ON SIMPLIFIED SPACE-TIME RECEIVER STRUCTURES FOR GSM

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ABSTRACT

Space-time receiver structures for GSM have been proposed in [1]–[4]. This paper puts them onto a solid theoretical basis by means of estimation theory. Hereby, it becomes obvious where sub-optimalities arise from. In addition, an alternate structure is discussed that is advantageous for real-time implementation. The receiver structure exploits antenna diversity through a narrow-band combiner in the baseband signal processing. Adopting the minimum mean squared error criterion, the spatial weighting coefficients of the combiner are optimized jointly with the desired impulse response of the equivalent channel after combining. The solution of the optimization is equivalent to solving a generalized eigenvalue problem.
The performance of the resulting GSM receiver in a Generalized Hilly Terrain environment is evaluated by numerical simulation. Exploiting antenna diversity with two antenna elements at spacing $\lambda/2$ gives a 2 dB gain. Adding a third antenna element in a triangular array configuration yields an additional gain of 1 dB as compared to the case of just two receive elements.

1. INTRODUCTION

Algorithms which use a user-specific training sequence to identify the propagation channel are called "temporal reference algorithms". The training sequence is a sequence of known symbols, often called "pilot bits" or "midamble". Note that using training sequences not only allows estimation of the user's own channel impulse response, but also that of interfering users. Similarly, a user-specific spreading-sequence can be exploited, even if the sequence of transmitted symbols is not known at the receiver.
For single antenna systems the Viterbi equalizer (VE) receiver is the widespread receiver for mobile communication systems when ISI dominates the channel, e.g. for the GSM/DCS standard. For antenna array systems the receivers can exploit the spatial dimension. Based on the optimization criteria, space-time receivers can be grouped into two main classes [5, 6]: the space-time Viterbi equalizer (ST-VE) and the space-time minimum mean square error (ST-MMSE).
The ST-VE needs estimates of the channel impulse response $\boldsymbol{H}(t)$ and of the covariance matrix of the co-channel interference (CCI) [2]. Although these parameters can be estimated from a sufficiently long training sequence, problems can arise for non-stationary[1] CCI and fast Doppler. The MMSE receiver, on the other hand, allows the equalization of the channel and rejection of the CCI but this also requires the estimation of a high number of parameters (higher than the number of degrees of freedom in estimates for $\boldsymbol{H}(t)$).
For long training sequences the ST-VE achieves better performance than the ST-MMSE if the Viterbi algorithm is based on a consistent estimate of the channel and interference parameters. For a short training sequence the gain of using an ST-VE instead of an ST-MMSE becomes considerably smaller because the estimation of a high number of parameters becomes unstable. Therefore, it is not clear in such a case whether the higher numerical effort of the ST-VE is worth the gain. The 26 symbols of the GSM/DCS midamble are considered as being "short". A brief analysis is given in the following section.

2. SIGNAL MODEL

For an $N \times M$ communication system comprising N receiver antenna-elements and M transmit elements, the base-band equivalent channel is written as

$$\boldsymbol{x}(t) = \sum_k \boldsymbol{H}(t - kT)\boldsymbol{s}_k + \boldsymbol{u}(t) \ , \qquad (1)$$

where $\{\boldsymbol{s}_k\}$ is the M-dimensional transmitted vector symbol sequence and $\boldsymbol{H}$ is the $(N \times M)$ channel impulse response matrix of the desired signal. The individual symbols $\boldsymbol{s}_n$ stem from a finite alphabet. The multi-channel noise–plus–interference process

[1]e.g. cyclo-stationary

("impairment") is denoted by $\boldsymbol{u}(t)$ and it is assumed that

$$\boldsymbol{C}_{\boldsymbol{u}}(t_1, t_2) = \nu \boldsymbol{J} \delta(t_1 - t_2) \ , \qquad (2)$$

where $\boldsymbol{J}$ denotes the Hermitian positive-definite spatial coherence matrix of the noise–plus–interference. It is normalized such that ν denotes the one-sided single-antenna power spectral density, resulting in $\mathrm{tr}(\boldsymbol{J}) = N$. Thus, $\boldsymbol{u}(t)$ is assumed to be temporally white, but spatial correlation is allowed. It is briefly noted that temporal whiteness is not a necessary assumption and can be dropped.

3. OPTIMUM MIMO RECEIVER

The Viterbi algorithm, originally developed for decoding convolutional codes [7], has also led to a fundamental result in the optimum demodulation of channels exhibiting ISI [8].
The Gaussian prefilter in GMSK modulation introduces ISI that spreads over several bit intervals, thus degrading performance from MSK when coherent symbol-by-symbol detection is used. The optimal VE demodulator for GMSK requires $4(2^{L-1})$ states on AWGN channels [9, 10], where L is the ISI duration in bit intervals. The presence of severe multipath fading and narrow-band receive filtering to reduce Adjacent Channel Interference (ACI) further increases the number of states. Utilizing a linear representation of GMSK signals [11], it is possible to derive an VE GMSK demodulator that requires only 2^{L-1} states and achieves essentially the same BER performance as MSK [12].
The following derivation is based on Ref. [13] which describes the case of a single transmitter and receiver antenna ($M = N = 1$). The multiple input, multiple output (MIMO) ($M \geq 1$, $N \geq 1$) extension to the multi-antenna case follows the ideas in Ref. [14]. The generalization to the case of spatially colored interference plus noise is briefly described here.
The digital signal is characterised by a sequence of Dirac impulses

$$\boldsymbol{s}(t) = \sum_{k=0}^{n-1} \delta(t - kT)\boldsymbol{s}_k$$

where $\mathbf{s} = \{\boldsymbol{s}_k\}$ is a sequence (or equivalently a *vector*) from a finite alphabet, e.g. $\boldsymbol{s}_k \in \{1, j, -1, -j\}^M$, in what follows. The total transmission sequence is taken to be of arbitrary length n. Then the first part of the channel (characterised by a linear matrix-valued impulse response $\boldsymbol{H}(t)$) has output

$$\boldsymbol{v}(t; \mathbf{s}) = \sum_{k=0}^{n-1} \boldsymbol{H}(t - kT)\boldsymbol{s}_k. \qquad (3)$$

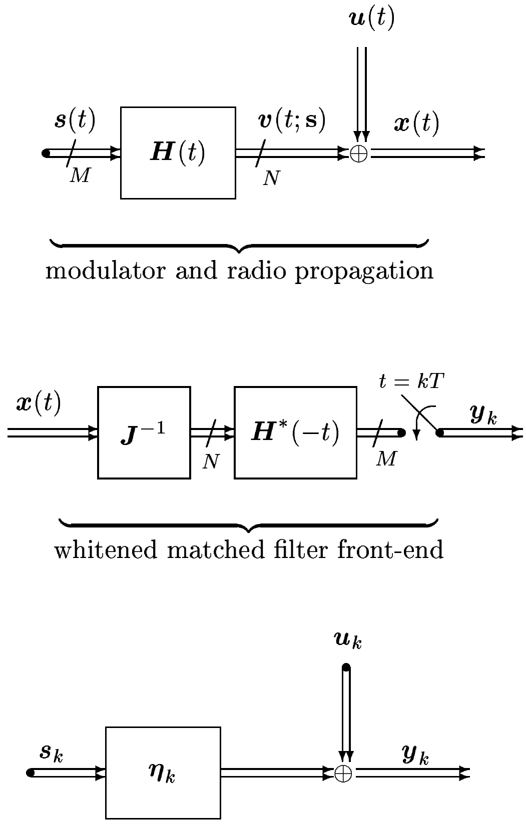

Figure 1: Top: Baseband model of transmitter, MIMO channel, and optimum receiver. The equivalent discrete-time channel is also shown at the bottom.

The additive noise–plus–interference process $\boldsymbol{u}(t)$ is taken to be a stationary complex Gaussian process with zero mean and temporally white with spatial covariance matrix $\boldsymbol{C}_{\boldsymbol{u}} = \nu \boldsymbol{J}$. Thus, the model from Eq.(2) is adopted.[2] If the noise–plus–interference is also assumed to be spatially white then $\boldsymbol{J}$ reduces to the identity matrix, $\boldsymbol{J} = \boldsymbol{I}$. Otherwise, $\boldsymbol{J}$ is assumed to be Hermitian positive-definite. The received vector signal at the individual antenna elements is denoted as

$$\boldsymbol{x}(t) = \boldsymbol{v}(t; \mathbf{s}) + \boldsymbol{u}(t) \ .$$

The decision rule which minimises the error probability based on the *entire received vector signal* $\boldsymbol{x}(t)$ can be

[2] This is an essential loss of generality in this derivation. If $\boldsymbol{u}$ is composed of at least some interfering signals with finite support, multi-user detection methods [15] can be applied to improve performance

shown to *minimise the squared Mahalanobis distance* [16] over all admissible input sequences $\mathbf{s}$. In the case of Gaussian $\boldsymbol{u}(t)$, this distance corresponds to the (negative) exponent inside the Gaussian probability density function, $d^2(\mathbf{s}) :=$

$$\int_{-\infty}^{\infty} (\boldsymbol{x}(t) - \boldsymbol{v}(t;\mathbf{s}))^* \boldsymbol{J}^{-1} (\boldsymbol{x}(t) - \boldsymbol{v}(t;\mathbf{s}))\, \mathrm{d}t \ .$$

By inserting (3) and neglecting terms that do not depend on the input symbols $\mathbf{s}$, it is seen that this is equivalent to maximising the quantity

$$\Lambda_n(\mathbf{s}) = 2\,\mathrm{Re} \sum_{k=0}^{n-1} \boldsymbol{s}_k^* \boldsymbol{y}_k - \sum_{k=0}^{n-1}\sum_{\ell=0}^{n-1} \boldsymbol{s}_k^* \boldsymbol{\eta}_{k-\ell} \boldsymbol{s}_\ell$$

where we have defined the *sampled output of the whitened matched filter bank*

$$\boldsymbol{y}_k = \left. \int_{-\infty}^{\infty} \boldsymbol{H}^*(t-\tau) \boldsymbol{J}^{-1} \boldsymbol{x}(t)\, \mathrm{d}t \right|_{\tau = kT} \tag{4}$$

and the *intersymbol interference matrices*

$$\boldsymbol{\eta}_{k-\ell} = \int_{-\infty}^{\infty} \boldsymbol{H}^*(t-kT) \boldsymbol{J}^{-1} \boldsymbol{H}(t-\ell T)\, \mathrm{d}t \tag{5}$$

The variables $\boldsymbol{y}_k$ are the observables on which all decisions will be based. Note that they can be viewed as sampling the output of the whitened matched filter. Thus, the received waveform vector-signal $\boldsymbol{x}(t)$ is first spatially whitened and finally convolved with the matched filter at the symbol-rate T.
Note that the equivalent discrete-time noise–plus–interference sequence $\boldsymbol{u}_k$ in Fig. 1 is a filtered version of $\boldsymbol{u}(t)$: it is filtered by the whitened matched filter bank at the receiver front-end

$$\boldsymbol{u}_k = \left. \int_{-\infty}^{\infty} \boldsymbol{H}^*(t-\tau) \boldsymbol{J}^{-1} \boldsymbol{u}(t)\, \mathrm{d}t \right|_{\tau = kT} \tag{6}$$

Thus, $\boldsymbol{u}_k$ is temporally correlated and the covariance is determined by the ISI coefficients,

$$\mathsf{E}\left[\boldsymbol{u}_k \boldsymbol{u}_\ell^*\right] = \nu \boldsymbol{\eta}_{k-\ell} \ . \tag{7}$$

Although the ISI coefficients $\boldsymbol{\eta}_i$ are potentially nonzero for all i, in practice, for sufficiently large i, we will have $\boldsymbol{\eta}_i \approx \mathbf{0}$. We shall accept this approximation to limit the dimensionality of the problem. Thus, we take

$$\boldsymbol{\eta}_i = \mathbf{0} \qquad \text{for } i \geq L \qquad \text{where } L \ll n$$

Also, by virtue of the conjugate-symmetry of the coefficients $\boldsymbol{\eta}_i$, the symmetric quadratic form inside the expression for Λ_n can be written as the sum of the diagonal terms plus twice the real part of the upper triangular quadratic form, i.e.

$$\sum_{k=0}^{n-1}\sum_{\ell=0}^{n-1} \boldsymbol{s}_k^* \boldsymbol{\eta}_{k-\ell} \boldsymbol{s}_\ell = \sum_{k=0}^{n-1} \boldsymbol{s}_k^* \boldsymbol{\eta}_0 \boldsymbol{s}_k + 2\mathrm{Re}\left\{ \sum_{k=0}^{n-1} \boldsymbol{s}_k^* \sum_{i=1}^{k+n} \boldsymbol{\eta}_i \boldsymbol{s}_{k-i} \right\}. \tag{8}$$

This can be substituted into the expression for Λ_n and truncated for $i \geq L$. The key observation is that Λ_k can be evaluated recursively in time (i.e. the index k steps through sampling points kT)

$$\Lambda_{k+1} = \Lambda_k + \lambda_{k+1} \tag{9}$$

such that

$$\Lambda_n(\mathbf{s}) = \sum_{k=0}^{n-1} \lambda_k \qquad \text{with} \quad \lambda_k = 2\,\mathrm{Re}\left\{ \boldsymbol{s}_k^* \left(\boldsymbol{y}_k + \boldsymbol{\eta}_0 \boldsymbol{s}_k - \sum_{i=0}^{L-1} \boldsymbol{\eta}_i \boldsymbol{s}_{k-i} \right) \right\}.$$

These expressions define the path-metric Λ_n of the trellis diagram as the sum of the individual branch-metrics λ_k which establishes the decoding criterion on the MIMO channel. Note that the branch metric of the Viterbi equalizer depends on the vector-valued symbol-rate sequence $\boldsymbol{y}_k$. The number of internal states of the Viterbi algorithm becomes b^{ML} where b is the size of the transmitter alphabet.

4. ALTERNATE STRUCTURE FOR SINGLE USER CASE

It is concluded from the previous section that the single-user ST-VE receiver (i.e. $M = 1$) consists of a whitened matched-filter bank for the individual antennas as a front-end. The front-end output signals are added and sampled at the symbol-rate. Finally, a scalar Viterbi equalizer using the branch metric (9) is responsible for eliminating the remaining ISI in time [1]–[4].
Aside from the great computational complexity of the optimum MIMO receiver described in Section 3, the VE is quite sensitive to co-channel interferers or non-Gaussian noise. At the same time, the length of the estimated channel has a major impact in VE in terms of complexity and decoder delay. In summary, any pre-processing aiming at the reduction of the channel

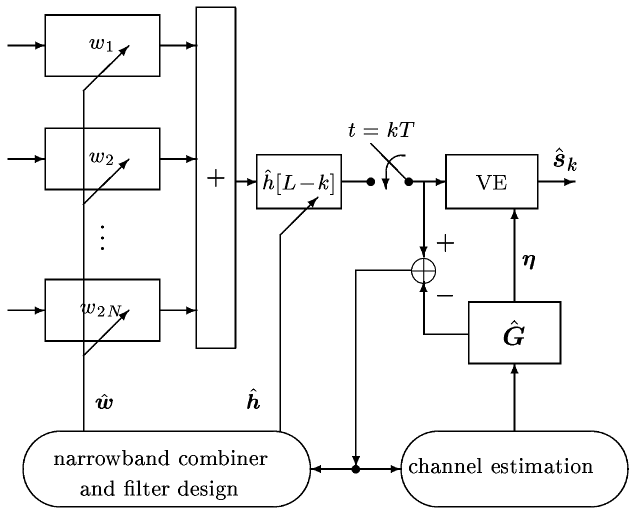

Figure 2: MDIR receiver with narrowband combining for N antennas. The derotation at the receiver front-end is not shown. The MDIR architecture operates on $2N$ equivalent real-valued input channels.

length and removal of co-channel interferers is welcome. The MDIR (Matched Desired Impulse Response) [4] takes advantage of both spatial combining and VE: the array combiner plus symbol-time sampler cancels co-channel interference and yields white Gaussian noise, while the VE block optimally deals with ISI and detects symbols. Figure 2 depicts the MDIR structure with narrowband combiner. Although suboptimal, the MDIR with narrowband combiner results in a good compromise between reduction in complexity and performance.

Concerning the the spatial pre-whitening operation $\boldsymbol{J}^{-1}$, it can be dealt either at the VE block, by using its covariance matrix in the computation of the metrics in the Viterbi trellis, or it can be moved to the front-end of the receiver [17]. Finally, note that the length of the desired impulse response used by the Viterbi Equalizer determines the maximum delay (ISI order) allowed to pass the beamformer. In other words, if the length equals to L symbol intervals, then only arrivals later than L symbol intervals will be considered as co-channel interference at the beamformer stage. The early arrivals will be managed as desired signal at this stage. Simulation results show that if the matched filter length is 4, then the impulse response is usually between 4 and 6 in order to bound the complexity and delay of the Viterbi Equalizer.

4.1. Derotation at receiver front-end

The investigated receiver for GMSK waveforms samples the I– and Q–phases jointly at all N antenna elements at the symbol rate. Next, the (I,Q) components are derotated which exploits the inherent redundancy in the GMSK waveform and removes the differential encoding of the GSM data bits. This results in real-valued channel impulse responses for the created sub-channels with binary inputs and all calculations from now on are performed with real-valued quantities. The number of equivalent real-valued diversity channels becomes $2N$.

4.2. Joint channel– and weight estimation

The joint design of the spatial combiner $\boldsymbol{w}$ and the impulse response $\boldsymbol{h}$ for the Viterbi equalizer is based on the Minimum Mean Squared Error (MMSE) criterion

$$\varepsilon^2 = \varepsilon^2(\boldsymbol{w}, \boldsymbol{h}) = \|\boldsymbol{X}\boldsymbol{w} - \boldsymbol{D}\boldsymbol{h}\|^2, \qquad (10)$$

where $\|\cdot\|^2$ is the Euclidian norm. The real-valued observation matrix $\boldsymbol{X}$ of the midamble portion of the received data has dimension $(n - L + 1) \times 2N$

$$\boldsymbol{X} = \begin{pmatrix} x_1(L-1) & x_2(L-1) & \cdots & x_{2N}(L-1) \\ x_1(L) & x_2(L) & \cdots & x_{2N}(L) \\ \vdots & \vdots & \ddots & \vdots \\ x_1(n-1) & x_2(n-1) & \cdots & x_{2N}(n-1) \end{pmatrix} . \qquad (11)$$

Thus, $\boldsymbol{X}$ contains the received midamble after dropping the first $L-1$ samples. The first $L-1$ samples are dropped because they depend on the transmitted user bits through the channel memory of length $L-1$. For GSM the midamble length n is 26.

Let the Toeplitz matrix $\boldsymbol{D}$ containing the bipolar $\{-1, 1\}$ symbols of the GSM training sequence code be defined as

$$\boldsymbol{D} = \begin{pmatrix} d_{L-1} & d_{L-2} & d_{L-3} & \cdots & d_1 & d_0 \\ d_L & d_{L-1} & d_{L-2} & \cdots & d_2 & d_1 \\ d_{L+1} & d_L & d_{L-1} & \cdots & d_3 & d_2 \\ \vdots & \vdots & \vdots & \ddots & \vdots & \vdots \\ d_{n-1} & d_{n-2} & d_{n-3} & \cdots & d_{n-L+1} & d_{n-L} \end{pmatrix} . \qquad (12)$$

The dimension of $\boldsymbol{D}$ is $(n-L+1) \times L$. Then we obtain the least-squares estimates of the $2N$ parallel diversity channel impulse responses by

$$\hat{\boldsymbol{G}} = (\boldsymbol{D}^T\boldsymbol{D})^{-1}\boldsymbol{D}^T\boldsymbol{X} \ . \qquad (13)$$

The space–time channel impulse response matrix $\hat{\boldsymbol{G}}$ has dimension $L \times 2N$.

Note that the optimization of $\varepsilon^2(\boldsymbol{w}, \boldsymbol{h})$ over all $\boldsymbol{w}, \boldsymbol{h}$ is aimed at keeping the multipath content of the signal so it can be used by the Viterbi Equalizer. A constraint has to be imposed in order to avoid the trivial solution. It is chosen to constrain the desired energy at the output of the spatial combiner [4]

$$E = \|\boldsymbol{D}\hat{\boldsymbol{G}}\boldsymbol{w}\|^2 = \boldsymbol{w}^T\hat{\boldsymbol{G}}^T\boldsymbol{D}^T\boldsymbol{D}\hat{\boldsymbol{G}}\boldsymbol{w} \ . \tag{14}$$

Introducing a Lagrangian multiplier λ, we arrive at the following cost function of the unconstrained optimization problem

$$f(\boldsymbol{w},\boldsymbol{h},\lambda) = \varepsilon^2(\boldsymbol{w},\boldsymbol{h}) - \lambda(\boldsymbol{w}^T\hat{\boldsymbol{G}}^T\boldsymbol{D}^T\boldsymbol{D}\hat{\boldsymbol{G}}\boldsymbol{w} - E) \ . \tag{15}$$

A closed-form solution for $\boldsymbol{h}$ is available

$$\hat{\boldsymbol{h}} = \hat{\boldsymbol{G}}\boldsymbol{w} \ . \tag{16}$$

Finally, we obtain a generalized eigenvalue problem for the vector $\boldsymbol{w}$

$$[\boldsymbol{X}^T\boldsymbol{P}_{\boldsymbol{D}}^{\perp}\boldsymbol{X}]\boldsymbol{w} = \lambda[\hat{\boldsymbol{G}}^T\boldsymbol{D}^T\boldsymbol{D}\hat{\boldsymbol{G}}]\boldsymbol{w} \tag{17}$$

where the projection matrix $\boldsymbol{P}_{\boldsymbol{D}}^{\perp}$ is defined as

$$\boldsymbol{P}_{\boldsymbol{D}}^{\perp} = \boldsymbol{I} - \boldsymbol{D}(\boldsymbol{D}^T\boldsymbol{D})^{-1}\boldsymbol{D}^T \ .$$

If the training sequence and the noise-plus-interference are uncorrelated processes then we can adopt (for sufficiently large $n - L$) the following approximation

$$\boldsymbol{D}^T\boldsymbol{X} \approx \boldsymbol{D}^T\boldsymbol{D}\hat{\boldsymbol{G}}\boldsymbol{w} \ . \tag{18}$$

This approximation allows to rewrite (16b) as

$$\boldsymbol{h} = \hat{\boldsymbol{G}}\boldsymbol{w} \tag{19}$$

$$\hat{\boldsymbol{R}}\boldsymbol{w} = \lambda\hat{\boldsymbol{G}}^T\boldsymbol{D}^T\boldsymbol{D}\hat{\boldsymbol{G}}\boldsymbol{w} \tag{20}$$

where the noise-plus-interference covariance matrix is defined as

$$\hat{\boldsymbol{R}} = (\boldsymbol{X} - \boldsymbol{D}\hat{\boldsymbol{G}})^T(\boldsymbol{X} - \boldsymbol{D}\hat{\boldsymbol{G}}) \approx \boldsymbol{X}^T\boldsymbol{X} - \hat{\boldsymbol{G}}^T\boldsymbol{D}^T\boldsymbol{D}\hat{\boldsymbol{G}} \ . \tag{21}$$

Up to a constant factor, the matrix $\boldsymbol{X}^T\boldsymbol{X}$ is an estimate of the cross spectral density matrix of the array observations

$$\hat{\boldsymbol{C}}_{\underline{\mathrm{x}}} := \frac{\boldsymbol{X}^T\boldsymbol{X}}{n - L + 1} \ . \tag{22}$$

The effective impulse response $\boldsymbol{h}$ (which the VE block tries to equalize) belongs to the space-time channel $\hat{\boldsymbol{G}}$ filtered by the array combiner with weight $\boldsymbol{w}$. The SINR at the output of the sampler is given by

$$\mathrm{SINR} = \frac{\|\boldsymbol{D}\hat{\boldsymbol{G}}\boldsymbol{w}\|^2}{\varepsilon^2_{\mathrm{opt}}} = \frac{\boldsymbol{w}^T\hat{\boldsymbol{G}}^T\boldsymbol{D}^T\boldsymbol{D}\hat{\boldsymbol{G}}\boldsymbol{w}}{\boldsymbol{w}^T\hat{\boldsymbol{R}}\boldsymbol{w}} = \frac{1}{\lambda} \tag{23}$$

Therefore, the coefficients of the linear combiner $\boldsymbol{w}$ are given by the generalized eigenvector of Eq.(20) associated to the *minimum* generalized eigenvalue. It is concluded from Eq.(21) that the eigenvalues are positive. The impulse response of the channel that is to be used in the VE block is the matched response of the linear combiner plus the sampler to the physical baseband channel.

Care has to be taken to guarantee that the matrix $\hat{\boldsymbol{R}}$ is full rank. Otherwise the computation of the generalized EVD is not a well conditioned problem. When the number of rows in matrix $\boldsymbol{X}$ is smaller than the number of columns $2N$ then the product $\boldsymbol{X}^T\boldsymbol{X}$ is rank deficient and so is $\hat{\boldsymbol{G}}^T\boldsymbol{D}^T\boldsymbol{D}\hat{\boldsymbol{G}}$. It has been observed that diagonal loading on $\hat{\boldsymbol{R}}$ leads to improved performance:

$$\hat{\boldsymbol{R}} \leftarrow \hat{\boldsymbol{R}} + \sigma^2\boldsymbol{I}$$

with a value of σ^2 just below the noise level. The degree of load is not critical but leads to a biased estimate for the optimum weighting coefficients and a distorted directional response of the antenna array. A large value for σ^2 implies that the most significant undesired signal is spatially white.

5. PERFORMANCE EVALUATION

The performance of the described receiver was evaluated numerically by simulations. These were performed with the directional channel model recommended by the European research initiative COST–259. This model ist geometry-based, stochastic, and assumes single scattering. The Generalized Hilly Terrain (GHT) scenario was selected for evaluation and the parameters are given in Table 1. For details of the implementation, we refer to [18]

The receiver was implemented for a channel length $L = 5$ resulting in a VE with 16 states.

The Bit Error Ratio (BER) over Signal– to Noise Ratio (SNR) is shown in Figure 3. The three curves correspond to the case of 1, 2, and 3 antenna elements at the receiver. Increasing the number of receive antennas from 1 to 2 with spacing $\lambda/2$ gives 2 dB gain in this environment. Thus, we observe a 1 dB combining loss compared to the optimum of 3 dB due to a loss in coherency of the wavefronts. Adding a third antenna element in a triangular array configuration yields an additional gain of 1 dB as compared to the case of 2 receive elements.

6. CONCLUSION

The MDIR receiver structure is advantageous for real-time implementation. This paper investigates the theoretical basis of the MDIR receiver by means of estimation theory and the performance is evaluated by means of simulation. It turns out that two-branch antenna diversity with spacing $\lambda/2$ yields roughly 2 dB gain in SNR for the MDIR receiver whereas three-branch diversity yields 3 dB as compared to the single antenna case.

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Channel model parameter	Value
Center frequency	1845 MHz
Range BS—MS	4 km
Mobile speed	3 km/h
Base station antenna height	50 m
No. of Tx antennas	1
No. of Rx antennas	1, 2, 3
Rx antenna element spacing	$\lambda/2$
Sampling rate	3.7 μs
No. of scaterers per cluster	50

Table 1: COST–259 Parameters for GHT environment

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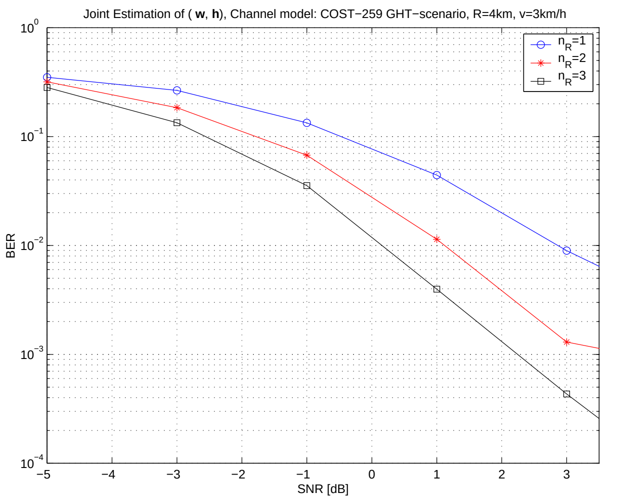

Figure 3: Bit error ratio versus SNR for the MDIR receiver operating with $N = 1 \ldots 3$ antennas at $\lambda/2$ spacing. Channel model according to COST–259 "Generalized Hilly Terrain", see Table 1 for details.