Space-time Equalization with Smart Antenna and Hard Limiter Receiver

Frank Althaus
CTL ETHZ
CH-8092 Zuerich
Telephone: +41 1 632 6918
Fax: +41 1 632 1209
Email: althaus@nari.ee.ethz.ch

Michael Kuhn
Ascom Systec AR&T
CH-5506 Maegenwil
Telephone: +41 62 889 5248
Fax: +41 62 889 5290
Email: michael.kuhn@ascom.ch

Armin Wittneben
CTL ETHZ
CH-8092 Zuerich
Telephone: +41 1 632 3611
Fax: +41 1 632 1209
Email: wittneben@nari.ee.ethz.ch

***Abstract*— Both, adaptive antenna arrays for ISI cancellation and non-linear hard limiter receivers are well known as key technologies for high data rate low-cost receivers [1][2]. However, the combination of both technologies is a big challenge since non-linear receivers only provide phase or frequency samples. The amplitude information, which is required for the adaptation of the antenna array, is not accessible. Therefore, adaptive antenna arrays are most often applied in combination with linear receivers.**

In this paper we present a new very low cost receiver which combines adaptive antenna arrays and hard limiter receivers. The training of the adaptive antenna array is performed on basis of the complex channel impulse response (CIR), which is estimated only by phase samples. The performance of this approach is shown by means of simulations using a simple hard decision device and Maximum Likelihood equalization. Three decoding structures are compared: Optimum combining with hard decision, optimum combining with reduced state Viterbi equalizer and a single antenna receiver with full state Viterbi equalizer.

I. INTRODUCTION

Hard amplitude limiting receiver structures are known as very low cost receivers. Compared to linear receivers power consumption and hardware effort can be reduced immensely. Due to the amplitude limitation the dynamic range of those systems is minimized and no automatic gain control (AGC) circuit is required. However, channel equalizing in digital domain is difficult for non-linear receivers since the amplitude information is not available. Thus, non-linear receivers are usually clearly outperformed by linear receivers, especially in severe frequency selective mobile communication channels. Smart antennas are known as low cost technology for intersymbol interference (ISI) cancellation in the analogue domain [2][3][4]. Up to now, the training of the adaptive antenna array (Combiner) is performed using linear receiver structures because an estimate of the complex channel impulse response (CIR) is required.

In this paper a method to calculate the combiner coefficients on basis of phase samples is shown. Thus, a new receiver structure, namely combiner with cascaded phase detector, is able to compete even severe ISI situations (Fig. 1). The hard limiter is inherently included because equalization and channel estimation are based only on phase samples. The complex weighting coefficients of the adaptive antenna array are represented by c_i with $i = 1..N_A$ where N_A equals the number of antennas in the receiver. The receive signals are amplified (or damped), phase shifted and summed up in the analogue domain. Thus, only one receive chain is required. The resulting signal is filtered by a bandpass prior to phase detection and sampling. An equalizer can be cascaded to obtain the estimated information symbols $\vec{\alpha}$. In combination with a Viterbi decoder a new class of space-time equalizers with different degrees of complexity can be deduced. The performance of these equalizers is shown by means of simulations.

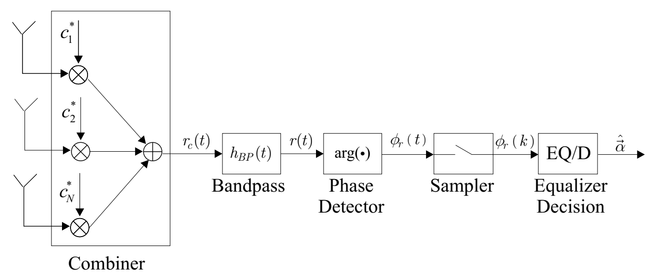

Fig. 1. Receiver model

II. SYSTEM MODEL

Figure 2 shows a block diagram of the discrete system model. Phase modulation of the input symbols $\vec{\alpha}$, $\vec{\alpha}[\nu] \in$

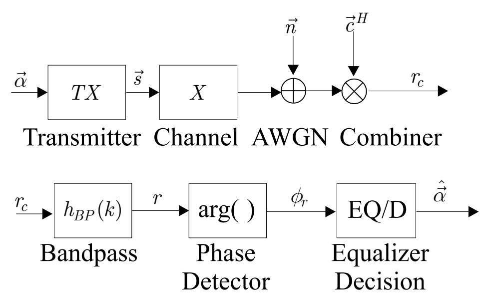

Fig. 2. Discrete system model

$\{-1, 1\}$, yields the transmit sequence $\vec{s}$. The rows of the channel matrix X describe the N_A multipath propagation channels from the transmit antenna to the receive antennas. The combined signal r_c can be written as:

$$r_c = \vec{c}^H \cdot (X\vec{s} + \vec{n}) \tag{1}$$

$\vec{n}$ is an additive symmetric white Gaussian noise (AWGN) component with two sided spectral power density N_0. The resulting impulse response of channel matrix and combiner is:

$$\vec{h}_c^T = \vec{c}^H \cdot X \tag{2}$$

The received signal behind the bandpass filter $h_{BP}(k)$ is denoted r and ϕ_r is the phase of the received signal. The phase detector *arg(·)* is assumed to provide a $\mod 2\pi$ distributed phase signal.

Two types of decoders are investigated: a differential phase threshold detector, where the ν-th symbol is decided with

$$\hat{\vec{\alpha}}[\nu] = \text{sign}\left\{\vec{\phi}_r[\nu] - \vec{\phi}_r[\nu-1]\right\} \tag{3}$$

and a Viterbi decoder which performs a Maximum Likelihood sequence estimation [7]:

$$\begin{aligned} \hat{i} &= \arg\{\max_i p(\vec{\phi}_r \mid \vec{d}^{(i)})\} \\ \hat{\vec{\alpha}} &= \vec{\alpha}^{(i)} \end{aligned} \tag{4}$$

$p(\vec{\phi}_r \mid \vec{d})$ is the conditional probability density function and $\vec{d}$ is the complex desired sequence.

III. Combining

In this paper we investigate optimum combining (OC) and examine the impact on the system performance in combination with the cascaded equalizer. OC maximizes the signal-to-interference-plus-noise ratio $S/(I+N)$. The combiner coefficients are given as:

$$\vec{c} = \left(XX^{\mathrm{H}} + I\sigma_n^2\right)^{-1} \cdot X[:, k_{opt}] \tag{5}$$

σ_n^2 is the variance of the symmetric complex additive white Gaussian noise, I is the unity matrix and $X[:, k_{opt}]$ means the k_{opt} column of the channel matrix. The parameter k_{opt} is a synchronization parameter which selects the signal tap. All other taps are considered as interference and are suppressed.

The combiner coefficient optimization requires an estimate of the complex channel impulse matrix X for the adaptation of the antenna array. Since the channel impulse response is generally unknown in wireless transmission systems it has to be estimated.

IV. Maximum Likelihood Sequence Estimation

The Maximum Likelihood sequence estimation (MLSE) is performed using some idealized assumptions: The bandpass is supposed to be an ideal bandpass filter. This means the noise samples to be statistically independent and the phase errors to be caused by an AWGN component. Furthermore, the knowledge of the complex desired sequence $\vec{d}$ is assumed . Then, the probability density function in (4) can be factorized:

$$\begin{aligned} &\max_l p\left(\vec{\phi}_r \middle| \vec{d}^{(l)}\right) \\ &= \max_l \prod_{k=1}^{K} p\left(\vec{\phi}_r[k] \middle| \vec{d}^{(l)}[k]\right) \\ &= \max_l \prod_{k=1}^{K} p\left(\overrightarrow{\Delta\phi}^{(l)}[k] \middle| |\vec{d}^{(l)}[k]|\right) \end{aligned} \tag{6}$$

with the phase error $\overrightarrow{\Delta\phi}^{(l)}[k] = \vec{\phi}_r[k] - \vec{\phi}_d^{(l)}[k]$. The probability density $p\left(\Delta\phi^{(l)} \middle| |d^{(l)}|\right)$ is given by Pawula in [5]:

$$\begin{aligned} p_{\Delta\phi|\rho}(\Delta\phi|\rho) = \frac{\exp(-\rho)}{2\pi} + \sqrt{\frac{\rho}{4\pi}} \exp\left(-\rho \sin^2(\Delta\phi)\right) \\ \cdot \cos(\Delta\phi) \cdot \operatorname{erfc}\left(-\sqrt{\rho}\cos(\Delta\phi)\right) \end{aligned} \tag{7}$$

$$\rho = \frac{|d|^2}{\sigma_n^2} \tag{8}$$

With (6) and (7) the MLSE can be evaluated using the Viterbi algorithm [6].

V. Channel estimation

Wireless LAN standards (e.g. HiperLAN) often apply a known training sequence at the beginning of a data burst for CIR estimation and synchronization. Non-linear hard limiter receivers can estimate the complex CIR on basis of phase samples of this training sequence as shown in [7]: The training sequence (TS) consists of M consecutive m-sequences. This leads to an M fold observation of the same desired phase sequence $\vec{\phi}_d$:

$$\vec{\phi}_r^{(m)} = \vec{\phi}_d + \vec{\phi}_n^{(m)}\ , m = 1...M \tag{9}$$

A phasor average yields an estimate $\hat{\vec{\phi}}_d$. The associated amplitude sequence $|\vec{d}|$ is estimated based on the variance of the observed phase values. These estimates yield a first estimate $\hat{\vec{d}}_1$ of the complex desired sequence:

$$\hat{\vec{d}}_1 = |\hat{\vec{d}}| \cdot \exp(j\hat{\vec{\phi}}_d) \tag{10}$$

A subspace approach yields an estimate $\hat{\vec{h}}_c$ of the complex CIR:

$$\hat{\vec{h}}_c = \arg\left\{\min_{\vec{h}_c} \left|\mathbf{S}_{BP}\vec{h}_c - \hat{\vec{d}}_1\right|^2\right\} \tag{11}$$

$\mathbf{S}_{BP}$ is the convolution matrix of the transmitted training sequence and the bandpass impulse response. The CIR estimation procedure can easily be adapted to the N_A antenna receiver by switching the antennas of the antenna array. Note, the training sequence must be transmitted N_A times.
For Viterbi decoding according to (4) the complex sequence $\vec{d}$ is required. A second, more precise estimate $\overset{\leadsto}{d}_2$ is given by:

$$\overset{\leadsto}{d}_2 = \mathbf{S}_{BP}\hat{\vec{h}}_c \tag{12}$$

VI. Results

Three receiver structures are considered here. First, a one antenna phase detection receiver without combiner and with full state Viterbi decoder is examined. The number of states is determined by the overall system memory which is given by the memory of the bandpass, channel and transmitter. With the MLSE approach from (4) this is a quasi optimum decoder for the phase detection receiver.
The second structure is a three antenna receiver ($N_A = 3$) with combiner. Optimum combining is used to combat the ISI of the propagation channel. A differential phase detector according to (3) is used for the decision of the symbols. However, this structure suffers from the ISI introduced by the bandpass filter which cannot be combated by the combiner.
Therefore, the third structure uses combiner and phase detection with Viterbi decoder to equalize this ISI. The number of states is reduced compared to the first structure and this way the complexity is reduced.

The performance of these receiver structures are examined by bit error ratio (BER) simulations. The curves are shown over E_b/N_0 which is energy per bit to noise power density. For the three antenna receivers this is the mean E_b/N_0 of all branches. Here, MSK modulation and $N_A = 3$ antennas are investigated. (The Trellis representation of MSK can be found in [6].) The bandpass is a second order butterworth filter with $B \cdot T = 0.4$; B the one sided 3dB-bandwidth.
First the performance for an exemplary channel matrix X

$$X = \begin{bmatrix} 1 & 0 & 1 \\ 1 & -1 & 0 \\ 0 & 1 & -1 \end{bmatrix}$$

is examined and compared with the performance of state-of-the-art decoders. For a single antenna differential phase detection (*DPD*) receiver all three channels are very critical. Fig. 3 shows the simulation results for the DPD with Viterbi decoder and Euclidean metric (*DPD+Viterbi*) as described in [1]. Obviously, this receiver fails and also simpler structures without equalizer or with decision feedback equalizer fail in these channels. The reason for that

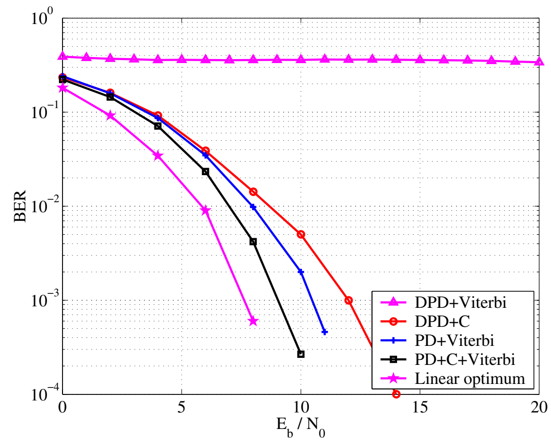

Fig. 3. Performance of the proposed receivers with perfect CIR knowledge.

are high amplitude ¤uctuations caused by the propagation channel. The special choice of X allows a very good result of the combiner combating the ISI. With $\vec{c}^H = [1, 1, 1]$ the resulting impulse response equals an ideal propagation channel:

$$\vec{h}_c^T = \vec{c}^H \cdot X = [2, 0, 0] \tag{13}$$

The results in Fig. 3 are achieved for this combining vector which also implies a perfect channel estimate. As expected the DPD with combiner (*DPD+C*) achieves good performance. The single antenna receiver with full state Viterbi and phase detection (*PD+Viterbi*) achieves better performance: about 1.5dB at a BER of 10^{-3}. Note, using 3 antennas would increase the performance due to the diversity gain. The gain of the reduced state Viterbi (*PD+C+Viterbi*) is again 1.5dB at a BER of 10^{-3}. As a reference the performance of an optimum linear receiver with one antenna (*Linear optimum*) is also shown. It achieves a gain of about 1.5dB compared to the best phase detection receiver.
The impact of the nonlinear channel estimation (according to Section V) is shown in Fig. 4. The CIR estimates are performed for each antenna separately by switching the antennas. This means a 3 times longer training sequence (TS) than for the single antenna system. With the system memory L the length of the training sequence (TS) in times of the symbol period is:

$$N_A \cdot M \cdot 2^{L+1} \tag{14}$$

In the simulations here $L = 4$ and $M = 5$is used and the length of the TS results to 480 symbols. Since T-spaced sampling is used at the receiver the number of samples for the channel estimate is also 480 which means 160 samples per antenna. The results show a loss of only about 1dB and below for BER $\leq 10^{-3}$.
Experiences have shown that the quality of the CIR estimate is essentially determined by the number of samples of the TS. So, with a higher sampling rate the performance

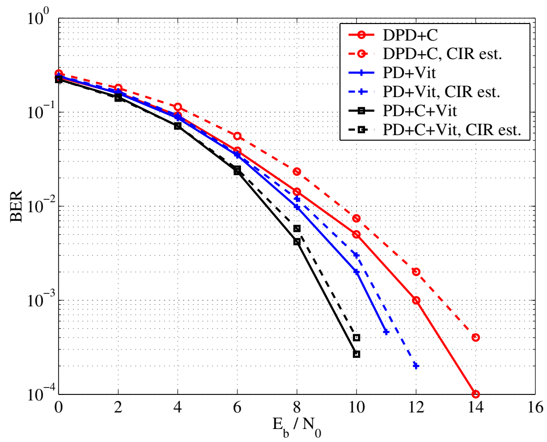

Fig. 4. Performance of the proposed receivers with estimated CIR.

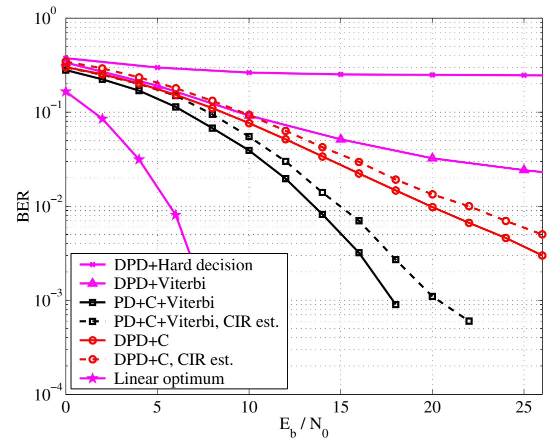

Fig. 5. Performance of the proposed receivers for channel model with exponential power delay profile.

of the receiver can be increased or otherwise the length of the TS can be reduced without giving up the performance.

Finally, a more general channel model is used to examine the performance of the receiver structures with combiner. The taps of the channel impulse response are modeled as randomly generated complex numbers with Rayleigh distributed amplitudes A and uniformly distributed phases ϕ. The channel length is chosen to have 3 taps. The delay spread DS is set to $0.5T$. With the symbol period T the taps of the channel matrix are given:

$$X[k,\nu] = A_{k,\nu}\exp(j\phi_{k,\nu})\sqrt{\exp\left(-\frac{\nu}{\mathrm{DS}/T}\right)} \qquad (15)$$

Fig. 5 shows the mean BER of this channel model evaluated for 200 channels at each E_b/N_0. Again the combiner with Viterbi outperforms the simpler structure. The loss of the channel estimation compared to the BER with perfect channel knowledge is up to 2dB for BER $\leq 10^{-3}$. The performance of a linear optimum receiver and a differential phase detection receiver are shown as references. Of course, compared to the complex optimum linear receiver the performance of any phase detection receiver must be worse. However, differential phase detection with antenna combiner is a very low cost receiver structure that provides very useful capabilities also in severe ISI propagation. Compared to common nonlinear receiver structures with differential phase detection or limiter discriminator detection the coverage of the phase detector with combiner will be much higher. As shown in the plot the differential phase detector fails in these ISI channels. Also, with additional (full state) Viterbi equalizer only a small performance improvement can be achieved. However, the combiner structures –in particular with reduced state Viterbi– outperform these common structures. Therefore, they could be an attractive alternative to common expensive linear receivers.

References

[1] J. P. Fonseka. Noncoherent detection with viterbi decoding for gmsk signals. *IEE Proc.-Commun.*, 143(6):373–378, December 1996.

[2] A. Wittneben and U. Dettmar. A low cost noncoherent receiver with adaptive antenna combining for high speed wireless lans. *Vehicular Technology Conference*, 1997.

[3] M. Kuhn. *Smart Antennas as Key Technology for ISI-Cancellation in Low-Cost/Low-Power Receivers*. PHD Thesis, Universitaet des Saarlandes, 2002.

[4] M. Kuhn and A. Wittneben. Very low complexity space-time decision feedback equalization of intersymbol interference on severe frequency-selective fading channels. *Vehicular Technology Conference Spring 2002 (in press)*, 2002.

[5] R. F. Pawula, S. O. Rice, and J. H. Roberts. Distribution of the phase angle between two vectors perturbed by gaussian noise. *IEEE Transactions on Communications*, 30(8):1828–1841, August 1982.

[6] J. Anderson, T. Aulin, and C.-E. Sundberg. *Digital Phase Modulation*. Plenum Press, New York, 1986.

[7] Frank Althaus. *A New Low Cost Approach to Wireless Communication Over Severe Multipath Fading Channels*. Dissertation thesis, in press, Universitaet des Saarlandes, 2002.