PATH-DIVERSITY FOR PHASE DETECTION IN LOW-COST SENSOR NETWORKS

Frank Althaus, Thomas Zasowski and Armin Wittneben

Swiss Federal Institute of Technology (ETH) Zurich
Communication Technology Laboratory, Sternwartstrasse 7, CH-8092 Zuerich
Email: {althaus, zasowski, wittneben}@nari.ee.ethz.ch

ABSTRACT

Wireless sensor networks are composed by a high number of very simple nodes. A paradigm for the nodes transceiver technology is to be energy aware and cost conscious. In this paper we consider a nonlinear transceiver structure which is a key technology for low-cost and power efficient network nodes. We use a phase detection receiver with a branch dependent metric which benefits from the knowledge of the channel fading coefficients. Therefore we introduce a method to estimate the complex coefficients based on samples of the phase detector output. Since most routing concepts for sensor networks envisage multipath routing with simple forwarding strategies (e.g. flooding), it is not far from that to make use of path diversity techniques in order to improve the network reliability. We show that remarkable diversity gains can be achieved also with the nonlinear receiver structure. By means of simulations we examine path diversity techniques for 2PSK, 4PSK and 8PSK modulation where we make use of common transmit diversity schemes. For 2PSK we propose a slight modification of the Alamouti scheme [1] to improve the performance.

1. INTRODUCTION

WIRELESS sensor networks are characterized by a high density of sensor nodes, each providing sensor information which must be communicated through the network. In contrast to many other mobile ad-hoc network topologies there is no access point which can be reached by all nodes and which manages communication. In sensor networks the information is carried across the network via multi-hop links over several paths and many nodes to a certain data sink (Fig. 1). Therefore, sensor nodes have transmitter and receiver capabilities to be able to forward data packets. However, links in sensor networks are unreliable for several reasons: In general sensor nodes are located randomly and/or are mobile. Geographic distances and obstacles between the nodes can decrease the communications performance. A very important point is the energy consumption of the nodes. Since the power supply of the nodes is provided by batteries their lifetime is limited and with it the lifetime of links and the network. When the battery of a node is depleted this node cannot longer communicate and all network paths passing this node fail. Mostly the batteries cannot be replaced or recharged and nodes with empty batteries will be dumped. Hence, wireless technology for sensor nodes must be extremely cost conscious and energy aware. In the next section we introduce a wireless transceiver structure of highest simplicity, power efficiency and cost consciousness. The nonlinear receiver structure is based on a phase detector. In Section 3 we propose a branch dependent metric which uses the complex channel coefficients. In Section 4 we introduce a technique to estimate the complex channel impulse response based on the phase detector output. This method simplifies for the memoryless channel coefficients which are assumed in Section 5 where we examine path diversity techniques with phase detection receiver. For some specified assumptions this path diversity problem converges to a simple transmit diversity problem which is familiar in literature.

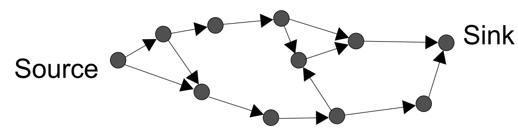

Fig. 1. Sensor network

2. SENSOR NODE TRANSCEIVER STRUCTURE

Since the lifetime of a sensor node depends directly on its power consumption, power efficient transceiver architectures are very important for sensor nodes [2][3]. The main power consumers of the transceiver are the transmit power amplifier stages. Using power efficient hard limiting amplifiers is a paradigm for an energy aware transmitter technology. This implies using (nonlinear) constant envelope modulation schemes as FSK and CPM. Fig. 2 shows the structure of the transceiver where the transmitter block is characterized more general by a nonlinear power amplifier. For constant envelope signals the most appropriate receiver structure is a hard limiter receiver with phase or frequency detection [4]. This nonlinear structure is known as very simple and power efficient because requirements to linearity and dynamic range leading to the high complexity of linear structures can be dropped. Furthermore, costly automatic gain control circuits (AGC) can also be dropped. Since amplitude information is lost with this receiver type only phase detectors, differential phase detectors and frequency detectors can be used for a data decoding. In Fig. 2 this is indicated by a phase detector and a subsequent filter which can perform a signal integration to get frequency or differential phase detection output. In this paper we focus on one special case of the transceiver structure. The model of the considered phase detection receiver is shown in Fig 3.

To forward data packets the input data must be stored. Transmitting and receiving simultaneously is not possible. A very simple procedure is ”analog store and forwarding”. In this case the analog data is retransmitted without decoding them first. This scheme reduces power consumption for the forwarding to a minimum. The signal processing block in Fig. 2 is characterized by a

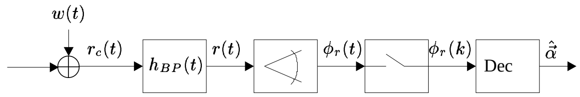

Fig. 3. Phase detection receiver

time delay which is representative for the forwarding mode of the node.

3. BRANCH DEPENDENT METRIC FOR PHASE DETECTION

We propose the use of a Maximum Likelihood Sequence Estimator (MLSE) with a branch dependent metric for the phase detection receiver in Fig. 3. The complex channel coefficients are known from the estimation process (Section 4). Therefore, the possible complex desired sequences

$$\vec{d}^{(l)} = |\vec{d}|^{(l)} \cdot \exp(\phi_d^{(l)}) \tag{1}$$

are known at the decoder. This knowledge allows to formulate the following MLSE approach:

$$\begin{aligned} \hat{l} &= \arg\{\max_l p(\vec{\phi_r} \mid \vec{d}^{(l)})\} \\ \hat{\vec{\alpha}} &= \vec{\alpha}^{(\hat{l})} \end{aligned} \tag{2}$$

$p(\vec{\phi_r} \mid \vec{d})$ is the conditional probability density function and $\vec{d}$ is the complex desired sequence. Since the metric function of $\vec{\phi_r}$ and $\vec{\phi_d}$ depends on the amplitude $\vec{d}^{(l)}$ it differs in the l branches and the metric is denoted *branch dependent*. The 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. The probability density function in (2) can be factorized:

$$\begin{aligned} &\max_l p\left(\vec{\phi}_r \left| \vec{d}^{(l)}\right.\right) \\ &= \max_l \prod_{k=1}^{K} p\left(\vec{\phi}_r[k] \left| \vec{d}^{(l)}[k]\right.\right) \\ &= \max_l \prod_{k=1}^{K} p\left(\overrightarrow{\Delta\phi}^{(l)}[k] \left| |\vec{d}^{(l)}[k]|\right.\right) \end{aligned} \tag{3}$$

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

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

with the instantaneous signal-to-noise ratio

$$\rho = \frac{|d|^2}{\sigma_n^2} \tag{5}$$

In Section 6 we compare this metric approach with the best known metric for phase detection as proposed in [6], where (4) is used as the metric where ρ is assumed to be constant. That is why this metric is branch independent.

4. ESTIMATION OF THE CHANNEL FADING COEFFICIENTS

Nonlinear hard limiter receivers can estimate the complex channel impulse response (CIR) on basis of phase samples from an appropriate training sequence [7]: The training sequence (TS) consists of M consecutive m-sequences. This leads to an M fold observation of the same desired phase sequence $\vec{\phi}_d$:

$$\vec{\phi}_r^{(m)} = \vec{\phi}_d + \overrightarrow{\Delta\phi}_n^{(m)}, \; m = 1...M \tag{6}$$

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. Both estimates are used to achieve a first estimate $\hat{\vec{d}}_1$ of the complex desired sequence:

$$\hat{\vec{d}}_1 = |\hat{\vec{d}}| \cdot \exp(j\hat{\vec{\phi}}_d) \tag{7}$$

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

$$\hat{\vec{h}}_c = \arg\left\{\min_{\vec{h}_c} \left|\mathbf{S}_{BP}\vec{h}_c - \hat{\vec{d}}_1\right|^2\right\} \tag{8}$$

$\mathbf{S}_{BP}$ is the convolution matrix of the transmitted training sequence and the bandpass impulse response. (8) can be solved using QR-decomposition which finally leads to a linear equation system [8].

In flat fading the impulse response $\vec{h}_c$ reduces to a complex scalar h_c and $\mathbf{S}_{BP}$ reduces to a vector $\vec{s}_{BP}$.

$$\begin{aligned} \hat{h}_c &= \arg\left\{\min_{h_c} \left|\vec{s}_{BP}h_c - \hat{\vec{d}}_1\right|^2\right\} \\ &= \arg\left\{\min_{h_c} \left(\vec{s}_{BP}h_c - \hat{\vec{d}}_1\right) \cdot (\vec{s}_{BP}h_c - \hat{\vec{d}}_1)^H\right\} \end{aligned} \tag{9}$$

where H denotes the conjugate complex transposed. Performing the derivation $\frac{d}{dh_c}$ we get the following condition for h_c

$$\begin{aligned} \vec{s}_{BP}h_c^H\vec{s}_{BP}^H - \vec{s}_{BP}\hat{\vec{d}}_1 &= 0 \\ \Leftrightarrow h_c^H\vec{s}_{BP}^H - \hat{\vec{d}}_1 &= 0 \\ \Leftrightarrow h_c^H &= (\vec{s}_{BP}^H)^{-1}\hat{\vec{d}}_1^H \\ \Rightarrow h_c &= (\hat{\vec{d}}_1^H(\vec{s}_{BP}^H)^{-1})^H \\ \Leftrightarrow h_c &= \vec{s}_{BP}^{-1}\hat{\vec{d}}_1 \end{aligned} \tag{10}$$

With $\vec{s}_{BP}^{-1}$ meaning the pseudo inverse of $\vec{s}_{BP}$ we get

$$h_c = \frac{1}{|\vec{s}_{BP}|^2}\vec{s}_{BP}^H\hat{\vec{d}} \tag{11}$$

5. PATH AND TRANSMIT DIVERSITY WITH PHASE DETECTION

As mentioned in Section 1 links in sensor networks are often unreliable. To improve the reliability we investigate path diversity schemes for sensor nodes with the simple transceiver structure described in Section 2. Path diversity means that a data packet is transmitted to several neighboring (or inter-located) nodes which simultaneously forward this packet to a certain receiver node. Fig. 4 shows this scheme for the case of two inter-located nodes. Since

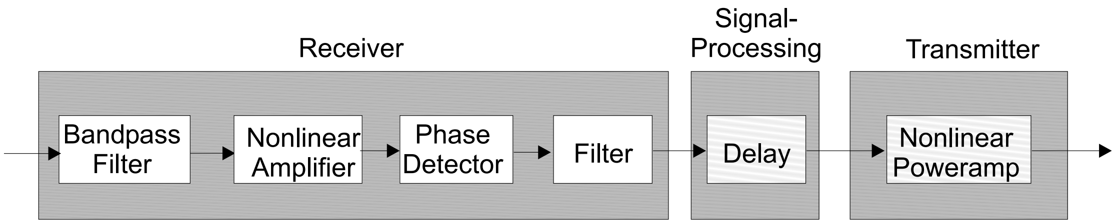

Fig. 2. Transceiver structure

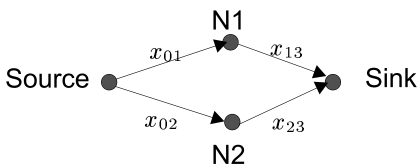

Fig. 4. Path diversity

the fading coefficients of both paths are different a diversity gain can be achieved. This path diversity is similar to transmit diversity if we neglect the noise component of the inter-located node. Then, the transmit antennas are located at different neighbors (macro diversity). For a simple analog forwarding scheme the fading coefficient of each path results from the fading coefficients of the two hops and from the amplification gain of the forwarder:

$$\begin{aligned} h_1 &= x_{01} \cdot \alpha_1 \cdot x_{13} \\ h_2 &= x_{02} \cdot \alpha_2 \cdot x_{23} \end{aligned} \quad (12)$$

If we assume that the attenuation of the first hop is compensated by the amplification:

$$|x_{0i}| \cdot |\alpha_i| = 1 \quad (13)$$

the resulting fading coefficient h_i is only a phase shifted form of x_{i3}:

$$h_i = x_{i3} \cdot e^{j\phi_i} \quad (14)$$

On the other hand if the nodes decode the data packet we assume that the fading of the first hop is compensated by the amplification

$$x_{0i} \cdot \alpha_i = 1 \quad (15)$$

and the resulting fading coefficients are given by:

$$\begin{aligned} h_1 &= x_{13} \\ h_2 &= x_{23} \end{aligned} \quad (16)$$

We assume the coefficients to be statistically independent Rayleigh fading variables. Then, the path diversity problem is identical to the transmit diversity problem which is very familiar in literature. However, the nonlinear structure of the receiver forbids a direct transfer of these results to our problem. Therefore, in this paper we investigate the performance of transmit diversity schemes for nonlinear reception.

We compare the performance of different transmit diversity schemes for PSK modulation and phase detection receiver. Therefore we consider Alamouti's transmit diversity scheme [1] and Wittneben's delay diversity scheme [9].

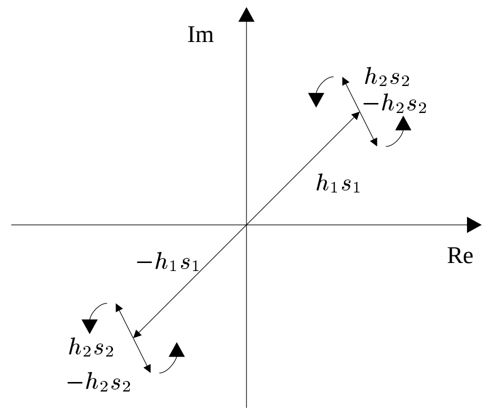

Fig. 5. Possible receive signals using Alamouti scheme with 2 PSK

5.1. Alamouti scheme

In [1] Alamouti presented a transmit diversity technique for systems with two transmit antennas and one receive antenna that achieves full diversity. The transmit signals s_1 and s_2 will be sent alternately on both transmit antennas according to (17). At time t_1 the symbol s_1 is sent on antenna T_{x1} and the symbol s_2 is sent on antenna T_{x2}, at time t_2 the symbol $-s_2^*$ is sent on antenna T_{x1} and the symbol s_1^* on antenna T_{x2}. Then the receive signal vector $\vec{r} = [r_1,\, r_2]^T$ can be written as

$$\vec{r} = \begin{bmatrix} s_1 & -s_2^* \\ s_2 & s_1^* \end{bmatrix}^T \cdot \begin{bmatrix} h_1 \\ h_2 \end{bmatrix} + \begin{bmatrix} w_1 \\ w_2 \end{bmatrix} \quad (17)$$

h_1 and h_2 denote the Rayleigh fading coefficients between the two transmit antennas and the receive antenna. It is assumed that h_1 and h_2 are constant during two time steps and mutually independent. w_1 and w_2 denote the complex additive white Gaussian noise components. s_1^* denotes the conjugate complex of s_1.

To get an insight in the problem of transmit diversity with phase detection we consider critical channel conditions. Fig. 5 illustrates the possible received signals for a certain h_1 and $|h_2| < |h_1|$ assuming a 2PSK signal constellation. Rotating the phase of h_2 we can figure out the critical constellations for phase detection. Obviously, critical constellations are given for $\phi_{h2} = \phi_{h1}$ and $\phi_{h2} = \phi_{h1} + \pi$, respectively ($h_i = |h_i| \cdot \exp(j\phi_{hi})$). For these constellations the phase differences disappear and extreme values of the amplitude appear. This means very small amplitudes are also possible.

These effects are demonstrated by means of an example. We assume $s_1 = 1$, $\phi_{h2} = \phi_{h1}$ and $|h_1| \geqslant |h_2|$. Using these parameters we get the receive signals for the time steps t_1 and t_2 displayed

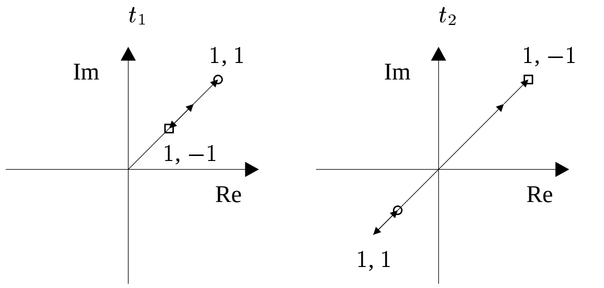

Fig. 6. Receive signals using Alamouti scheme for $s_1 = 1$ and $s_2 = \pm 1$

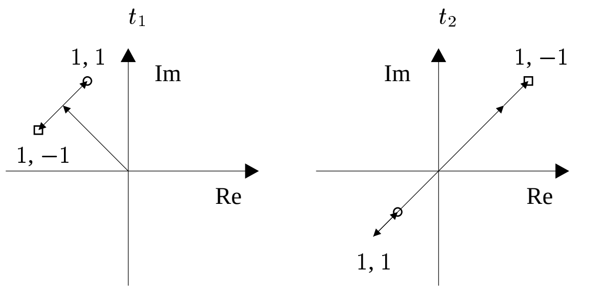

Fig. 7. Receive signals using modified Alamouti scheme for $s_1 = 1$ and $s_2 = \pm 1$

in Fig. 6. At time step t_1 the different receive symbols have the same phase angles and cannot be distinguished by the phase detector. If we regard time slot t_2 the phase detector is able to distinguish the receive signals. But if we assume that $|h_2|$ approaches $|h_1|$ the amplitude of the receive signal for $s_1 = 1; s_2 = 1$ converges to zero and therefore the phase detector performs worse. So it is obvious that the AWGN channel demonstrates a worst case scenario for the Alamouti scheme with phase detection.

These effects can be reduced by rotating transmit symbol s_1 at time t_1 by $\frac{\pi}{2}$. With this modification we get receive signals as shown in Fig. 7. If we assume that $|h_2|$ approaches $|h_1|$ the phase difference for different transmit symbol constellations approaches $\frac{\pi}{2}$ and the amplitude approaches $\sqrt{|h_1|^2 + |h_2|^2}$. At time t_1, $h_2 s_2$ is orthogonal to $h_1 s_1$ and the phase difference gets its maximum. This modified scheme shows two effects:

- The phase difference is increased,
- the amplitude obtains a positive interference i.d. $|r_1| > |h_1 s_1|$ and so the smallest possible amplitudes are higher than before.

The modified Alamouti scheme is shown in (18).

$$\vec{r} = \begin{bmatrix} e^{-j\frac{\pi}{2}} s_1 & -s_2^* \\ s_2 & s_1^* \end{bmatrix}^T \cdot \begin{bmatrix} h_1 \\ h_2 \end{bmatrix} + \begin{bmatrix} w_1 \\ w_2 \end{bmatrix} \qquad (18)$$

The new receive signal constellations are pictured in Fig. 8.

5.2. Delay scheme

In [9] Wittneben introduced a delay diversity technique. Every signal sent on transmit antenna T_{x1} is transmitted on antenna T_{x2} with a delay of one time step:

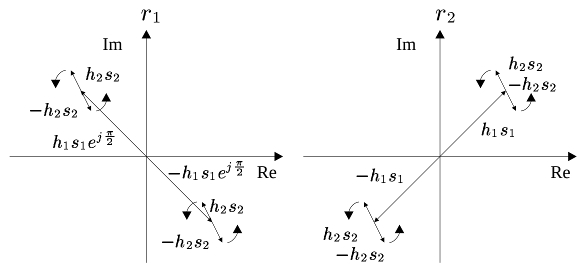

Fig. 8. Possible receive signals using modified Alamouti scheme with 2 PSK

$$\vec{r} = \begin{bmatrix} s_1 & s_2 & \cdots & s_n & 0 \\ 0 & s_1 & \cdots & s_{n-1} & s_n \end{bmatrix}^T \cdot \begin{bmatrix} h_1 \\ h_2 \end{bmatrix} + \begin{bmatrix} w_1 \\ \vdots \\ w_{n+1} \end{bmatrix}$$

Note that the receive vector $\vec{r}$ consists of $n + 1$ elements, i.d. the transmission of n symbols requires $n + 1$ time steps. The bandwidth efficiency of this scheme decreases with decreasing block length n since the ratio of transmitted symbols to used time steps $\frac{n}{n+1}$ decreases. The Rayleigh fading coefficients h_1 and h_2 are assumed to be constant during $n+1$ time steps. $w_1, \cdots, w_n$ denote the complex additive white Gaussian noise components.

For $n = 1$ only one symbol is transmitted and the delay diversity scheme passes over into a repetition scheme which achieves maximum diversity but has minimum efficiency $\frac{n}{n+1} = \frac{1}{2}$.

6. SIMULATION RESULTS

Using 2PSK, 4PSK and 8PSK we simulate the symbol error ratios (SER) for different diversity techniques over the signal-to-noise ratio per bit: (E_b/N_0). E_b means the energy per bit and N_0 denotes the noise power density.

In Fig. 9 the SER performance for 2PSK is displayed. As reference the SER of a linear detector using Alamouti scheme is given which achieves full diversity. For 1 receive and 2 transmit antennas full diversity means a SER's decreasing of two decades per 10dB signal-to-noise ratio. The phase detector with branch dependent metric and Alamouti scheme does not reach full diversity. But compared to the best known (branch independent) metric [6] a remarkable gain can be achieved. The modified Alamouti scheme improves the performance and achieves almost full diversity. The diversity gain which can be achieved with the delay scheme (block length $n = 3$) is almost identical but the SER is about 1dB worse. With growing block length the SER degrades which is exemplarily pictured for $n = 8$. Note that the bandwidth efficiency of these delay schemes is worse compared to the Alamouti scheme. A fair comparison with the Alamouti scheme can be performed with a 4PSK repetition code that also transmits one bit per time step. This scheme outperforms the 2PSK schemes with phase detection and reaches almost the SER performance of the linear receiver. However, the 4PSK with repetition code needs a more complex decoder than the 2PSK with modified Alamouti scheme.

Since the modified Alamouti scheme is not appropriate for 4 PSK we compare only the remaining schemes as shown in Figure 10. Again the Alamouti scheme with linear receiver achieves full diversity and gives a lower bound for the SER. Using a phase detector with branch dependent metric full diversity cannot be achieved

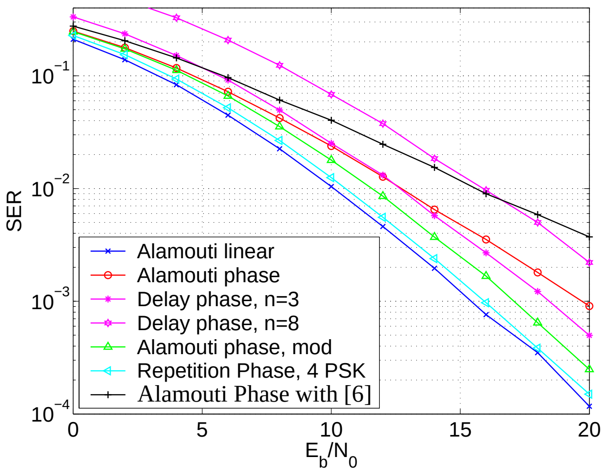

Fig. 9. Comparison of different transmit diversity schemes with 2 PSK and repetition code with 4 PSK

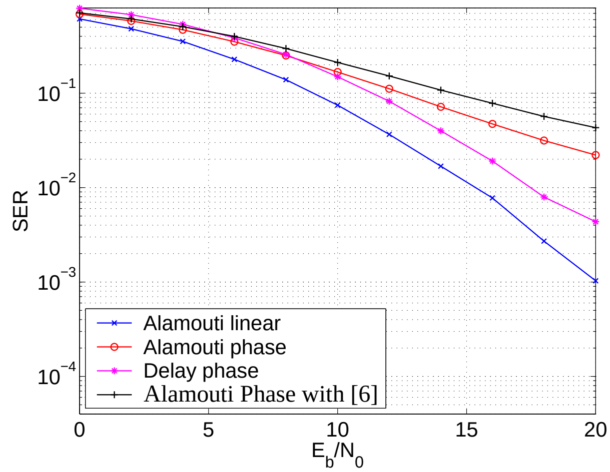

Fig. 11. Comparison of different transmit diversity schemes with 8 PSK

but the diversity is obviously larger than that of the branch dependent metric which achieves no diversity. The SER is crossed by the SER of the delay scheme ($n = 3$) which has higher SER at low E_b/N_0 but achieves almost full diversity. A repetition code with 16 PSK with the same bandwidth efficiency gains full diversity but the performance is worse than that of 4 PSK schemes. Obviously, the distances for the phase-space equalization decrease severely with the growing symbol alphabet.

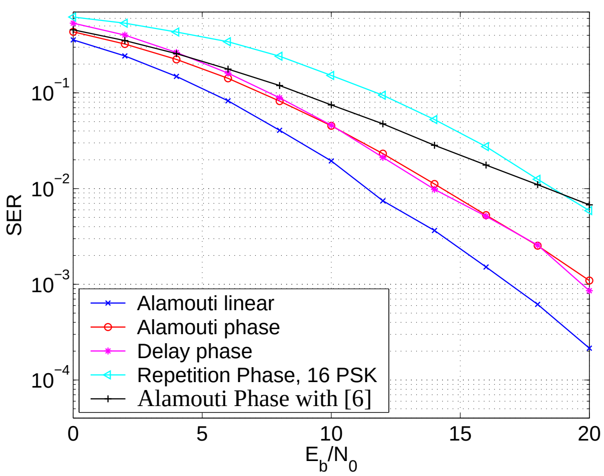

Fig. 10. Comparison of different transmit diversity schemes with 4 PSK and repetition code with 16 PSK

In Figure 11 the performance for 8 PSK is pictured. The phase detector with branch dependent metric achieves only poor diversity but shows still better performance than the branch independent metric. In contrast to that the delay scheme ($n = 3$) achieves nearly full diversity.

7. CONCLUSION

We proposed a very low-cost and energy aware transceiver structure for sensor network nodes. A new decoding and channel estimation method for this nonlinear phase detection receiver was presented. The performance of different diversity schemes for 1 receive and 2 transmit antennas were examined by means of simulations for 2PSK, 4PSK and 8PSK modulation. It could be shown that with this nonlinear structure diversity gains close to 2 can be achieved which allows to improve the network reliability.

8. REFERENCES

[1] Siavash M. Alamouti. A simple transmit diversity technique for wireless communications. *IEEE Journal on Selected Areas in Communications*, October 1998.

[2] F. Akyildiz, W. Su, Y. Sankarasubramaniam, and E. Cayirci. A survey on sensor networks. *IEEE Communications Magazine*, pages 102–114, August 2002.

[3] V. Raghunathan, C. Schurgers, S. Park, and M. Srivastava. Energy-aware wireless microsensor networks. *IEEE Signal Processing Magazine*, pages 40–50, March 2002.

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

[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, aug 1982.

[6] John P. Fonseka. Soft-decision phase detection with viterbi decoding for cpm signals. *IEEE Transactions on Communications*, 47(12):1802–1810, dec 1999.

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

[8] Simon Haykin. *Adaptive Filter Theory*. Prentice Hall, 3. edition, 1996.

[9] Armin Wittneben. Basestation modulation diversity for digital simulcast. *Vehicular Technology Conference, 1991. Gateway to the Future Technology in Motion., 41st IEEE*, 1991.