmission model of (7) the estimates $\hat{\underline{\mathbf{d}}} = \left(\hat{\underline{d}}_1, \hat{\underline{d}}_2 \ldots \hat{\underline{d}}_N\right)^{\mathrm{T}}$ of the transmitted data vector $\underline{\mathbf{d}}$ are obtained by solving the detection problem

$$\hat{\underline{\mathbf{d}}} = \underline{\mathbf{M}}\,\underline{\mathbf{e}} \tag{8}$$

using linear detection algorithms [11] which are represented by $\underline{\mathbf{M}}$ in (8). In order to specify the ZF equalizer the following notations are introduced: $\vec{\underline{\mathbf{h}}}^{(k_{\mathrm{a}})} = \underline{\mathbf{F}}\,\underline{\mathbf{h}}^{(k_{\mathrm{a}})}$ is the frequency response of channel k_{a}, $\vec{\underline{\mathbf{n}}}^{(k_{\mathrm{a}})} = \underline{\mathbf{F}}\,\underline{\mathbf{n}}^{(k_{\mathrm{a}})}$ and $\vec{\underline{\mathbf{e}}}^{(k_{\mathrm{a}})} = \underline{\mathbf{F}}\,\underline{\mathbf{e}}^{(k_{\mathrm{a}})}$ are the interference vector and the received signal vector of the k_{a}-th antenna in the frequency domain, respectively. The total interference vector $\vec{\underline{\mathbf{n}}}$ and the total received signal vector $\vec{\underline{\mathbf{e}}}$ are build according to (5) and (6), respectively. Furthermore, the matrix

$$\vec{\underline{\mathbf{H}}}_{\mathrm{z}}^{(k_{\mathrm{a}})} = \underline{\mathbf{F}}\,\underline{\mathbf{H}}_{\mathrm{z}}^{(k_{\mathrm{a}})} \underline{\mathbf{F}}^{-1} = \mathrm{diag}\left(\vec{\underline{h}}_1^{(k_{\mathrm{a}})}, \vec{\underline{h}}_2^{(k_{\mathrm{a}})} \ldots \vec{\underline{h}}_N^{(k_{\mathrm{a}})}\right) \tag{9}$$

represents the k_a-th channel in the frequency domain, and the respective total channel matrix is

$$\overleftrightarrow{\underline{\mathbf{H}}}_{\mathrm{z}} = \left(\overleftrightarrow{\underline{\mathbf{H}}}_{\mathrm{z}}^{(1)\mathrm{T}}, \overleftrightarrow{\underline{\mathbf{H}}}_{\mathrm{z}}^{(2)\mathrm{T}} \ldots \overleftrightarrow{\underline{\mathbf{H}}}_{\mathrm{z}}^{(K_\mathrm{a})\mathrm{T}} \right)^{\mathrm{T}} . \qquad (10)$$

Finally, after introducing the covariance matrix

$$\overleftrightarrow{\underline{\mathbf{R}}}_{\mathrm{n}} = \mathrm{E}\left\{ \underline{\vec{\mathbf{n}}}\, \underline{\vec{\mathbf{n}}}^{*\mathrm{T}} \right\}, \qquad (11)$$

the ZF equalizer for the considered OFDM system becomes

$$\hat{\underline{\mathbf{d}}} = \left(\overleftrightarrow{\underline{\mathbf{H}}}_{\mathrm{z}}^{*\mathrm{T}} \overleftrightarrow{\underline{\mathbf{R}}}_{\mathrm{n}}^{-1} \overleftrightarrow{\underline{\mathbf{H}}}_{\mathrm{z}} \right)^{-1} \overleftrightarrow{\underline{\mathbf{H}}}_{\mathrm{z}}^{*\mathrm{T}} \overleftrightarrow{\underline{\mathbf{R}}}_{\mathrm{n}}^{-1} \underline{\vec{\mathbf{e}}} . \qquad (12)$$

Since $\overleftrightarrow{\underline{\mathbf{H}}}_{\mathrm{z}}$ is composed of simple diagonal matrices, the detection problem of (12) may be simplified by considering $\overleftrightarrow{\underline{\mathbf{R}}}_{\mathrm{n}}$ in more detail. Utilizing

$$\overleftrightarrow{\underline{\mathbf{R}}}_{\mathrm{n}}^{(u,v)} = \mathrm{E}\left\{ \underline{\vec{\mathbf{n}}}^{(u)} \underline{\vec{\mathbf{n}}}^{(v)*\mathrm{T}} \right\}, \; u,v = 1 \ldots K_\mathrm{a} , \qquad (13)$$

$\overleftrightarrow{\underline{\mathbf{R}}}_{\mathrm{n}}$ can be expressed as

$$\overleftrightarrow{\underline{\mathbf{R}}}_{\mathrm{n}} = \begin{pmatrix} \overleftrightarrow{\underline{\mathbf{R}}}_{\mathrm{n}}^{(1,1)} & \overleftrightarrow{\underline{\mathbf{R}}}_{\mathrm{n}}^{(1,2)} & \cdots & \overleftrightarrow{\underline{\mathbf{R}}}_{\mathrm{n}}^{(1,K_\mathrm{a})} \\ \overleftrightarrow{\underline{\mathbf{R}}}_{\mathrm{n}}^{(2,1)} & \overleftrightarrow{\underline{\mathbf{R}}}_{\mathrm{n}}^{(2,2)} & \cdots & \overleftrightarrow{\underline{\mathbf{R}}}_{\mathrm{n}}^{(2,K_\mathrm{a})} \\ \vdots & \vdots & & \vdots \\ \overleftrightarrow{\underline{\mathbf{R}}}_{\mathrm{n}}^{(K_\mathrm{a},1)} & \overleftrightarrow{\underline{\mathbf{R}}}_{\mathrm{n}}^{(K_\mathrm{a},2)} & \cdots & \overleftrightarrow{\underline{\mathbf{R}}}_{\mathrm{n}}^{(K_\mathrm{a},K_\mathrm{a})} \end{pmatrix} . \qquad (14)$$

Assuming that the interference at different antennas is uncorrelated, only the main diagonal of (14) remains. Moreover, if the interference on different subcarriers is uncorrelated, then

$$\overleftrightarrow{\underline{\mathbf{R}}}_{\mathrm{n}}^{(k_\mathrm{a},k_\mathrm{a})} = \mathrm{diag}\left(\left|\underline{\vec{n}}_1^{(k_\mathrm{a})}\right|^2, \left|\underline{\vec{n}}_2^{(k_\mathrm{a})}\right|^2 \cdots \left|\underline{\vec{n}}_N^{(k_\mathrm{a})}\right|^2 \right), \quad k_\mathrm{a} = 1 \ldots K_\mathrm{a} , \qquad (15)$$

i.e. $\overleftrightarrow{\underline{\mathbf{R}}}_{\mathrm{n}}$ is also a diagonal matrix. Therefore, under the given assumptions the detection problem of (12) can be simplified, and with the SNRs

$$?_i^{(k_\mathrm{a})} = \left|\underline{\vec{h}}_i^{(k_\mathrm{a})}\right|^2 \Big/ \mathrm{E}\left\{ \left|\underline{\vec{n}}_i^{(k_\mathrm{a})}\right|^2 \right\}, \; k_\mathrm{a} = 1 \ldots K_\mathrm{a} , \qquad (16)$$

at the K_a antenna elements the estimates $\hat{\underline{d}}_i$, $i = 1 \ldots N$, of $\hat{\underline{\mathbf{d}}}$ are given by

$$\hat{\underline{\mathbf{d}}}_{i,\mathrm{ZF}} = \frac{\sum_{k_\mathrm{a}=1}^{K_\mathrm{a}} ?_i^{(k_\mathrm{a})} \underline{\vec{h}}_i^{(k_\mathrm{a})-1} \underline{\vec{e}}_i^{(k_\mathrm{a})}}{\sum_{k_\mathrm{a}=1}^{K_\mathrm{a}} ?_i^{(k_\mathrm{a})}} , \; i = 1 \ldots N , \qquad (17)$$

independently of each other, i.e. a subcarrier-wise detection is possible.

D. SIMULATION RESULTS

The performance of the adaptive antenna concept is evaluated at the link level in terms of the PDU (Protocol Data Unit) Error Rate (PER) vs. Signal-to-Noise-Ratio (SNR). The considered PDU type is the Long Channel (LCH) and the filters, the channel estimation as well as the frame- and frequency synchronization are assumed to be ideal. The HIPERLAN/2 channel models A (typical office) and E (typical large open space), developed by ETSI BRAN, are applied. Furthermore, the channel impulse response (CIR) is assumed to be time invariant within a burst. HIPERLAN/2 provides seven transmission modes. In this evaluation only the most robust mode (BPSK, code rate $R_\mathrm{c} = 1/2$, data rate $R_\mathrm{d} = 6\mathrm{Mbps}$) and the least robust mode (64 QAM, $R_\mathrm{c} = 3/4$, $R_\mathrm{d} = 54\mathrm{Mbps}$) are considered.

In Fig. 2, which shows the results for channel model A the gain by doubling the number of antennas from one to eight is 7 dB, 6.5 dB and 4 dB in the case of 54 Mbps and 5.7 dB, 5.2 dB and 4.5 dB in the case of 6 Mbps at the target PER of 10^{-2}. These high gains are explained as follows: Doubling the number of antennas yields an SNR increase of 3 dB, since the noise at different antennas is uncorrelated. Additionally, frequency diversity is provided on the particular subcarrier, since the different channels are uncorrelated. The achieved gain due to frequency diversity is higher for the 54 Mbps mode than for the 6 Mbps mode, since the capability of this mode to correct errors on weak subcarriers by Forward Error Correction (FEC) is relatively low. Fig. 3 shows the results for channel model E, which provides multipath diversity, but also introduces intersymbol interference (ISI) and, therefore, intercarrier interference (ICI) due to the large delay spread. For the 54 Mbps mode the degradation by the ICI is higher than the gain due to multipath diversity. Therefore, this mode performs worse than with model A. In contrast to that, the 6 Mbps mode performs better, since the ICI is negligible at the low SNRs and only the benefit of multipath diversity is visible. In Fig. 3 in the case of 6 Mbps the achieved gain by doubling the number of antennas is 1-2 dB smaller than in Fig. 2, because the multipath diversity also provides some frequency diversity. With increasing number of antennas the benefits of multipath diversity decrease. In the case of the 54 Mbps mode considered in Fig. 3 the achieved gain by doubling the number of antennas is higher than in the case of Fig. 2, since the ICI is also reduced by the adaptive antenna concept. With increasing number of antennas the degradation caused by the ICI decreases.

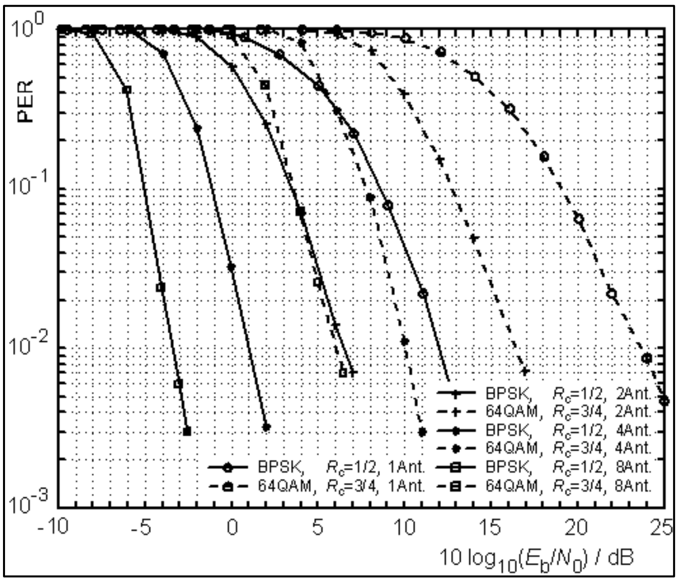

Fig. 2: PER vs. SNR for $K_a = \{1, 2, 4, 8\}$ antennas with $R_d = \{6, 54\}$ Mbps; channel model A

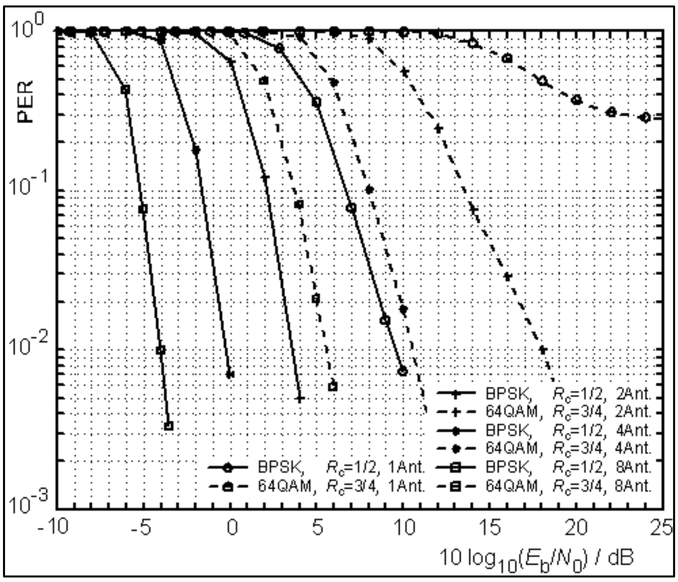

Fig. 3: PER vs. SNR for $K_a = \{1, 2, 4, 8\}$ antennas with $R_d = \{6, 54\}$; channel model E

E. CONCLUSIONS

It has been shown that the considered adaptive antenna concept improves the link level performance of HIPERLAN/2 significantly. Using only two antennas enables the usage of the 54 Mbps mode with channel model E, which would be impossible with a single antenna. However, the presented results are upper bounds due to the ideal assumptions. Nevertheless, justified by the high maximum gain, the authors expect that this adaptive antenna concept very significantly improves also the performance of a real HIPERLAN/2 system.

F. ACKNOLEDGEMENT

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